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# PROCEEDINGS

*of*

# The Institute of Radio Engineers

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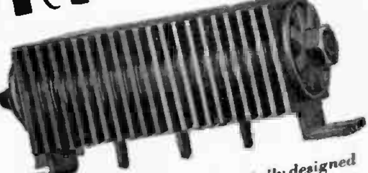
# SMOOTH Power for Dynamic Speakers from the AC Line with the "ELKON" DRY Rectifier



*Dynamic Speaker Series, especially designed to meet the needs of Manufacturers*

THE Elkon Engineers have worked exceedingly closely with many of the manufacturers of dynamic speakers. The result of this work is an intimate knowledge of the problems of supplying smooth power to moving coils, and a complete series of rectifiers supplying just the voltage and current desired.

Already the Elkon Dry Rectifier has been specified by a majority of the Dynamic Speaker manufacturers. We would like to work with you as we have with them. Simply tell us your specifications, and we will submit a sample rectifier — May be supplied with or without power transformer.

Self healing, the Elkon Rectifiers have an exceptionally long life, are noiseless and require no attention or adjustment.

Let Rectifier Headquarters solve your problems. Send the coupon which will bring you the Engineering Bulletins on the Elkon Rectifiers and High Capacity Dry Condensers.

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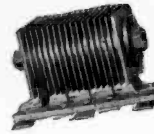

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## Elkon Rectifiers are Self Healing

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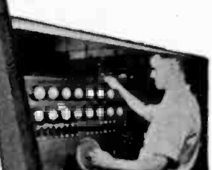
*Row upon row of Elkon Rectifiers being seasoned under the careful eyes of skilled workers*



*Every junction must have just the right pressure — 16 meters to watch in this operation*

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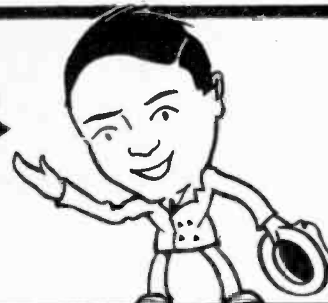
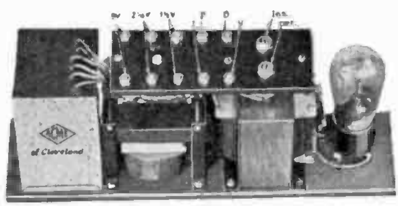


# POWER PACK



MANUFACTURERS MODEL POWER PACK ABC
FOR AC TUBES USING UX280 OR CX380 TUBES
WIDTH 3 INCHES. LENGTH 12 INCHES,
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CAPACITY 40 MILLS. AT 160-180 VOLTS
OPERATES ON 110 115 VOLTS, 50-60 CYCLES

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*Established in 1917*

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REPRESENTING

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MADE BY
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Radio - Is - *BETTER* - With - Dry - Battery - Power

# made to run the *full* race!

*ANY* horse can make a good start . . . . . . But it takes real stamina to *finish!* ¶So it is with batteries. *Staying* power is the quality to look for—unfailing power over a long period of service. Millions prefer Burgess *Chrome* Batteries for just this reason. They hold up . . . . . . . They *last*. ¶Next time, buy black and white striped Burgess *Chrome* Batteries. You are certain to get longer and better service for your money.

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*Ask Any Radio Engineer*

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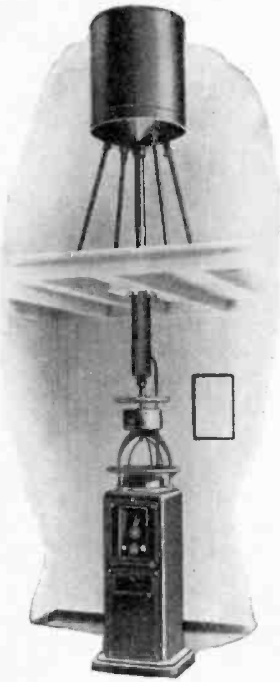
# BURGESS BATTERIES

# The new Kolster Radio Compass provides

## *Greater Safety— Visual Bearings— Simple, Positive Operation*



From its aluminum pedestal to the new enclosed loop, the new Kolster Radio Compass [Type AM-4490] embodies every improvement radio science has to offer. The insulated cylindrical housing of the loop affords complete protection against wind, ice, snow, and spray. A tiny lamp flashes the signals of nearby stations. While for long-distance bearings, a Kolster eight-tube receiver with a new circuit especially designed for radio compass work provides the maximum of selectivity and sensitiveness.

No knowledge of radio is necessary to operate the Kolster Radio Compass. It is built for the navigator. The positive unidirectional indicator gives instantaneous readings without guesswork or computation. A new and improved Kolster automatic compensator corrects all natural errors, and makes direct readings absolutely accurate.

*Kolster Radio is also setting new standards of performance in brose.*

## KOLSTER RADIO CORPORATION

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# RADIOTRONS for modern A. C. OPERATION

## POWER AMPLIFIER RADIOTRON UX-171

*Filament—5 Volts—.5 Amperes*

| | | | | |
|---|---|---|---|---|
| Plate Voltage | 90 | 135 | 180 | Volts |
| Negative Grid Bias | 16½ | 27 | 40½ | Volts |
| Plate Current | 10 | 16 | 20 | Milliamperes |
| Plate Resistance (A.C.) | 2500 | 2200 | 2000 | Ohms |
| Mutual Conductance | 1200 | 1360 | 1500 | Micromhos |
| Voltage Amplification Factor | 3.0 | 3.0 | 3.0 | |
| Max. Undistorted Output | 130 | 330 | 700 | Milliwatts |

## R. F. & A. F. AMPLIFIER RADIOTRON UX-226

*Filament (A. C.) 1.5 Volts—1.05 Amperes*

| | | | | | |
|---|---|---|---|---|---|
| Plate Voltage | 90 | 135 | 135 | 180 | Volts |
| Negative Grid Bias | 6 | 12 | 9 | 13½ | Volts |
| Plate Current | 3.7 | 3 | 6 | 7.5 | Milliamperes |
| Plate Resistance (A.C.) | 9400 | 10,000 | 7400 | 7000 | Ohms |
| Mutual Conductance | 875 | 820 | 1100 | 1170 | Micromhos |
| Voltage Amplification Factor | 8.2 | 8.2 | 8.2 | 8.2 | |
| Max. Undistorted Output | 20 | 60 | 70 | 120 | Milliwatts |

## DETECTOR RADIOTRON UY-227

*Heater (A. C.) 2.5 Volts—1.75 Amperes*

| | | | |
|---|---|---|---|
| Plate Voltage | 45 | 90 | Volts |
| Grid Leak | 2-9 | ¼-1 | Megohms |
| Plate Current | 2 | 7 | Milliamperes |
| Plate Resistance (A.C.) | 10,000 | 8000 | Ohms |
| Mutual Conductance | 800 | 1000 | Micromhos |
| Voltage Amplification Factor | 8 | 8 | |

## FULL WAVE RECTIFIER RADIOTRON UX-280

| | | |
|---|---|---|
| A.C. Filament Voltage | 5.0 | Volts |
| A.C. Filament Current | 2.0 | Amperes |
| A.C. Plate Voltage (Max. per plate) | 300 | Volts |
| D.C. Output Current (Maximum) | 125 | Milliamperes |
| Effective D.C. Output Voltage of typical Rectifier Circuit at full output current as applied to Filter | 260 | Volts |

## HALF WAVE RECTIFIER RADIOTRON UX-281

| | | |
|---|---|---|
| A.C. Filament Voltage | 7.5 | Volts |
| A.C. Filament Current | 1.25 | Amperes |
| A.C. Plate Voltage (Max. per plate) | 750 | Volts |
| D.C. Output Current (Maximum) | 110 | Milliamperes |
| Effective D.C. Output Voltage of typical Rectifier Circuit at full output current as applied to Filter | 620 | Volts |

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New York Chicago San Francisco

# RCA Radiotron

MADE BY THE MAKERS OF THE RADIOLA

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**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.

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**Abbreviations**—Write a.c. and d.c., kc, $\mu$f, $\mu\mu$f, emf, mh, $\mu$h, 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.

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PROCEEDINGS OF

# The Institute of Radio Engineers

Volume 16 | June, 1928 | Number 6

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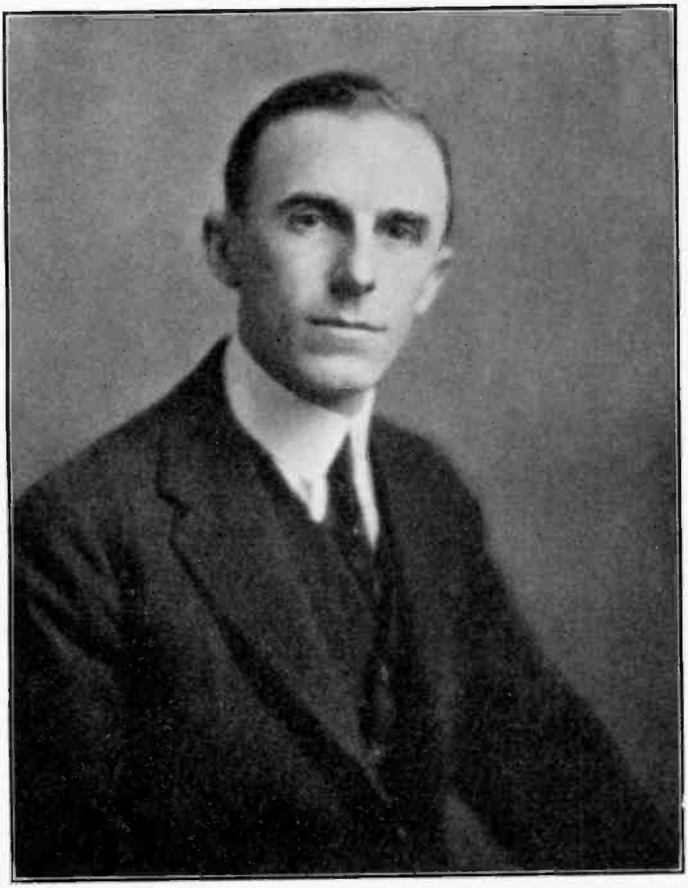


Ray H. Manson

Member of Board of Direction of the Institute, 1928

## Ray H. Manson

MEMBER OF THE BOARD OF DIRECTION OF THE INSTITUTE, 1928

Ray H. Manson was born in Bath, Maine, on August 25, 1877. He graduated from the University of Maine in 1898, and received the E.E. degree from the same institution in 1921. The year following graduation he was an assistant in the Electrical Engineering Department of the University of Maine.

In 1899–1900 Mr. Manson was employed in the Telephone Manufacturing Department and the Electrical Laboratory of the Western Electric Company in Chicago. From 1901 to 1904 he was in the Engineering and Sales Departments of the Kellogg Switchboard and Supply Company at Chicago.

During the period of 1904 to 1916 Mr. Manson was connected with the Dean Electric Company and its successor, the Garford Manufacturing Company, at Elyria, Ohio, the latter four years being Chief Engineer. Since 1916 he has been Chief Engineer of the Stromberg-Carlson Telephone Manufacturing Company at Rochester, New York.

During his association with the manufacturing business over fifty U. S. patents on telephone, phonograph, and radio subjects have been granted him.

Mr. Manson has been active in radio standardization work in the Radio Division of the National Electrical Manufacturers' Association. He is Chairman of the Technical Committee of that organization. For several years he has been Chairman of the Electro-Acoustic Subcommittee on the I. R. E. Committee on Standardization.

Mr. Manson was appointed a Manager of the Institute by the Board of Direction in January of 1927, and was elected by the Membership to the Board for a three-year period on January 3, 1928. He is a Member of the Institute, a member of the American Institute of Electrical Engineers, and a Member of the Society of Automotive Engineers.

## CONTRIBUTORS TO THIS ISSUE

**Dahl, O.:** (See PROCEEDINGS for March, 1928).

**Espenschied, Lloyd:** Born April 27, 1889 at St. Louis. E. E. course at Pratt Institute, 1909; amateur experimenter and radio operator on ships, 1905–09; assistant engineer Telefunken Wireless Telegraph Company of America, 1909–10; engineer, Department of Development and Research, American Telephone and Telegraph Company since 1910. In charge of transmission of high frequency in wire carrier and radio systems. Responsible for a number of inventions relative to communication systems, both wire and radio; also in field of railway signal and speed control systems and phonograph systems. He has contributed frequently to the PROCEEDINGS; is a charter member and Fellow of the Institute, and served on the Board of Direction of the Institute from 1913 to 1927.

**Nelson, J. R.:** Born on October 27, 1899 at Murray, Utah. Power Inspector, Western Electric Company, 1922–23; B.S. in E.E., University Southern California, 1925; Engineering Record Office, Bureau of Power and Light, Los Angeles, Calif., 1925; Radio Test, Radio Development Laboratory and Tube Research Laboratory, 1925–27. M.S. in E.E. degree, Union College, 1927. July, 1927 to date, Engineering Department, E. T. Cunningham, Incorporated. Associate member of the Institute.

**Owens, Raymond B.:** A.B. degree Randolph-Macon College, 1917. Post-graduate work at George Washington University. In naval radio service 1918–19; Radio Test Shop, Washington Navy Yard, 1919–21; Naval Radio Laboratory, Bureau of Standards, 1921–23; Research work in connection with precision apparatus for Naval frequency standardization, Naval Research Laboratory, 1923 to date. Member of the Institute.

**Pickard, Greenleaf W.:** (See PROCEEDINGS for May, 1928).

**Prince, D. C.:** Received B.S. in E.E. degree at University of Illinois, 1912; M.S. in E.E., University of Illinois 1913; General Electric Company, Test and Railway Motor Department, 1913–14; State Public Utilities Commission of Illinois, Assistant to Electrical Engineer, 1914–17; Ordnance Department, A.E.F., 1917–19; Research Laboratory, General Electric Company, 1923 to date. Fellow, American Institute of Electrical Engineers and American Physical Society.

**Terman, Frederick Emmons:** (See PROCEEDINGS for April, 1928).

**Turner: H. M.:** Born at Hillsboro, Illinois, July 20, 1882. Received B. S. degree, University of Illinois, 1910; M. S. degree, 1915; assistant instructor in electrical engineering, 1910–12. Instructor, University of Minnesota, 1912–18. Assistant professor, Yale University, 1918–26; Associate professor, 1926– . Frequent contributor to the leading engineering journals. Member I.R.E., A.I.E.E., and International Union of Scientific Radiotelegraphy.

**Tuve, M. A.:** Born on June 27, 1901 at Canton, South Dakota. Amateur radio operator, 1915–21; Received B.S. in E.E. degree and M.A. degree from University of Minnesota; Ph. D. degree from Johns Hopkins, 1926; former Instructor in Physics, Princeton University and Johns Hopkins University. At present, associate physicist, Department of Terrestrial Magnetism, Carnegie Institution of Washington.

**Van Dyke, K. S.:** Born at Brooklyn, N. Y., December 8, 1892; Received B.S. degree from Wesleyan University, 1916; M.S. degree, 1917. Ph.D. degree, University of Chicago, 1921. Engineering Department, American Telephone and Telegraph Company, 1917–19; Assistant in physics department, University of Chicago, 1919–21; Assistant Professor of Physics, Wesleyan University, 1921–25; 1925 to date Associate Professor of Physics, Wesleyan University. Member of the Institute. Secretary (1927) and Vice-Chairman (1928) Connecticut Valley Section of the Institute.

**Williams, N. H.:** Ph.D. degree, 1912. Research Laboratory, General Electric Company, 1923–24, engaged in development of screen-grid tube and measurement of the charge of the electron by the "shot-effect." On staff of Physics Department, University of Michigan, 1909 to date. At present Professor of Physics.

**Worrall, Robert H.:** Educated at Lowell Textile School and George Washington Unversity. With Manchester Traction and Power company, Manchester, New Hampshire, 1912–16 in charge of electrical machine shop; assistant in charge of Transmitter Test Laboratory, Marconi Wireless Telegraph Company, 1916–17; in charge of Receiver Test Laboratory, Washington Navy Yard, 1917–23; research and design pertaining to frequency standardization, Naval Research Laboratory, 1923 to date. Member of the A. I. E. E. and Associate member of the Institute.

**Yagi, Hidetsugu:** Born, January 28, 1886 at Osaka City, Japan. Graduated and received degrees from Tokio Imperial University, 1909, and Department of Education, Japanese Government, 1919. Lecturer at Sendai Higher Technical School, 1909-10; Professor in electrical engineering, Sendai Higher Technical College, 1910-1912; travelling Fellow, Department of Education of Japanese Government, studying in Berlin, Dresden, London, and at Harvard, 1913-16; Professor of electrical engineering, Tohoku Imperial University, Sendai, Japan, 1916 to date. Fellow of the Institute.

# INSTITUTE ACTIVITIES

## May Meeting of the Board of Direction

A meeting of the Board of Direction was held in the office of the Institute on May 2nd. The following were present: L. E. Whittemore, Vice-President; Melville Eastham, Treasurer; Donald McNicol, Junior Past-President; W. G. Cady, J. H. Dellinger, R. A. Heising, J. V. L. Hogan, R. H. Marriott, and J. M. Clayton, Secretary.

The following were transferred or elected to higher grades of membership in the Institute: Transferred to the grade of Fellow: C. M. Jansky, Jr.; Transferred to the grade of Member: W. E. Brindley, J. E. Fetzer, C. W. Horn, C. M. Howard, and S. R. Montcalm. Elected to the grade of Member: R. W. Ackerman, A. J. Chesterton, C. R. Hanna, A. H. Morton, O. A. Pearson, and A. M. Trogner.

One hundred and ten Associate members and ten Junior members were elected.

A petition from members residing within the vicinity of New Orleans, Louisiana was submitted to the Board for approval of the formation of a New Orleans Section of the Institute. The petition was approved. Prominent in the organization work of the New Orleans Section were Pendleton E. Lehde, T. G. Deiler, L. J. N. duTreil, and A. Schiele.

Lewis M. Hull was appointed the Institute's representative on the Sectional Committee on Radio of the American Engineering Standards Committee, to succeed Professor L. A. Hazeltine.

## 1928 Membership Cards

As announced on several occasions in past issues of the Proceedings, membership cards are available for *all* members this year. Cards can be secured only upon specific request to the office of the Secretary, and are not mailed automatically upon payment of member dues.

## New Style of I. R. E. Emblem

A new membership emblem, in the form of a screw-back lapel button, approximately one-half the size of the present emblem, will be available within approximately thirty days. This emblem will be finished in the various colors signifying the different grades of membership. It is of 14 k. gold and can be purchased from the Secretary for $2.75 postpaid.

### 1928 Year Book

The 1928 Year Book of the Institute, containing general information concerning the Institute, the Constitution and By-Laws, a catalog of the membership listed both alphabetically and geographically, and other information of value, is at the printer's and should be mailed as a supplement to either the June or July issue of the PROCEEDINGS.

### References to Radio Literature

Through the cooperation of the Bureau of Standards, with the July issue the PROCEEDINGS will contain the monthly list of references to current radio literature as prepared by the Bureau and printed in the Radio Service Bulletin. The first printing will consist of a recapitulation of all references compiled since the first of 1928, to be followed by the monthly list, in each issue of the PROCEEDINGS.

### Papers in Pamphlet Form

On page XXXIV of this issue will be found a list of papers which are in pamphlet form and are available for distribution, free of charge, to members of the Institute. Kindly address requests to the Secretary, indicating the papers wanted numerically and *not* by titles. To non-members of the Institute these reprint copies are sold at fifty cents each.

### Errata

An error exists in Fig. 2 of the paper "Broadcast Control Operation" printed on page 502 of the April, 1928 issue of the PROCEEDINGS. The total resistances of the potentiometers shown should be paralleled and connected to the upper end of their corresponding coil secondaries so that the impedance facing the amplifier input remains sensibly constant, the audio input to any of the potentiometers varying according as the microphone in question is to play a greater or lesser part in the final output.

### Material on Formation of Sections

The 1927 and 1928 Committees on Sections have prepared a document outlining the steps to be taken in organizing a Section of the Institute, containing suggestions relative to the successful operation of a Section and describing many of the important functions of a Section. This material was prepared, primarily, for the information and guidance of members interested in the organization of a Section of the Institute. Members so interested may obtain a copy upon application to the Institute office.

## Institute Meetings

### New York Meeting

On May 2nd a meeting of the Institute was held in the Engineering Societies Building, 33 West 39th Street, New York. L. E. Whittemore, Vice-President of the Institute, presided.

K. S. Van Dyke presented two papers. The first, "The Piezo-Electric Resonator and Its Equivalent Network," is published in this issue of the Proceedings.

The second, "Some Experiments with Vibrating Quartz Spheres," is summarized as follows: Quartz spheres, mounted between parallel plate electrodes which may have much greater separation than the diameter of the sphere, respond as piezo-electric resonators to frequencies in the following ratios: 1, 1.47, 1.61, 1.83, 2.12, 2.18, and 4.12. Some of these response frequencies appear as doublets which depend for their relative intensity on the orientation of the sphere in the field. The lowest frequency corresponds to a radio wavelength of about 108 meters per millimeter diameter of the sphere. The decrement of the sphere is very small and the sphere behaves as an oscillator as well as a resonator for some of these modes. The geometry of the modes listed is indicated by the Giebe and Scheibe type of luminous discharge on, and near, the surface of the sphere when it is driven by a sufficiently intense field. Photographs of these discharge patterns were shown for the first three of the above modes for a sphere about 5 cm. in diameter. In strong electric fields in which the sphere is resonant, the sphere becomes very slippery to touch and slides around quite freely, without rolling, on a flat supporting electrode, or slides along a track without rolling, or runs around a ring. Though apparently quite still when resting on a smooth concave electrode, nevertheless, if centered by a small hole, it selects an axis for rotation and rotates steadily about this axis, the rotation being apparently due to the reaction of the sphere on the support rather than to any winds of the Meissner type which it may give off. Winds of the latter type are present, however, but in the vibration modes observed, apparently have no moment for rotation.

In the discussion which followed these papers, the following, among others, took part: F. K. Vreeland, L. E. Whittemore, K. S. Van Dyke, L. M. Hull, and W. C. Bohn.

It is hoped that the second paper may appear in some future issue of the PROCEEDINGS.

Some three hundred members and guests attended this meeting.

### ATLANTA SECTION

A meeting of the Atlanta Section was held on May 9, 1928 in the Civic Room of the Hotel Ansley, Atlanta. Preceding the meeting an informal dinner was held in the Ansley Hotel.

F. H. Schnell, of the Burgess Battery Company, presented a paper, "Aircraft Radio."

Officers to serve until December 31, 1928 were announced as follows: Chairman, Walter Van Nostrand; Vice-Chairman, D. C. Alexander; Secretary-Treasurer, George Llewellyn.

This was the last meeting of the Section until fall.

### BUFFALO-NIAGARA SECTION

A joint meeting of the Canadian Section, Rochester Section, and Buffalo-Niagara Section, to which members of the Niagara Frontier Section of the A. I. E. E. were invited, was held on April 11th in Foster Hall, University of Buffalo. L. C. F. Horle, Chairman of the Buffalo-Niagara Section, presided.

J. M. Thompson, of the Ferranti Meter and Transformer Company of Canada, presented a paper, "Characteristics of Output Transformers." The paper dealt with the operating characteristics of the output transformer, which are developed in terms of the known speaker, tube and transformer constants. In the first part of the paper the general formula for the speaker current is developed and the effect of varying the transformer constants shown. The turn-ratio of the transformer for maximum speaker current is considered in relation to the commonly used impedance ratio formula. The limitation of the impedance ratio formula is then pointed out and limits set for its general use. The general form of the current frequency characteristic for exponential horns and dynamic cone speakers is then obtained and a general method for matching the speaker to the output tube is given. In the latter part of the paper, curves are given to show the results obtained in the mathematical part of the paper. The curves also include the results of tests made in the Toronto laboratory of Ferranti, Ltd., to check the fundamental formula. The effect of the turn-ratio on the form of the current frequency characteristic is shown and a method of using the turn-ratio of the output transformer to match the speaker to the output tube is given. A perfect transformer is also compared with a good commercial transformer and the general effect of the leakage inductance and the self capacity of the transformer was shown.

Messrs. Manson, Million, Horle, Klumb, Hector, and others participated in the discussion which followed.

The second paper of the evening, "Shielding of Radio Receivers" was presented by M. L. Levy, of the Stromberg-Carlson Telephone Manufacturing Company. The paper pointed out the necessity for electrostatic and electromagnetic shielding in radio-frequency amplifiers of high gain. No attempt was made to explain why this shielding was necessary as it was the author's intention to cover practical applications of shielding to receivers of commercial types. With the aid of exceptionally well-prepared slides, Mr. Levy described the fundamentals involved in both electrostatic and electromagnetic shielding and how, in certain applications, combination effects were obtained by the use of the single shields. Then followed a description of several types of commercial receivers employing both types of shielding.

Messrs. Horle, Jones, Manson, Graham, Lidbury, and others participated in the ensuing discussion.

One hundred and seventeen members and guests attended this meeting.

### Canadian Section

In the Electrical Building of the University of Toronto a meeting of the Canadian Section was held on April 18th. A. M. Patience presided.

John P. Minton, of White Plains, New York, presented a paper, "Soft Magnetic Materials in Radio." Messrs. Patience, Thomson, Hepburn, Smith, Richardson, Mott, Price, Pipe, and others participated in the discussion which followed.

Forty-eight members and guests attended this meeting.

On May 2nd a meeting of the Canadian Section was held in the Electrical Building of the University of Toronto. A. M. Patience presided. Mr. Clark, of the Bell Telephone Company of Montreal, presented a paper, "Carrier Current Transmission."

Messrs. Patience, Bagly, Soucy, V. G. Smith, A. H. R. Smith, Lowry, Price, and others participated in the discussion.

Fifty-nine members and guests attended this meeting.

In the election of officers for the Canadian Section it was announced that the results were as follows: A. M. Patience, re-elected Chairman; V. G. Smith, Vice-Chairman; C. C. Meredith, Secretary-Treasurer; and J. M. Leslie, Assistant Secretary.

On April 18th a "Ladies' Night Dinner and Concert" was held in conjunction with the Toronto branch of the A. I. E. E.

### Cleveland Section

At the May 4th meeting of the Cleveland Section, held in the Case School of Applied Science, Cleveland, Professor John R. Martin presided. Five speakers were presented. H. W. Fay, of the Incandescent Lamp Division of the General Electric Company, gave a review of the recent I. R. E. paper on the UX-250 Power Tube. The essential features of the paper were set forth in a very clear and understandable manner. The reasons for the choice of the various design characteristics were explained. Comparison with the preceding power tubes, 210, 171, 112, and 120 was made.

M. V. King, of the Sterling Manufacturing Company, gave a review of the recent I. R. E. paper on "The Piezo-Electric Resonator and Its Equivalent Network." A striking feature was the simplicity of the equivalent circuit and the extremely low values of capacities involved. The sharpness of resonance obtained from the crystal oscillator was clearly shown from properties of the circuit.

Dean S. Kintner, radio editor of the *Cleveland Plain Dealer*, gave a "Will Rogers" talk on what a radio editor thinks about. Fortunately, a lot of it never gets into print. Some of the thoughts touched upon were the screen-grid tube, television, and broadcast programs.

Kelvin Smith, of the France Manufacturing Company, demonstrated the effect of suppressing certain frequencies, overloaded amplifiers and the meaning of transmission units by means of a new set of records issued by the Bell Telephone Laboratories. Some of the demonstrations bore a striking resemblance to the familiar "sidewalk" radio.

Ralph Worden, radio editor of the *Cleveland News*, gave a critical review of an article "Television Comes to the Home" which appeared in a recent radio periodical. The old familiar static will soon take on new terrors for the radio-looker of the future. The Kennelly-Heaviside layer will also soon come into prominence due to the appearance of a secondary or reflected image.

Forty-six members and guests attended the meeting.

In the discussions Messrs. Curtis, Catterall, Gimy, and others took part.

### Connecticut Valley Section

The Connecticut Valley Section, W. G. Cady, Chairman, presiding, held a meeting on April 26th in the auditorium of the Hartford Electric Light Association. B. J. Thompson, of the General Electric Company, presented a paper, "Characteristics and Uses of Screen-Grid Tubes." The paper centered around approximately fifty lantern slides which gave details of construction of the UX-222 screen-grid tube, as well as characteristics with different screen and plate voltages and with different bias voltages on the control grid. A very detailed account of how and why the tube acts as it does was given. Typical circuit diagrams for radio- and audio-frequency amplification were shown. The speaker mentioned the advantages of this type of tube over the standard three-element type for certain applications in radio reception, and the almost imperative use in the laboratory where circuits are required in which changes in the plate circuit must not affect the grid circuit.

Sixty members and guests attended this meeting.

### Los Angeles Section

On April 16th a meeting of the Los Angeles Section was held in Los Angeles. Dr. R. C. Burt, of California Institute of Technology, presented a paper on "The Photo Electric Cell." The paper pointed out, among other things, that the action of a photo electric cell can, so far, be explained only by the corpuscular theory, while the phenomenon of radio, light, and heat waves requires the wave theory. It is probable that both theories are correct, but the exact relation or connection is not known. The construction of the cell was explained, and some of its uses described. Its uses as a light-operated relay are almost innumerable.

### Philadelphia Section

A meeting of the Philadelphia Section was held on April 27th in the Bartol Laboratories of the Franklin Institute. J. C. Van

Horn presided. E. L. Nelson, of the Bell Telephone Laboratories, presented a paper on "Some Recent Developments in Radio Broadcasting Apparatus," which among other things explained in detail the development work on some of the transmitting equipment in the experimental development group of stations located at Whippany, New Jersey.

Messrs. Wilson, Darlington, Frazier and others discussed the paper.

Eighty members and guests attended the meeting.

## San Francisco Section

The meeting of the San Francisco Section of March 14th was held in the Club Rooms of the Engineers' Club, preceded by a dinner attended by twenty-five members and guests. L. F. Fuller presided at the meeting. The speaker, A. H. Saxton, operating engineer of the Pacific Division of the National Broadcasting Company, delivered a paper on "The Tendencies of Radio Broadcasting." The paper was centered around the organization and operation of the National Broadcasting Company. The general layout of apparatus used in conjunction with the studios was described, the speaker explaining how different studios could be connected to different channels. The necessity of careful checking of facilities was pointed out. The system used by the Westinghouse Electric and Manufacturing Company in operating two stations on the same wavelength was briefly touched upon.

In the discussion which followed Messrs. Fuller, Lippincott, Brown, and others took part.

Following the meeting the members visited the studios of the National Broadcasting Company in the Hunter-Dulin Building. From thence they proceeded to the control room of the Pacific Telephone and Telegraph Company where, through the courtesy of C. H. Cole of that organization, a description of the apparatus used, together with demonstrations, was provided.

## Seattle Section

The Seattle Section held a meeting on March 31st. W. A. Kleist presided. Major H. C. K. Muhlenberg presented a paper on "Radio in Aviation." The paper covered the following points: A review of radio communication between planes and between planes and the ground; relative advantages of telegraph and telephone communication; physical problems of generator location on plane; beacon flying with particular emphasis upon accuracy, dependability, and distances covered; shielding effect of metal planes.

Messrs. Wilson, Kleist, Williams, Mason, Renfro, and others discussed the paper.

Twenty members of the Section were present.

## Committee Work

### Sectional Committee on Radio, A. E. S. C.

The following were recently appointed officers of the Sectional Committee on Radio, American Engineering Standards Committee: Alfred N. Goldsmith, Chairman; C. H. Sharp, Vice-Chairman, and Laurens E. Whittemore, Acting Secretary.

### Committee on Membership

A meeting of the Committee on Membership was held on May 2nd in the Fraternities Club Grill. The following were present: H. F. Dart, Chairman; I. S. Coggeshall, H. B. Coxhead, and F. R. Brick. The Committee considered a number of plans looking to a continued increase in membership in the Institute throughout the year. Its next meeting will be held on June 6th.

### Committee on Admissions

At its meeting held in the office of the Institute on May 2nd, the following members of the Committee on Admissions were present: R. A. Heising, Chairman; Lewis M. Hull, E. R. Shute, F. K. Vreeland. Approximately thirty-five applications for transfer or election to the higher grades of membership in the Institute were considered. The amount of material on hand for consideration by the Committee during the past few months has been such that it has been necessary for the Committee to work a number of hours at each meeting to consider all of the applications.

### Committee on Constitution and Law

The first meeting of the newly organized Committee on Constitution and Law was held at dinner at the Fraternities Club on May 2nd. The following were present: R. H. Marriott, Chairman; E. N. Curtis, H. E. Hallborg, and G. W. Pickard. Most of the meeting was occupied in outlining the future work of the Committee. The Committee is proceeding immediately with the revision of the Constitution.

## Personal Mention

C. F. Donbar is Chief Engineer of station KQV of Pittsburgh. Mr. Donbar was formerly with the Bell Telephone Company.

J. P. Putnam, recently engineer, radio department of the Martin Copeland Company, is now employed as engineer with the National Company of Malden, Mass.

F. B. Sternberg, formerly director and manager of the Dallas Radio Laboratories is now associated with the P. F. Collier and Son Company in the Newark Branch.

Captain Guy Hill, attached to the office of the Chief Signal Officer, War Department, has been detailed to duty with the Federal Radio Commission as technical advisor.

E. H. Guilford, late Chief Engineer of the Radiore Company of Los Angeles, has become General Manager of the Radiore Company of Canada, Limited, with Headquarters in Montreal.

Captain S. C. Hooper, Head of the Radio Division of the Bureau of Engineering, Navy Department, has been detailed to duty with the Federal Radio Commission in the capacity of technical advisor.

Charles M. Kelly, Jr. has recently become associated with the Electrical Research Products Corporation of New York City. Mr. Kelly was formerly assistant sales manager of the Amrad organization.

O. M. Dunning has recently joined the staff of the Acoustic Products Manufacturing Corporation of Stamford, doing general development work on both receivers and electrical phonograph equipment.

Daniel E. Harnett has recently become associated with the Acoustic Products Manufacturing Corporation of Stamford, Connecticut in the design of radio parts. He was formerly in charge of production of the Murdock Company.

# BEAM TRANSMISSION OF ULTRA SHORT WAVES*

By

Hidetsugu Yagi

(College of Engineering, Tohoku Imperial University, Sendai, Japan)

***Summary***—*Part I of this paper is devoted to a description of various experiments performed at wavelengths below 200 cm. Curves are given to show the effect of the earth and various types of inductively excited antennas called "wave directors." Part I is concluded with a discussion of beam and horizontally polarized radiation.*

*Part II is devoted chiefly to the magnetron tubes used for the production of very short wavelengths (as low as 12 cm.) and the circuit arrangements employed. It is shown that the geometry of the tube and its external connections are of great importance.*

*The effect of variation of plate voltage, magnetic field strength and other factors on the high-frequency output, is described.*

## Introduction

THE general term "short wave" loses much of its lucidness when the range of frequency involved is considered. For this reason, the term "ultra short waves" will apply to only those electro-magnetic waves whose length is less than ten meters.

One of the simplest ways of generating short waves by means of vacuum tubes is to use the push-pull circuit developed by M. Mesny. This connection has been fully described by Mr. Englund in the PROCEEDINGS of the Institute.

Waves shorter than ten meters may be produced with stability, but it is difficult to make ordinary tubes operate satisfactorily below two meters. While electro-magnetic coupling is successfully used in the method referred to above, it seems much better to resort to electrostatic coupling in circuits used for the generation of waves of the length described in this paper. Fig. 1 shows a circuit which has been used in the generation of waves shorter than 100 cm.

Stable oscillations were successfully produced using ordinary tubes in this circuit. Such waves have been utilized to determine the natural frequencies of the various forms of metallic bodies. The characteristics of "wave directors", which will be fully described later in the paper, were thoroughly studied with the short waves produced using this type of generator. However, it was

* Original Manuscript Received by the Institute, January 30, 1928; Revised Manuscript Received by the Institute, March 29, 1928. Presented before meetings of the Institute in New York, Washington and Hartford.

impossible to generate waves shorter than 60 cm. even with this circuit using electrostatic coupling within the tubes.

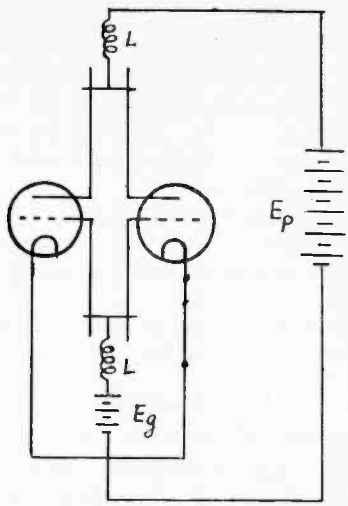



Fig. 1—Circuit Diagram of Oscillator; 60-200 cm.

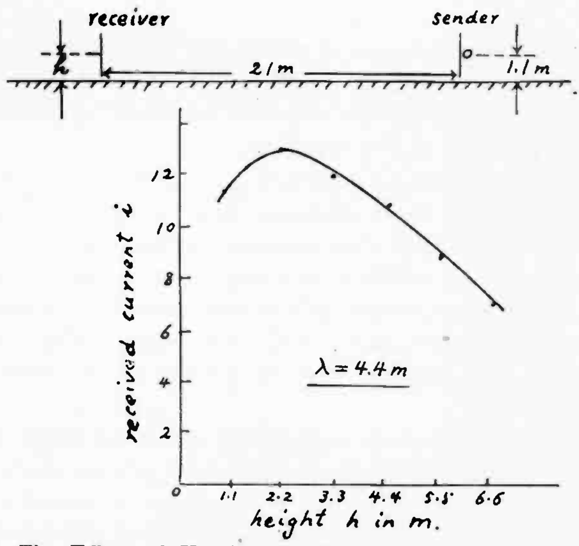



Fig. 2—The Effect of Varying Receiver Antenna Height. Sending Antenna Height Equals 1.1 m.

The method of Barkhausen and Kurz enables one to obtain much shorter waves. By this method, it was possible to reduce the minimum wavelength to 36 cm. using plate voltages in the order of 300 volts. Schafer and Merzkirch obtained waves of the order of 34 cm. with a plate voltage of 350 volts, and Scheibe has reported a stable minimum of 30 cm. With somewhat less stability, he has produced waves 24 cm. long.

Mr. K. Okabe, assistant professor at the Tohoku Imperial University, has succeeded in generating exceedingly short, sustained waves by introducing certain modifications in the so-called magnetron. These waves are the shortest which it has been

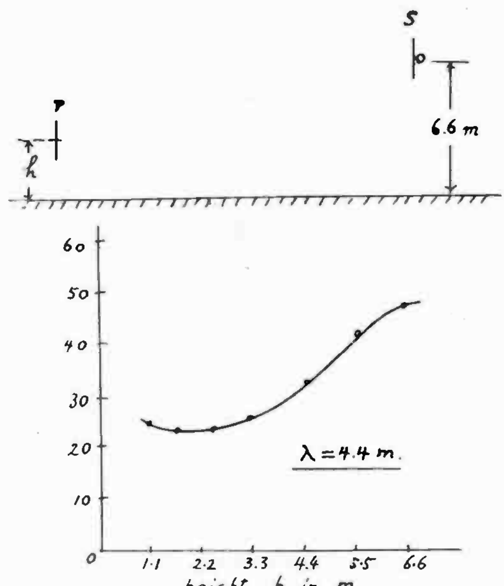



Fig. 3—The Effect of Varying Receiving Antenna Height. Sending Antenna Height Equals 6.6 m.

possible to generate so far as the author is aware. He was able to produce fairly strong radiation at a wavelength of 12 cm. and, by the use of harmonics, was able to obtain a minimum of 8 cm. The practical application of these ultra short waves will be dealt with in Part II of the paper.

## Part I

### Beam Radiation for 4-Meter Waves

Mr. S. Uda, assistant professor at the Tohoku Imperial University, has published nine papers in the *Journal* of the I.E.E. of Japan on beam radiation at a wavelength of 4.4 meters. Several papers by Mr. Uda and the author have been presented at the Imperial Academy of Japan and the Third Pan-Pacific Science

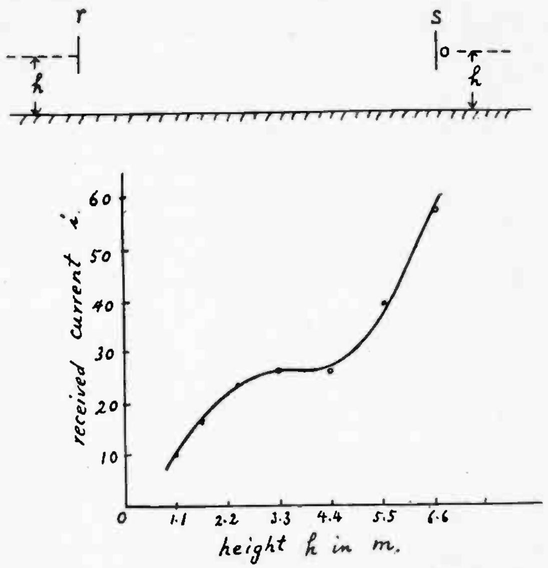



Fig. 4—The Effect of Varying Both Sending and Receiving Antenna Height Simultaneously.

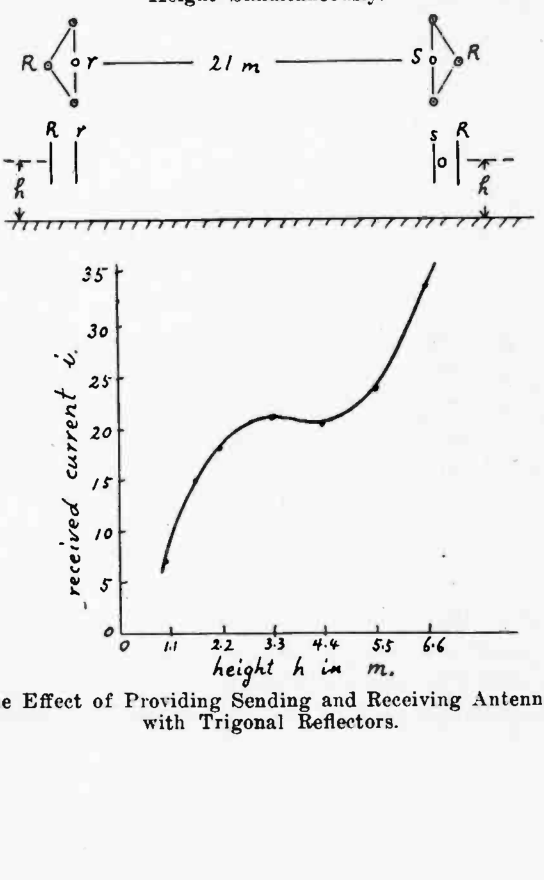



Fig. 5—The Effect of Providing Sending and Receiving Antenna in Fig. 4 with Trigonal Reflectors.

Congress held in Tokyo in 1926. In the following description, some of the much more notable points of the beam system used in this work will be explained. The photographs show some of the actual apparatus used.

## Wave Reflectors and Directors

Suppose that a vertical antenna is radiating electro-magnetic waves in all directions. If a straight oscillating system, whether it be a metal rod of finite length or an antenna with capacities at

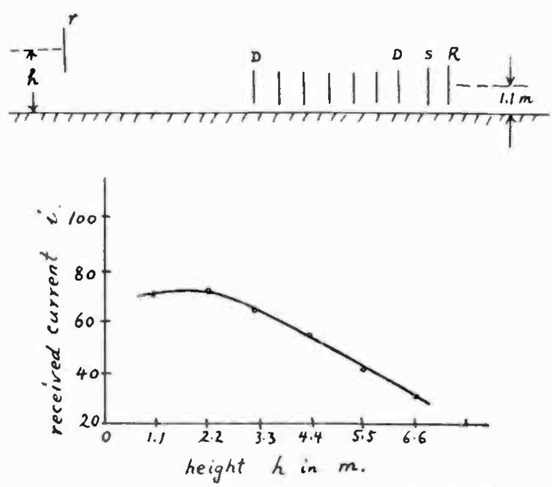



Fig. 6—The Effect of Varying Receiver Antenna Height When Wave Canal Is Applied at Sending Antenna. Sending Antenna Height Equals 1.1 m.

both ends and an inductance at the middle, is erected vertically in the field, the effect of this oscillator upon the wave will be as follows: If its natural frequency is equal to or lower than that of the incident wave, it will act as a "wave reflector." If, on the other hand, its natural frequency is higher than that of the incident wave, it will act as a "wave director." The field will converge upon this antenna, and radiation in a plane normal to it will be augmented. By utilizing this wave-directing quality, a sharp beam may be produced.

A triangle formed of three or five antennas erected behind the main or radiating antenna will act as a reflector. This system will be called a "trigonal reflector." In front of the radiating antenna, a number of wave-directors may be arranged along the line of propagation. By properly adjusting the distance between the wave-directors and their natural frequencies, it is possible to transmit a larger part of the energy in the wave along the row of directors. Adjustment of the natural frequency of the directors is made by simply changing their length or by adjusting the inductance inserted at the middle of these antennas.

The number of wave-directors has a very marked effect on the sharpness of the beam, the larger number of directors producing the sharper beam. It has been found convenient to designate such a row of directors as a "wave canal."

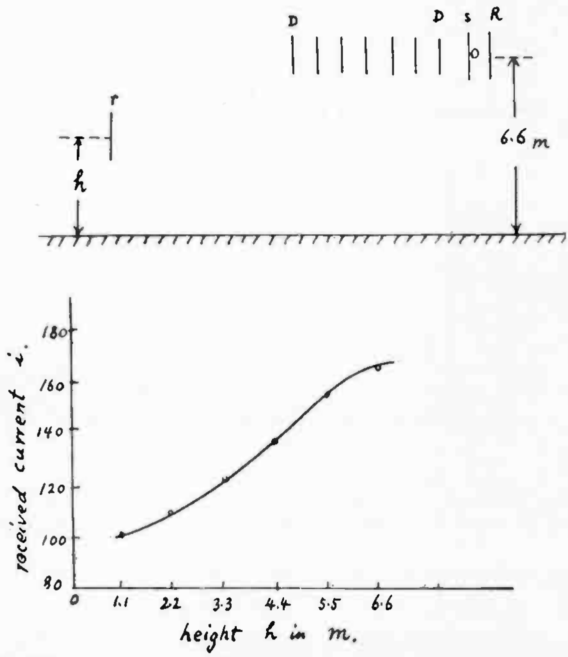



Fig. 7—The Effect of Varying Receiver Antenna Height When Wave Canal Is Applied at Sending Antenna. Sending Antenna Height 6.6 m.

The trigonal reflector and wave canal may also be employed at the receiving station. In this case, the reflector will be called "collector." Here again, the effect of the directors and the wave canal has been found to be considerable.

## Radio Beacon

These principles may be used in a radio beacon, by which a beam may be projected in any direction. This is not done by altering the position of the antennas or by revolving the whole system. A number of antennas which are fixed in position are employed and so arranged that their natural frequencies may be altered between two values. Thus, it may be made either a reflector or a director, depending upon its natural frequency.

The main or radiating antenna is situated at the center, and the others which are used for reflecting or directing the beam are located on two concentric circles whose radii are 1/4 and 1/2

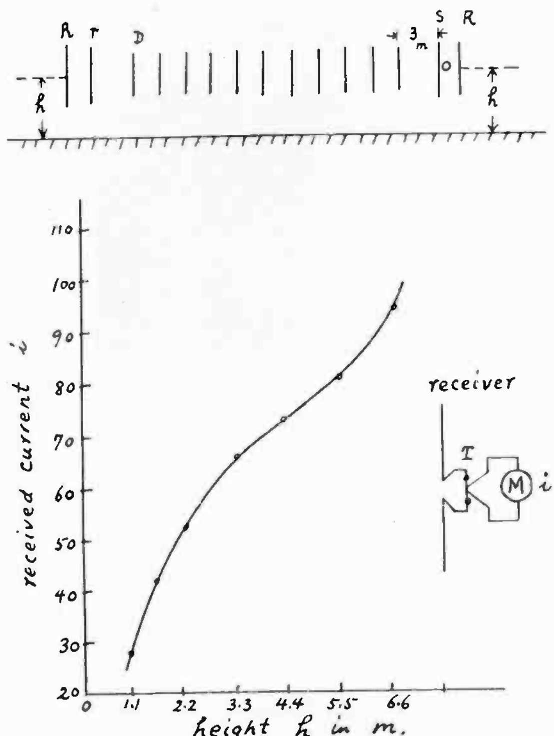



Fig. 8—The Effect of Varying Sending and Receiving Antenna Simultaneously When Wave Canal Is Located between Two Antennas.

wavelength respectively. The direction of radiation may be changed at will by properly controlling the functions of the antennas on these two concentric circles; that is, certain of them are made to act as reflectors while others are made to act as directors of the electro-magnetic wave.

## Radio Beacon Transmission

If the sending and the receiving antennas are both surrounded by reflecting systems, and these two structures, which

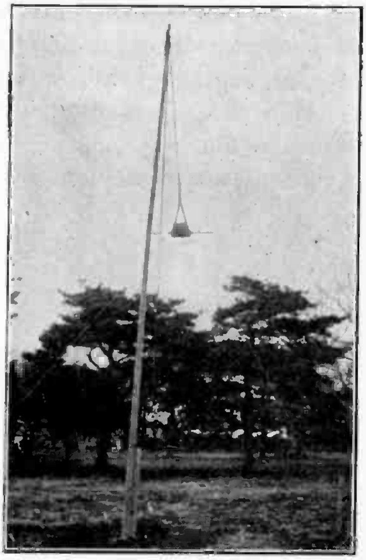


Fig. 9—Horizontally Polarized Wave Receiver in the Air.

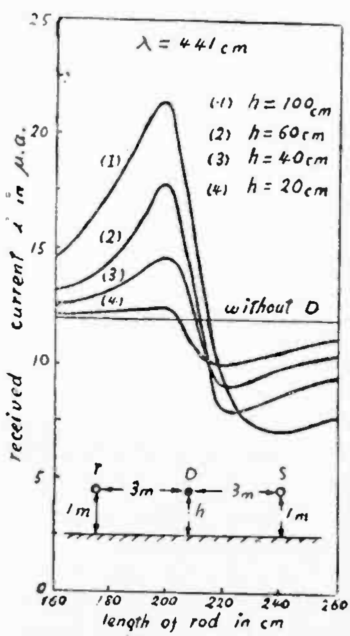
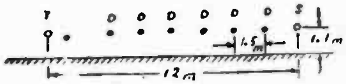



Fig. 10—The Effect of Varying Length and Height of Wave Director on Received Current.

are directed toward one another, are joined by a wave canal, the radio-frequency energy may be directed back and forth along this canal. All the directors forming the canal will have induced oscillations but the intensity and phase displacement will, in



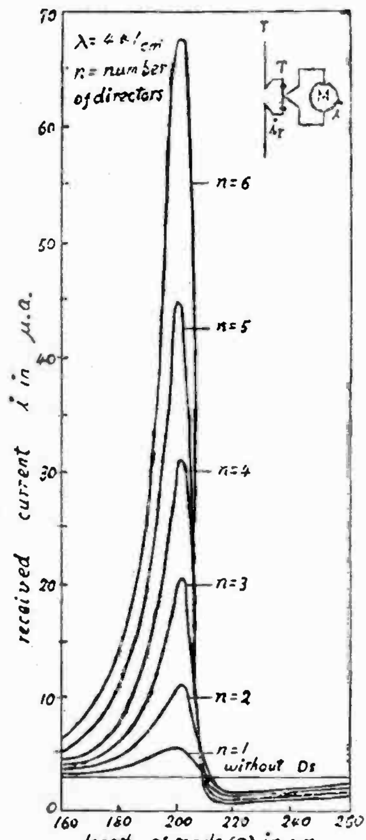




Fig. 11a Fig. 11b
The Effect of Varying Number and Length of Directors in Wave Canals on Received Current.

general, be different. A sort of standing wave will exist along the canal and the power will flow at a definite rate from the sending to the receiving station.

The wave energy received can be rectified by means of vacuum tubes or otherwise, and thus it may be used to charge a storage battery. It has been the experience of the author that rectification is very easily obtained even at very short waves.

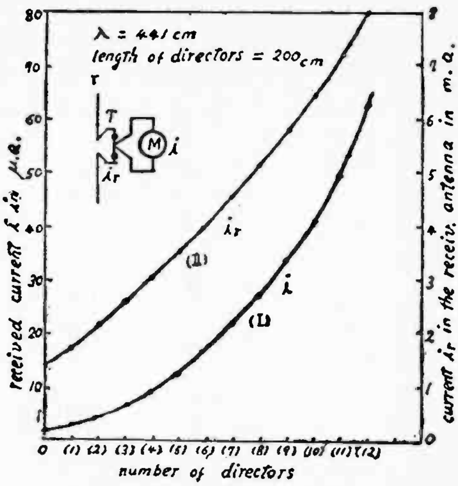



Fig. 12—The Effect of Varying Number of Directors on Received Current.

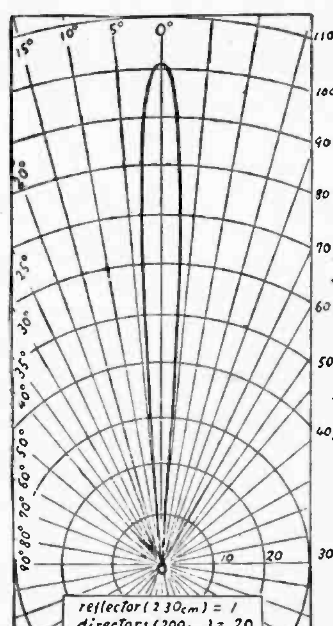



Fig. 13—Beam Radiation from a Radiator Utilizing a Wave Canal.

It appears that the wave collector at the receiving station may suppress to some extent the flow of energy from the sending antenna. It was found that in certain cases it was possible to transmit more power when a certain number of the directors in the middle of the wave canal were removed.

## Effect of the Earth

In ultra short wave work, the effect of the earth is very considerable. Some of the experimental results are given below to illustrate this. Figs. 2 to 8, which are self-explanatory, are for various conditions of transmitter and receiver antenna height, with and without trigonal reflectors and wave canals.

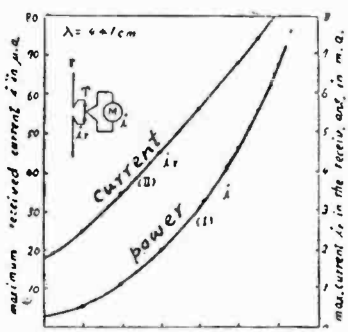



Fig. 14—The Effect of the Number of Directors on Received Current and Power.

It is interesting to note that the energy transmitted increases or is considerably increased when the height of the entire system is increased. As yet, no limit has been found for this effect.

## Projector of Horizontally Polarized Waves

A radiating antenna placed horizontally with the earth is naturally directive. The wave is radiated chiefly in a vertical plane bisecting the antenna and perpendicular to it. Various polar diagrams were taken with such an antenna using a receiving antenna such as is shown in the accompanying photograph, Fig. 9. A thermocouple and micro-ammeter located at the middle of this antenna were used to indicate the magnitude of the received power.

The results of these experiments are shown in Figs. 10 to 14. In Fig. 10, *S* and *R* are the sender and receiver respectively, while *D* is a wave director. The effect of varying the length of *D* on received energy is very pronounced, and is a maximum of about 200 cm., whereas, in the case of vertically polarized waves, this maximum occurred between 190 and 195 cm.

In Fig. 11a, a wave canal is introduced between the sending and receiving antennas and the effect of varying the length of all of the directors is more pronounced than was the case in Fig. 10.

In general, the effect of increasing the number of directors forming the canal is shown in Fig. 12, where $i$ is the current in the indicating meter and $i_r$ is the current in the antenna.

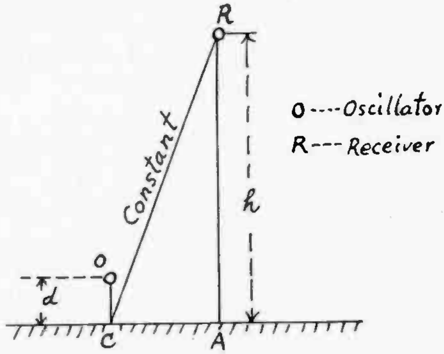



Fig. 15—Diagram Showing the Location of Antenna for Field Strength Measurements.

The length of the directors must be accurately adjusted; otherwise successful directing action will not be obtained. It has been found that the interval between adjacent directors must be adjusted to a suitable value. The most advantageous value for this interval seems to be approximately 3/8 wavelength.

A typical polar curve showing the beam radiation from such a projector is given in Fig. 13. The measurements were taken on a horizontal plane near the earth's surface. Here again, the advantage of utilizing the wave canal at the receiving station is demonstrated to be quite remarkable.

It has been found that power received increases nearly proportional to the square of the number of directors forming the canal. This effect is shown in the experimentally determined curves of Fig. 14.

## High-Angle Radiation of Horizontally Polarized Waves

Some experiments were performed in which the field strength around the sending antenna $O$ was measured by a receiving antenna $R$. The distance $CR$ from $R$ to a point on the surface of the earth directly beneath the sending antenna was kept constant. The wavelength employed was approximately 260 cm. and the length of the sending antenna $O$ was 135 cm. Fig. 15 shows the arrangement.

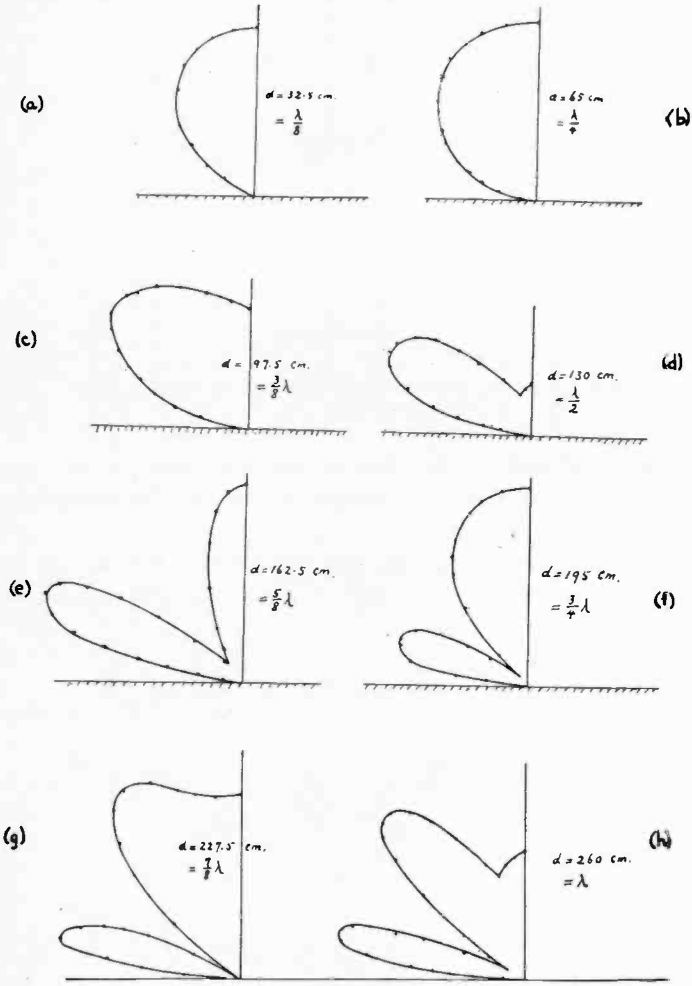



Fig. 16—Polar Diagrams.

The earth seems to act very much like a mirror to ultra-short waves, and reflection from its surface depends upon distance $d$ between antenna and earth as shown in Fig. 15. The experimentally determined polar diagrams shown in Fig. 16, (a), (b), (c), (d), (e), (f), (g), (h), illustrate this fact very well.

The effect of a wave canal upon high-angle radiation of horizontally polarized waves was then studied. A canal was arranged parallel to the surface of the earth in the first case and

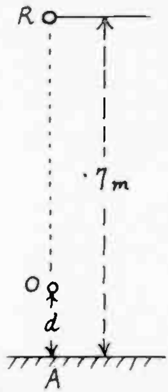



Fig. 17—Diagram Showing Location of Sending and Receiving Antenna for Fig. 18

along the line inclined 30 deg. to the horizontal in the second case. The actual set-up is shown in the two following photographs, Figs. 19 and 20.

It is evident from Fig. 21 that the canal is forcing the beam toward the horizontal direction. Thus, by the use of wave

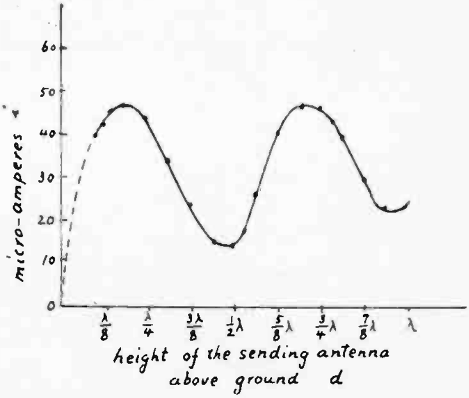



Fig. 18—The Effect of Height of Sending Antenna on Receiver Current.

canals, high-angle radiation may be propagated at various angles to the surface of the earth. This fact may find some practical application in long distance work.

## Theory

Theoretical calculations concerning the various experiments described above are naturally involved. Some of the previously mentioned papers presented to the I.E.E. of Japan contain theoretical descriptions of the research. Certain fundamental theories are to be found in a paper which will be published at some later date by the I.E.E. of Japan. This paper will be in English.

# Part II

## Magnetron Oscillators

A diode is capable of producing oscillations if the anode is a circular cylinder and the cathode is a straight filament at the

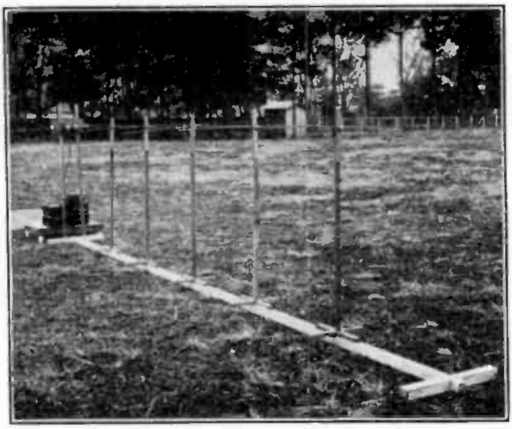


Fig. 19—Projector Horizontally Polarized Wave; 260 cm.

center, with the tube placed in a uniform magnetic field the direction of which coincides with the direction of the axis of the cylinder. When the strength of the magnetic field is increased past a critical value no current should flow through the vacuum tube because the electrons emitted from the filament and attracted by the anode describe circular orbits the diameter of which is less than the radius of the anode. However, when this is tried experimentally sometimes there is residual current flowing to the anode which can be detected by a hot wire instrument. This is evidence of the existence of high-frequency currents.

It has been found that any of the diodes or the triode shown in Fig. 23 can produce short-wave oscillations when sufficiently high anode voltage is applied and a magnetic field of appropriate intensity is employed. In order, however, that the oscillations be of extremely short wavelength with sufficient intensity, symmetrical construction and exact dimensioning are essential.

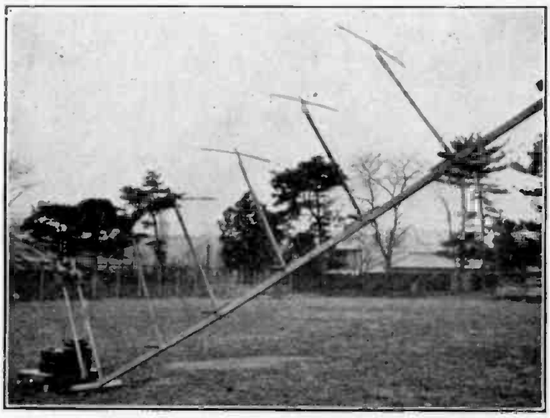


Fig. 20—High Angle Projector of the Horizontally Polarized Wave. 260 cm.

It has been found that the wavelength can be calculated roughly by the following semi-theoretical formula:

$\lambda_0 = 2\,ct$

$\lambda_0$ = semi-theoretical wavelength

Where $c$ = velocity of light

$t$ = the time required by an electron for travelling across the space between the cathode and the anode

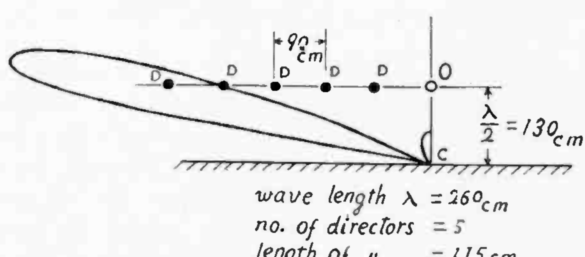



Fig. 21—Polar Diagram with Wave Canal Parallel to Surface of the Earth.

The results are given in the following tabulation. The second column gives the wavelength as measured by the Lecher wire method. The wavelength was practically independent of filament temperature.

MAGNETRON II

| $D_a = 1.32$ cm. Vacuum: | $L_a = 2.5$ cm. $10^{-3}$/bar. | $I_f = 3.5$ amp. Ni-Anode | $l_f = 3$ cm W-Fil. |
|---|---|---|---|
| Anode-Voltage (Volts) | λ (cm.) | Intensity of the Oscillations | $\lambda_e$ (cm.) |
| 190 | 150 | Weak | 87 |
| 230 | 122 | " | 79 |
| 280 | 88 | Middle | 72 |
| 450 | 63 | Strong | 58 |
| 500 | ..... | ........... | 55 |
| 1000 | 32 | Very Strong | 38 |
| 1300 | 26.5 | " " | 35 |
| 5000 | ..... | ........... | 17 |
| 20000 | ..... | ........... | 8.5 |

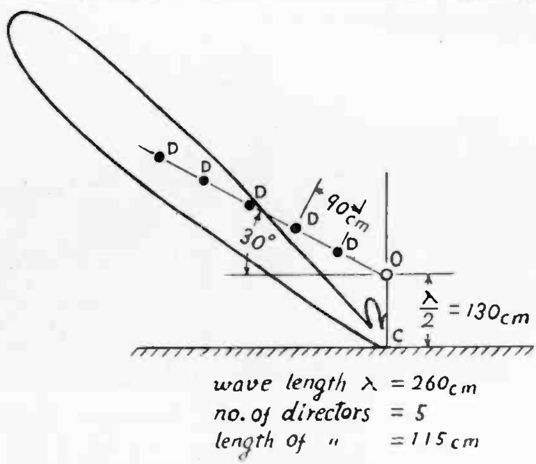



Fig. 22—Polar Diagram with Wave Canal at 30 deg. to Surface of the Earth.

The arrangement of the apparatus is shown in Fig. 24.

The variation of anode current with the magnetic field for a typical tube is shown in Fig. 25. Above the critical magnetic field strength there was still some current flowing which was a

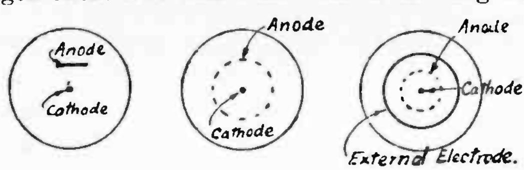



Fig. 23—Types of Diodes and Triode Used Experimentally.

result of the high-frequency oscillations. The most intense oscillation occurred at or near the critical field strength. The oscillations seemed to weaken with increasing magnetization.

When the anode diameter was kept constant larger diameter filaments seemed to give stronger oscillations. Moreover, with larger filament diameters, the anode current most favorable to the production of oscillations was smaller, which is decidedly an advantage.

To get the shortest waves, the anode diameter must be small. The result, however, is that the oscillations become less intense. It was found that the actual length of the anode must not be too short in proportion to its diameter, otherwise the oscillations were very feeble.

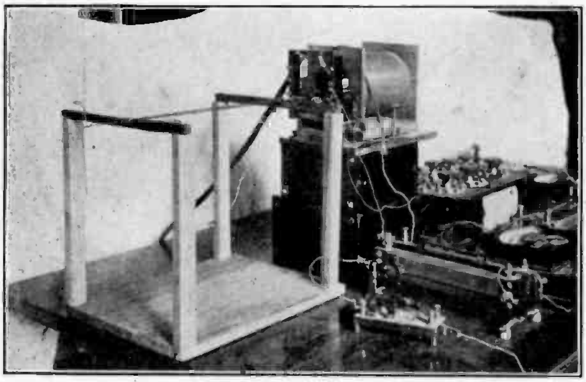


Fig. 24—Apparatus for 26.5-150 cm. Wavelength Production.

The position of the tube in the magnetic field is very important. It was found highly desirable to keep the tube in the most uniform portion of the field. As shown by Fig. 26, a slight deviation from the exact center of the magnetic field coil caused a marked decrease in the oscillation intensity.

## Shortest Waves Obtained

Two special tubes of small dimensions were constructed and tried.

| | | |
|---|---|---|
| No. I | $Da = 4.5$ mm. | $Df = 0.14$ mm. |
| No. II | $Da = 2.2$ mm. | $Df = 0.07$ mm. |

where $Da$ = anode diameter and $Df$ = filament diameter. For the test each tube was placed between the poles of a large electromagnet as shown in Fig. 27.

The relation between the anode voltage and the wavelength for tube No. I is shown in Fig. 28. Tube No. II gave a wave of 19 cm. with 840 volts on the anode and a minimum wavelength of 12 cm. with 1250 volts on the anode. These values of

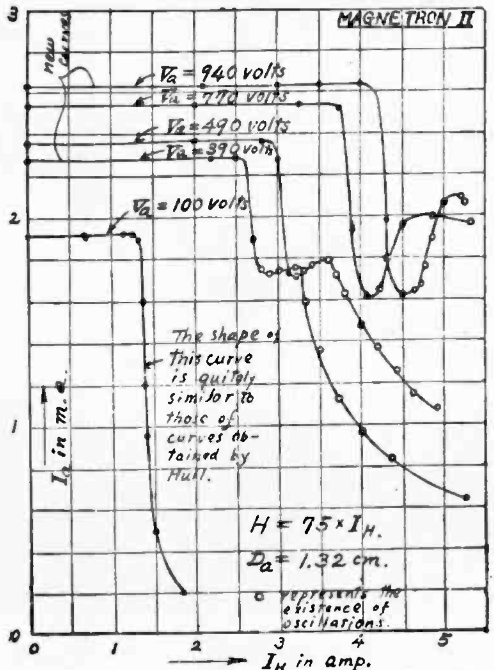



Fig. 25—Variation of Anode Current with Strength of Magnetic Field.

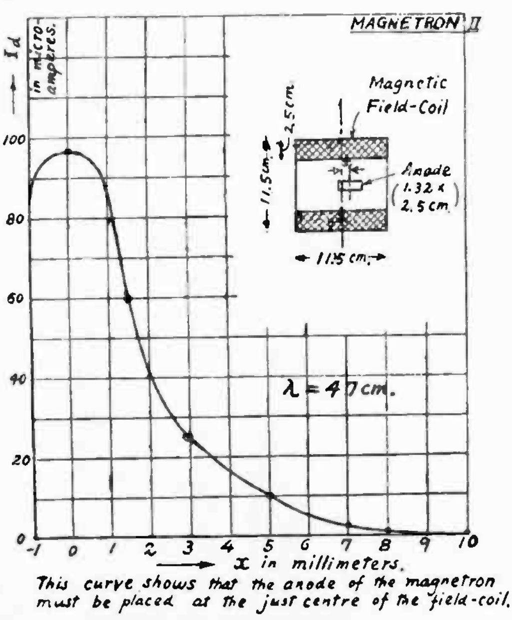



Fig. 26—Variation of Oscillation Intensity with Tube Position in Magnetic Field.

wavelength, however, do not agree very well with the semi-theoretical formula.

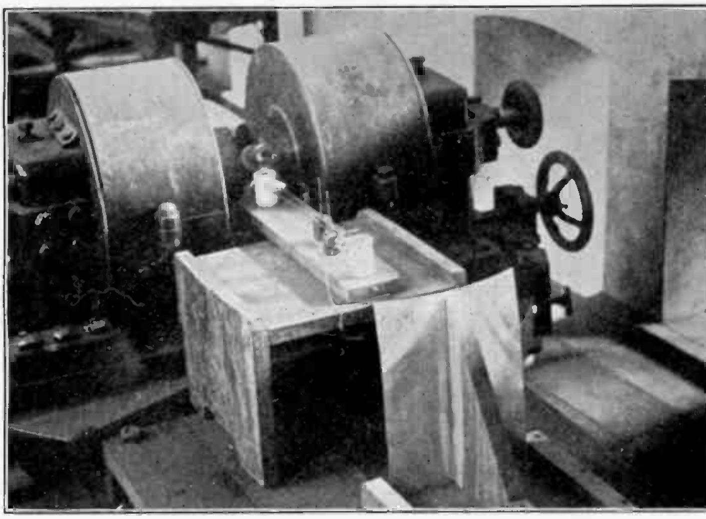


Fig. 27—Apparatus Set-up for the Shortest Waves Obtained (14-15 cm.).

The measurement of the wavelength on Lecher wires was not easy. Too strong a magnetic field seemed to disturb the

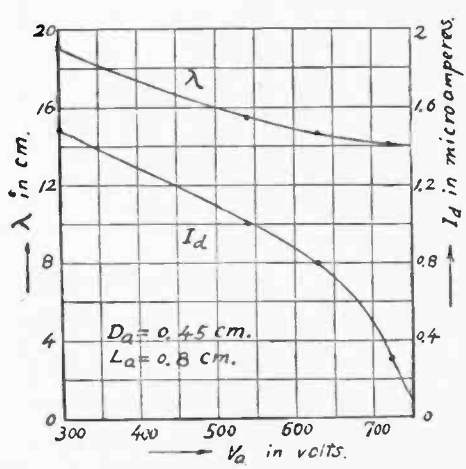



Fig. 28—Variation of Wavelength with Voltage.

steadiness of the oscillations and it was difficult to obtain the shorter waves as a fundamental oscillation. The stronger magnetic field had a tendency to produce oscillations rich in harmonics.

The most fruitful improvement made was to split up the cylindrical anode into two or more segments by narrow slits cut parallel to the axis of the cylinder. Fig. 29 shows the two-segment type and Fig. 30 the four-segment type of tube. Instead of bringing only one anode lead out of the tube a lead was brought out for each segment. These leads were then

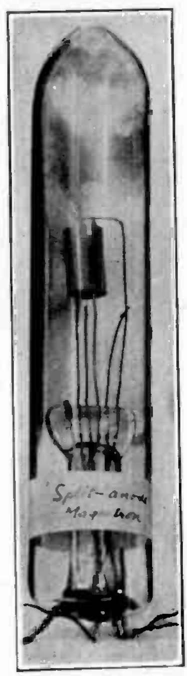


Fig. 29—Split-Anode Magnetron; 2-segment type.

brought together outside of the tube without directly touching each other and brought close to the cathode lead at a point $B$ in Fig. 31. After that the leads were all connected and led to the positive terminal of the high-voltage anode battery.

Each anode segment with its leads seems to form a resonant circuit, the natural frequency of which may vary with the length of the lead and the capacity of the segment.

The distance between the anode leads and the cathode lead must also be adjusted at the point $B$, so that maximum oscillation intensity may be obtained. Now, owing to the tuning action of these resonant circuits, the change of wavelength, due to the change of anode voltage, became inappreciably small. The

Fig. 30—Split-Anode Magnetron; 4-segment type.

wavelength was determined either by a Lecher system or by a receiving set used to indicate standing electromagnetic waves formed before a sheet metal screen.

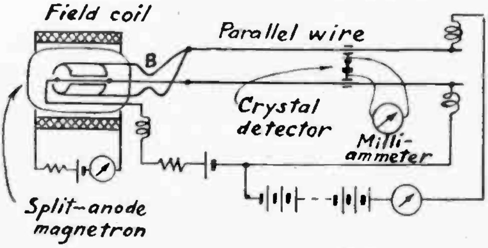



Fig. 31—Circuit Connection for Short Wave Oscillating Magnetron.

## 40 cm. Waves

A split anode magnetron was found to be especially well suited for the production of very intense oscillations of about 40 cm. wavelength. A typical case is given in the following table.

$Da = 14$ mm. $\qquad La = 26$ mm.
$Df = 0.14$ mm. $\qquad Lf = 30$ mm.

where $La$ = length of anode and $Lf$ = length of filament.

| Anode Voltage | Wavelength | Intensity (Arbitrary) |
|---|---|---|
| 951 | 34.5 | 5.3 |
| 724 | 41.5 | 15.5 |
| 670 | 42 | 16 |
| 500 | 42.5 | 6.7 |
| 400 | 42.5 | 4 |
| 320 | 42.5 | 1.8 |

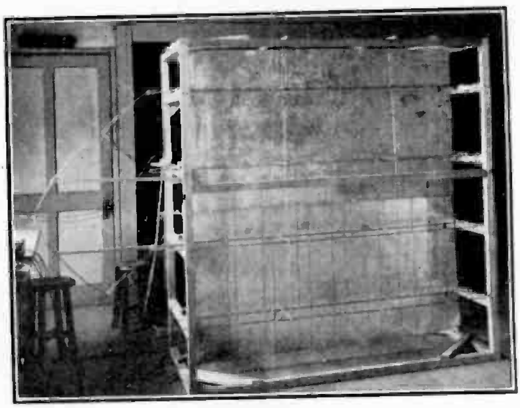


Fig. 32—Antenna System for 40 cm. Wave Transmitter.

The apparatus used in this experiment is shown in Fig. 32 (front view) and Fig. 33 (rear view).

Fig. 33—Magnetron Oscillator for 40 cm. Wave Transmitter.

In order to obtain various directive effects, antenna systems, as shown in Fig. 34, may be used and several of these may be combined, using metal plates as reflectors; or groups of reflectors as shown in Fig. 35 may be used with parabolic reflectors of sheet metal. Fig. 24 shows a radiating system of

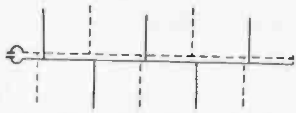


Fig. 34—Directive Antenna, 40 cm.

the type given in Fig. 34. The polar diagrams (Fig. 36) of the antenna system can be calculated from the arrangement of the various elements. The actual measurements showed good

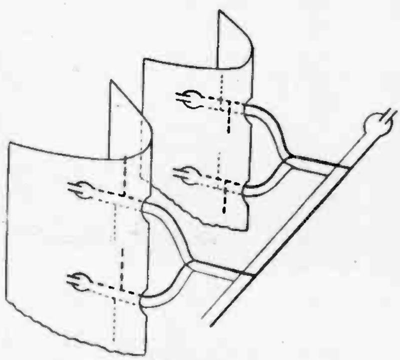


Fig. 35—Directive Antennas.

agreement with the theoretical values and the beam was confined within a small angle in the horizontal and vertical planes.

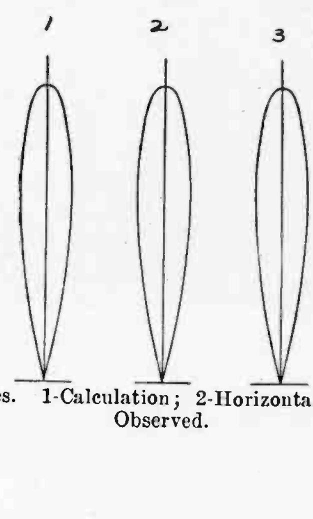



Fig. 36—Polar Curves. 1-Calculation; 2-Horizontal Observed; 3-Vertical Observed.

## Reception of Short Waves

For reception a crystal detector or thermocouple attached at the center of a straight antenna can be used. The currents from several such detectors may be combined in parallel or in series, according to the circumstances.

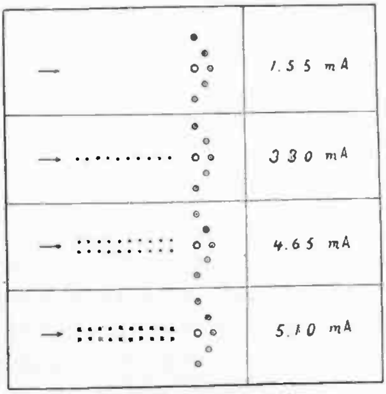



Fig. 37—Reception with a Collector and Wave Canals.

To increase the signal strength a wave collector was built, but its effect was not as remarkable as that of the wave canal. Wave directors on the transmitter proved to be astonishingly advantageous. It was found necessary to use a wave director in order to transmit signals at this short wavelength. The effect of the wave canal at the receiver is shown in Fig. 37.

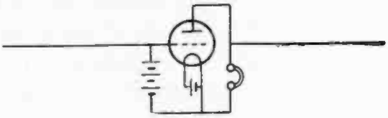


Fig. 38—Receiver (Barkhausen).

The maximum distance covered has so far not exceeded 1 kilometer. In this 1-kilometer experiment the wave was modulated at 900 cycles per second. The exact wavelength was 41 cm. and the anode voltage 1000. A single Hertzian resonator with a crystal detector at the center, a double row of 12-meter director chains and a three-stage amplifier for the modnlation frequency were employed. The results are shown in the following table:

| | |
|---|---|
| 1. Single Hertz resonator only | Signal not heard |
| 2. One director chain, no amplifier | Very weak |

| | |
|---|---|
| 3. Two director chains, no amplifier | Very weak |
| 4. One chain, with amplifier | Loud and clear |
| 5. Two chains, with amplifier | Loud and clear |

The type of receiver suggested by Barkhausen[1] and shown in Fig. 38 was also tried and gave better results than the crystal detector in detecting modulated waves in the neighborhood of $1\frac{1}{2}$ meters.

The experiments described in Part I were made by Mr. S. Uda, and those in Part II by Mr. K. Okabe, to the ingenuity of both of whom the successful development of the beam system is mainly due.

## Discussion

**J. H. Dellinger:**† Professor Yagi's remarkable work stimulates some thought of a radical order. I venture to suggest that before many years radio operations will generally be considered as divided into two classes, broadcasting and directive radio. Radio communication is to a large extent done the wrong way today. And before 1920 radio was all wrong. The only use of radio was for communication between two points, and it was always done by broadcasting in every direction. It was not until 1920 that we had the advent of broadcasting as such, transmission intended for reception by large numbers of receivers. In the eight years since 1920 we have successfully developed broadcasting. At present, therefore, the job of straightening radio out is half done.

It is interesting that 1920 marks not only the rise of broadcasting but also the beginning of directive radio. Ideally, radio transmissions should be broadcast in every direction only when intended for reception in every direction, and should be sent as nearly as possible in one line when intended for reception by one receiver. Since 1920 we have had the gradual and partial evolution of beam systems and other means of confining a communication more or less to the path desired. One instance is the use of a string of relay stations. Now Professor Yagi has shown us that one of the ways to accomplish the directive function is to use a string of absolutely automatic relay stations, viz., the simple devices he calls "directors." Not only in this ingenious suggestion but throughout a wide field of basic possibilities in directive radio,

[1] *Phys. Zeits.* **21**, 1, 1920.

† Chief of Radio Division, Bureau of Standards, Washington, D. C.

Professor Yagi has done exceptional fundamental work and has set forth a series of principles which will unquestionably guide much of the further development. While Professor Yagi's conclusions are validated by experiment, he has, as he says, in many directions only made a beginning and much remains to be done. I am sure that many of those who have heard and those who will read his paper will join him in further pursuit of a number of these interesting possibilities. When they have been fully developed we shall be a long way on the road toward the possibility of carrying on point-to-point communication by directive radio processes.

I have had the privilege of hearing Professor Yagi report not only on the part of his work included in his paper published in the PROCEEDINGS of the Institute of Radio Engineers, but also additional parts of it described in the *Proceedings* of the Third Pan-Pacific Science Congress. His work has included not only this development of wave projectors but also outstanding contributions to the technique of generating and using the shortest of radio waves, the development of the magnetron, and the amusing possibilities of radio power transmission. Whether the use of ultra-short radio waves will be important in long-distance communication, or whether Professor Yagi's ideas will have their principal application in methods of directing radio waves of more usual frequencies, time only can tell. In conclusion, I would like to say that I have never listened to a paper that I felt so sure was destined to be a classic.

# THE PIEZO-ELECTRIC RESONATOR AND ITS EQUIVALENT NETWORK*

By

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***Summary***—*The theory of the piezo-electric and the mechanical behavior of a quartz resonator is stated following Voight and Cady. The functions of the quartz as dielectric and as vibrator are shown to be separable and replaceable by a condenser in parallel with an electrical resonator, i.e., a series chain of inductance, resistance, and capacity. For a Curie-cut quartz rod excited lengthwise through the transverse piezo-electric effect into compressional vibration at the fundamental frequency the series elements become* $L = M/4\epsilon^2 b^2$; $R = N/4\epsilon^2 b^2$; $C = 4b^2\epsilon^2/g$. *This mode of vibration may be termed the "fundamental normal mode" since the vibration direction is normal to the electric field. For a Curie-cut quartz plate excited through the longitudinal piezo-electric effect into compressional vibration at the fundamental thickness frequency, the series elements become* $L = e^2 M/4\epsilon^2 l^2 b^2$; $R = e^2 N/4\epsilon^2 l^2 b^2$; $C = 4\epsilon^2 l^2 b^2/e^2 g$. *This mode of vibration may be termed the "fundamental parallel mode" since the vibration direction is parallel to the electric field.* $M$, $N$, *and* $g$ *are respectively the half-mass, mechanical resistance factor, and "equivalent stiffness" of the rod or plate whose thickness along the field is* $e$, *and dimensions normal to the field are* $l$ *and* $b$, *the latter being along the optic axis. The parallel capacity is shown to be less than that for a quartz dielectric which is free to assume piezo-electric strain and to be equal to that for unstrained quartz. Phase and amplitude variations of current to the resonator are shown as obtained with the cathode ray oscillograph.*

HALF a century ago the Curies discovered the piezo-electric effect. In the *direct effect* charges result from the compression of a crystal, which reverse in sign if the crystal is stretched instead of compressed. There are other types of stress under which similar charges appear and reverse in sign with the reversal of the stress. In the *converse effect*, which Lippmann showed must exist whenever a direct effect is thus reversible in sign, strains result when potential differences are applied to the crystal. These strains also reverse in sense with changes in the sign of the P.D.

The feebleness of these effects may be seen from the following data on a quartz plate, 1 mm. thick and 2.5 cm. square, the thickness lying along some optimum direction in the quartz for electrical effects from pressure. This direction is called *an electric axis* of the crystalline material, and *the electric axis* for a plate

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cut from the quartz as above stated. It was with plates cut in this fashion that the Curies measured the charges which result from pressure. A weight of 1 kilogram on the surface 2.5 cm. square should reduce the thickness by about two parts in ten million and set up in addition to this elastic strain a dielectric strain as well. This is the familiar Maxwell electric displacement due to polarization of the dielectric. With electrodes in contact with the square faces of the plate and connected to an external condenser of equal capacity the P.D. between the plates would be about 1 volt instead of 100.

The P.D. which results would be smaller as the condenser has larger capacity. In the converse effect, if a sufficient charge were applied to the electrodes to cause a P.D. of 100 volts the quartz plates would become thicker or thinner by two ten millionths of a millimeter, i.e., about one half of one thousandth of a light wave, if the plate were free to change its thickness. If not free, it would exert a force of a kilogram on its confines.

When a P.D. applied along the electric axis of a quartz plate tends to cause the quartz to contract along this axis, there is another direction at right angles to this electric axis along which the plate tends simultaneously to expand. Fractional changes in length along this second direction *transverse* to the applied field are equal to those which occur along (*longitudinal* to) the electric axis and are likewise reversible in direction with the sign of the field. It is with these transverse effects that Cady's[1] original development of the piezo-electric resonator was primarily concerned. Resonators cut from the quartz so as to utilize this transverse effect are used chiefly for the longer wavelengths. The two directions which have been referred to as *longitudinal* and *transverse* with respect to an electric axis determine a plane which is perpendicular to the well-known *optic axis* of quartz.

The illustrative examples given have been computed from formulas given by Voigt in his "Lehrbuch der Kristallphysik," formulas based on a theory which satisfactorily predicts the known facts of piezo-electrictification. In view of the minute effects described it is little wonder that piezo-electric phenomena were for so long of little interest except in connection with crystal theory. That so complete a theory of piezo-electricity as that of Voigt should have been developed in a field of so little apparent

[1] Proc. I.R.E. 10, 83, 1922.

practical usefulness is in itself a monument to theoretical science.

Given the phenomena of resonance in a piezo-electric crystal, the effects are no longer so small as to be unimportant. Since its introduction by Cady in 1921, interest in the *piezo-electric resonator* and its applications has developed rapidly until it is at the present day an essential element of most constant-frequency devices in the radio field.

The phenomena of resonance are fundamentally the same whether in a mechanical or an electrical system. A series electrical circuit is said to be in resonance with an applied alternating e.m.f. when the current is in phase with the e.m.f. The current may then be much larger for the same applied e.m.f. than at other frequencies and is given directly by Ohm's law. The natural frequency of free damped oscillations in the circuit is slightly lower than the resonance frequency by an amount depending on the decrement. When an alternating driving force has a frequency nearly the same as the natural free vibration frequency of the mechanical system being driven, then the resulting mechanical displacement may be many times larger than for a steady force of equal magnitude. Resonance strictly occurs for that driving frequency at which the mechanical vibrator so adjusts its response that its velocity is in phase with the driving force. The driving force is then applied directly against the resistance to the motion. For frequencies lower than this the velocity leads the driving force and the impedance which the mechanical vibrator offers to the force is said to have a negative angle. For frequencies above resonance the velocity lags behind the force and the mechanical impedance has a positive angle. The increase in amplitude at resonance is greater as the vibrating system has less damping.

In the quartz resonator the alternating mechanical forces which drive the mechanical resonator result, through the converse piezo-electric effect, from the alternating P.D. applied across its electrodes and are properly timed for resonance when the applied alternating potential has a frequency practically equal[2] to a natural free vibration frequency of the mechanical quartz resonator.

[2] Very slightly higher than the frequency of free damped vibrations. That this is so slight is due to the extremely low decrement of these resonators as compared with decrements met in electric circuits.

We shall see that the velocity of the vibrating crystal at resonance is practically in phase with the P.D. across its electrodes. This will suggest that the vibrating crystal responds like a tuned electric circuit with its current somehow masquerading as a mechanical velocity. Those who are familiar with mechanical impedance will note that the piezo-electric nature of quartz merely offers a coupling link between the mechanical impedance of the vibrator and current in the circuit in which the crystal is used.

Quartz resonators have such low damping that even when vibrating in air the increase in amplitude at resonance may be several thousand-fold. Cady found a 4000-fold increase for a typical plate even when damped by the presence of electrodes as close as possible without contact. If the 1 mm. thick quartz plate, which we have used in illustrating the feebleness of the static effect, should increase its amplitude a thousand times at resonance, an applied resonant P.D. of 100 volts across its electrodes would give it an amplitude of about half the wavelength of visible light. But to have attained a similar degree of compression from a steady applied force a weight of a ton would have been necessary, if the crystal did not break under the strain.

Cady in his original development of the theory of the resonator used the more or less familiar admittance circle diagram for studying the effects of the crystal at frequencies near resonance. The application of the resonance circle offers many advantages to those familiar with its use, as is true of most vector methods of solving alternating current problems. Some electrical engineers who are used to recognize circuit behavior from the arrangement of the elements in a circuit may understand the behavior of the resonator more easily if it can be replaced by a group of ordinary circuit elements. One purpose of the present paper is to develop a network which can, on paper at least, replace the crystal in all of its effects on the circuit in which it is placed. There is a certain fundamental reasonableness in the use of such an equivalent network since the resonator is used in electric circuits, where design problems involve relations between electric circuit elements only. Current flowing in a circuit can know only the effects of capacity, inductance, resistance, and electromotive force and their variations, and the current adjusts itself in frequency, magnitude, and phase to these elements.

For a resonator having a single degree of *mechanical* freedom the equivalent network is a simple one, in order that it may have a single degree of *electrical* freedom. The actual piezo-electric resonator under consideration is a rod or a plate of quartz which has a great many vibrating modes in which it may show resonance phenomena. Each of these modes may be considered by itself, and simpler mechanical resonators devised, each having a single degree of freedom, and each reproducing the mechanical vibration of the crystal in the vicinity of the resonant mode for which it was devised. The present paper will translate two of these simpler mechanical resonators into capacitance, inductance, and resistance in the crystal circuit. Corresponding to every other resonant frequency of the mechanical piece of quartz there is a similar network of constant elements which can completely represent the quartz in its effect on the circuit at all frequencies except in the neighborhood of other resonant frequencies of the crystal.

The two modes of crystal vibration with which the present paper deals are those lying in the plane perpendicular to the optic axis with end motions of the plate either at right angles to or along the electric axis of the plate and the applied field. Unfortunately there is some confusion between the terminology used by different writers arising from the use of "longitudinal," by some for compressional vibrations the long way of the plate, and by others for compressional vibrations which are set up in the thickness of the plate by the longitudinal piezo-electric effect. These latter call the vibration resulting from compressional strains of the transverse piezo-electric effect a "transverse" mode. Each of these modes of vibration represents the stationary wave pattern of a system of compressional waves (in mechanics and acoustics said to involve *longitudinal* vibrations), and neither vibration is concerned with waves which are transverse in the mechanical sense.

To avoid the confusion due to this double use of the term longitudinal to characterize both compressional vibrations and the piezo-electric effect which may cause the vibrations, the present paper will call modes where the vibration is parallel to the electric axis and the field *parallel modes* and will use the term *normal modes* where the vibration is perpendicular to the electric axis, i.e., at right angles to the field. The parallel modes are most directly driven by the longitudinal piezo-electric effect, and the normal modes by the transverse piezo-electric effect. The transverse and longitudinal piezo-effects may combine through elastic reactions in the quartz to drive the plate in a state of resonance for compressional waves along the third dimension of the plate, which is here the optic axis. Resonance frequencies for such vibrations are found by Hund.[3] To these *optic axis modes* the results of the analysis in the present paper are not directly applicable, though the method can be extended to apply to them.

*Fundamental modes* are those for which the wavelengths of these compressional waves are in the parallel case twice the thickness, and in the normal case twice the length, of the resonator. In either case the middle of the quartz is considered as a node with the ends in greatest motion. Higher overtones are not strictly in harmonic relation to these fundamental modes and have two or more nodal planes through the crystal. The networks to be derived are those for the parallel and normal fundamental modes of vibration of quartz plates or rods.

There are still other fundamental modes of vibration of quartz resonators, one of which is in common use. This Cady[4] has shown to be a shear mode. To excite it requires that the plate shall be cut from the original crystal with its face parallel to one of the six prismatic faces of the quartz instead of perpendicular to this face as for the plates with which this paper deals. In the plates with which the present paper is directly concerned, often described as "Curie Cut" plates, the electric field is along a natural electric axis of the quartz. Plates cut so as to use the shear are frequently referred to as "30 deg. crystals," as in their use the field makes an angle of 30 deg. with an electric axis. These latter are preferred by some for broadcasting station control. Their natural radio wavelength corresponds to about 140 meters per mm. thickness of the quartz as against about 110 meters per mm. with "Curie Cut" crystals. Though the analysis of the present paper is applicable only to "Curie Cut" crystals, a similar method may be applied to the shear vibrations of these "30 deg. crystals." Indeed it may be applied to any type of mechanical vibration which produces through known laws a change in an electric current, provided this change reverses in direction with the direction of the mechanical effect, and a network may be derived to have identical effects on the current.[5]

[3] Proc. I.R.E., 14, 459, 1926.
[4] *Phys. Review*, 29, 617, 1927.
[5] The general theory of such equivalent networks is given by Butterworth, *Proc.* Physical Society, 26, 264, 1914, and 27, 410, 1915.

The following familiar electrostatic theory[6] is assumed. Consider two parallel plates each of area $A$ and separation $e$ with an intervening medium of dielectric constant $k$. This dielectric may be said to have electric susceptibility $n$ such that $k=1+4\pi n$. This is analogous to the $\mu=1+4\pi k$ relating magnetic permeability to magnetic susceptibility. Opposite charges of surface density $\sigma$ exists on the inner surface of each plate. If the P.D. between the plates is $V$ the electric intensity, or strength of the field, between the plates is $V/e$. The electric displacement in the dielectric is $P=kV/4\pi e$ in units of Faraday tube per sq. cm., and is also equal to $\sigma$. The polarization $P$ is $nV/e$.

Voigt shows that a polarization $P$ may result in quartz either from an applied external mechanical pressure $p$ normal to proper faces of the plate, from mechanical shear in the quartz, and from an applied external field $V/e$. With shear and its associated polarization the present paper is not concerned. The polarization from pressure depends on the magnitude of the stray capacity of the closefitting electrodes and on any other capacity to which the electrodes are connected. Tubes of polarization in the quartz terminate on the inner surfaces of the electrodes resulting in charges of opposite sign on the two electrodes. These charges produce in the quartz dielectric between the electrodes their own polarization in a direction opposite to that from the pressure. The resultant of these two polarizations determines the force field between the charges on the inner surfaces of the electrodes and the force field determines the P.D. between the electrodes.

Furthermore a field applied from an external source produces a polarization in the quartz. A part of this polarization gives the usual dielectric displacement of a quartz condenser. An additional part results in a mechanical strain in the quartz, or else in a stress to balance some externally applied mechanical stress.

The properties of quartz which are responsible for these effects are given by the piezo-electric constant and the piezo-electric modulus, designated by $\epsilon$ and $\delta$ respectively. These should, strictly speaking, have subscripts to denote the directions to which they relate in the quartz. In this paper, the former, $\epsilon$, is the polarization along the electric axis resulting from unit mechanical compressional strains $s$ in the quartz. This associated strain may be either longitudinal or transverse to the electric

[6] All electric quantities used in this paper are expressed in c.g.s. electrostatic units unless specifically stated otherwise. Mechanical quantities are in c.g.s. units.

axis, but is always in the plane perpendicular to the optic axis. The latter, $\delta$, is the polarization along the electric axis associated with unit mechanical stress along either of these same two directions. The values of $\epsilon$ and $\delta$ are $-4.77\times10^4$ and $-6.45\times10^{-8}$ respectively in c.g.s. electrostatic units (Voigt's values). The quartz crystals is, of course, extremely complex in its elastic and its electrical behavior, showing different properties in different directions. Because of the very involved, and in fact unknown, exact nature of the strains involved in the vibrating rods and plates with which this paper is to deal, the rigorous equations which Voigt gives have been simplified to the extent of replacing algebraic sums of various component $\epsilon$'s and $\delta$'s by the single term most significant for our problem. For the complete expressions from which equations (1) to (4) are taken reference should be made to Voigt.

Voigt expresses the facts of piezo-electric polarization as they concern us here by the equations:[7]

$$P=(n+\epsilon\delta)V/e-\delta p=n'V/e-\delta p \tag{1}$$

$$\sigma=V(1+4\pi n')/4\pi e-\delta p=k'V/4\pi e-\delta p \tag{2}$$

$$Q=C'V-A\delta p=C'V-A\epsilon s \tag{3}$$

$$s=\delta V/e-up;\ p=\epsilon V/e \tag{4}$$

The two parts of the polarization are seen in (1). The first part involves the susceptibility $n'$ of a piezo-dielectric. As is seen in this equation the dielectric has susceptibility from two causes, $n$ which any ordinary dielectric would have, and $\epsilon\delta$ which is peculiar to its piezo-electric nature and accompanies its expansion or contraction due to the filed under constant end pressure. The effect of this latter is to increase slightly the apparent capacity of a condenser having a piezo-dielectric above that which the identical condenser would have if its dielectric were to be somehow robbed of its piezo-electric property, or if the crystal could be prevented from straining under polarization. This increase of capacity when extension at constant stress is permitted is of the order of one per cent and is probably of theoretical interest only; furthermore, ordinary measurement of the dielectric constant of quartz yields the *apparent* value $k'=1+4\pi n'=1+4\pi(n+\delta\epsilon)$. Hereafter the condenser functioning through this apparent dielectric constant will be referred

[7] Adapted from "Lehrbuch der Kristallphysik," 915–919.

to as a *piezo-condenser*, to distinguish it from the *piezo-vibrator* which is simultaneously involved in the same piece of quartz.

The second part of the polarization in the quartz is $\delta p$ of equation (1). This accompanies a compressional strain $s$ in the quartz, or, through the strain, its associated stress $p$. ($\delta p = \epsilon s$). The ratio $\epsilon/\delta = u$ is an elastic constant of the quartz and is roughly equivalent to Young's modulus in the plane perpendicular to the optic axis, in view of our approximations. Equations (2) and (3) express the density $\sigma$ of the charge on the inner surfaces of the electrodes, and the total charge $Q$, when the area of the electrodes is $A$. $C'$ is the measurable static capacity of the piezo-condenser when the crystal is under a constant stress. Equations (4) are fundamental equations for stress and strain in a piezo-electric substance, from which equations (1) to (3) are derived.

Our working formulas (3′) for the P.D. across a piezo-electric resonator come from a rearrangement of (3) above with allowance for additional capacity $C''$ which may exist in parallel with the resonator electrodes. $C''$ may perhaps represent a leakage capacity from the outer surfaces of the electrodes and the leads, or an extension of the electrodes beyond the area $A$ of the crystal face. Also a modified strain $s'$,[8] which is constant through the quartz, is used to avoid the actual strain distribution through the quartz.

$$V = \frac{(\sigma + \delta p)A}{C' + C''} = \frac{(\sigma + \epsilon s')A}{C' + C''} = \frac{Q + A\delta p}{C' + C''} = \frac{Q + A\epsilon s'}{C' + C''} \tag{3'}$$

The illustrative values already given to show the feebleness of the static effect were obtained by substitution in these formulas. The additional parallel capacity $C''$ may be considered to be included in $C'$.

When the mechanical quartz plate is vibrating in resonance with an applied P.D. it has been stated that strains may exist which are a thousand times as large as for the same static P.D. It is evident from (3′) that in static strains $\epsilon s'A$ now plays a part similar to that which charge plays in ordinary condensers. If, on the other hand, in considering vibratory phenomena, and thus, alternating strains, we regard $s'$ as the r.m.s. value of the strain, that part of the incident current which is responsible for resonant strains is $2\pi f_1 \epsilon s' A$ (effective or r.m.s. current) if $f_1$ is the frequency

[8] The modified strain $s'$ is explained later, and it is there shown that it may be used in place of $s$ in applying Eqs. (1–4) to the resonator.

at resonance. This amounts to a resonant current of 10 milliamperes or so to the $2.5 \times 2.5 \times 0.1$ cm. plate previously used in the static illustrations. It is here considered to be driven in resonance by a 100-volt 2800-kc. source.

The charging current to this same resonator due to its piezo-dielectric properties alone is an added 3 microamperes which flows even when the resonator is damped so that it cannot resonate. These two current values are computed from expressions $2\pi f_1 \epsilon s' A$ and $2\pi f_1 V C'$.

Thus it is seen that it is the actual straining of the quartz in its resonant mechanical vibrations that is responsible for most of the current to the resonator. This straining, and hence the current, assumes a given value for smaller applied P.D.'s at resonance than away from resonance, a fact which is also characteristic of currents in series resonant circuits. Current to the mounted crystal electrodes whether at resonance or not goes into two parallel channels, one channel that of the condenser made of piezo-dielectric and any parallel condensers, and the other channel that of the vibrating crystal. The relative magnitude of these at a given frequency and for given crystal area depends only upon the capacity of the condenser and the actual magnitude of the vibratory motions which the given crystal performs. This latter is limited by the mechanical impedance of the vibrator and by the driving P.D. This statement is analogous to the ordinary law for alternating currents in a divided circuit. For the divided circuit network, which is to be derived as equivalent to the piezo-electric resonator so far as circuit phenomena are concerned, a corresponding statement is: At any frequency, whether at resonance of one of the branches or not, the alternating current $I$ divides into two channels with relative magnitudes which depend only on the impedances of the two branches.

Caution is needed in applying equations (3′) to the vibrating crystal. These equations hold for instantaneous values of the P.D., charge, pressure, and strain. With sinusoidal variation of these quantities they must be used in proper phase relation, e.i., they must be treated as vector quantities. These same equations are true if vector quantities are used. The problem is simplified by using the time derivatives of this equation, made explicit in current, Eq. (6). This total current to the resonator and its associated capacities will then be seen to be the vector sum of that to the piezo-condenser and that to the piezo-vibrator. This vector sum, or a corresponding algebraic sum for instantaneous values of current, corresponds to Kirchoff's law for current at a junction. Thus the current to the vibrating crystal and its associated condensers behaves as if it flowed to a condenser and some other circuit in parallel. The problem is to find the nature of this parallel branch.

Before proceeding to the derivation of the network it is necessary to recall Cady's treatment of the theory of longitudinal vibrations of viscous rods. In this, as has been suggested, he shows that when lateral effects are neglected and compressional waves are travelling along the rod driving it at a frequency near the fundamental frequency, the analysis of the vibration may be simplified by substituting for the actual rod an equivalent vibrating system possessing a single degree of freedom. This equivalent system is so designed as to have an amplitude of vibration equal to that of one end of the actual rod (whose middle is fixed) when the driving force for the mechanical substitute system is equal to twice the force on one end (the sum of that on two ends) of the actual vibrating bar. His equation of motion is:

$$F_0 \cos \omega t = M\ddot{x} + N\dot{x} + gx \qquad (5)$$

Here $F_0$ is the maximum amplitude of the sinusoidal force applied to the equivalent system. $\omega$ is $2\pi$ times the frequency of the applied force; $M$ is half the mass of the rod; $N$ is the mechanical resistance coefficient, which is equal to $\pi^2\rho A'Q/2l'$, and $Q$ is the viscosity of the resonator material including any viscous effects of the atmosphere and the mounting in which it vibrates. $\rho$, $A'$, and $l'$ are respectively density, vibrating face area, and length of the rod. $g$ is the *equivalent stiffness* for the substitute system and is related to Young's modulus $G$ for the actual rod through the equation $g = \pi^2 A'G/2$. $x$ is the instantaneous displacement of one end of the rod. Dotted letters are used for first and second time derivatives. Cady also shows that the fictitious instantaneous pressure which may be supposed to act on each end of the vibrating bar to drive it in resonance is the same as the uniform real stress acting throughout the bar.

Most of the analysis which has thus far been presented is an adaptation from Voigt[9] or from Cady.[10] Cady proceeds to find the current to the resonator and its mounting both analytically

[9] Voigt, "Kristallphysik."
[10] Cady, *Phys. Rev.* 19, 1, 1919; PROC. I.R.E., 10, 83, 1922.

and graphically with the aid of the resonance circle. We shall, however, further interpret our equation (3′) for potential difference across the electrodes of the resonator. Differentiating (3) and solving for $I$ gives equation (6) the current to the resonator.

$$I = \dot{V}C' - \epsilon A\dot{s}' \tag{6}$$

In the mechanical effects with which this paper is concerned, those of rods vibrating near resonance, the strain $s$ obviously varies from point to point in the rod at any instant. A stress, which may be thought of as causing this strain at each point, varies similarly from point to point. Since this stress is $p$ of Eq. (1), that component of the polarization $P$ which is associated with this stress obviously varies according to the same law along the rod as does the strain. The corresponding component (by Eq. (3)) of the total charge on the electrodes is $2l'$ times the integral of this polarization over half the length of the plate. Since the instantaneous extension of the rod is the same function of strain through the half length of the rod, we can replace the point strain in Eq. (3) by a modified strain $s'$ which is uniform along the rod, but varying with time and which is defined by $s' = 2x/l'$. Here $x$ is the actual mechanical displacement of the end of the rod and $s'$ simply the average strain through the rod. In effect it replaces the strain of the actual rod by that for a mechanical resonator of a single degree of freedom.

This "average strain" is made up of two components at each instant in time. One part is that which accompanies the $\epsilon\delta V/e$ part of the polarization in Eq. (1) by virtue of the piezo-dielectric properties of quartz as distinct from ordinary dielectric properties, and is $\delta V/e$ by application of equation (4). It is obviously in phase with $V$ and cannot take part in the phase variations of the piezo-vibrator in the vicinity of resonance. The ability to vary in phase from the applied P.D. and thus to react on the rest of the circuit is restricted to the remaining part of the strain, namely, $s' - \delta V/e$. This other component of the strain when multiplied by $l'/2$ the distance from node to antinode in the vibrator, gives our $x$ the instantaneous displacement of one end of the vibrator. It should be noted that this is not quite as large as the actual displacement of the end of the vibrating bar for $x$ is to be measured from the non-reacting[11] displacement $l'\delta V/2e$. The latter is the only displacement when the driving

[11] Cady, W. G., Proc. I.R.E., 10, 83, 1922. This is here called "equilibrium elongation."

frequency is remote from resonance or when the plate is highly damped.

The instantaneous velocity of the piezo-vibrator face is $\dot{x}=0.5l'(\dot{s}'-\delta V/e)$ which when solved for $\dot{s}'$ and substituted in (6) gives:

$$I=\dot{V}C'-A\epsilon\delta\dot{V}/e-2\epsilon A\dot{x}/l'=\dot{V}(C'-A\epsilon\delta/e)$$
$$-2\epsilon A\dot{x}/l'=\dot{V}C_1-2\epsilon A\dot{x}/l' \qquad (7)$$

$A\epsilon\delta/e$ thus appears as a small correction to the capacity of the piezo-condenser (about 1 per cent). Considering the two terms of $C_1$ it is seen by $C'=k'A/4\pi e$ that the effect of $A\epsilon\delta/e$ is to reduce[12] the $k'/4\pi$ by the amount $\delta\epsilon$ which means reducing $n'$ by $\delta\epsilon$, i.e., back to $k$ and $n$ the dielectric constant and suscepti-

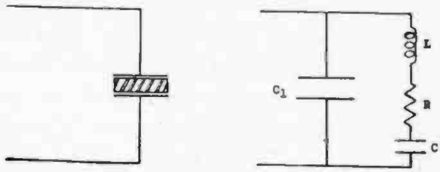


Fig. 1—Piezo-Electric Resonator and Its Equivalent Network

bility of unstrained quartz. We shall then, hereafter, consider the parallel condenser no longer as a piezo-condenser, but as having the capacity of an ordinary dielectric, i.e., as if the quartz dielectric were not piezo-electric, and call it $C_1$ instead of $C'$.

Eq. (7) is a statement of Kirchoff's current law, and may be considered either as a vector equation or an equation in instantaneous values. The current to the piezo-vibrator is proportional to and in phase with the rate of change of its modified resonating strain or to the velocity of the vibrating face of the crystal. From the analysis of mechanical resonators of a single degree of freedom[13] the instantaneous velocity $\dot{x}$ of the resonator is expressible in terms of the sinusoidal driving force of maximum value $F_0$ and the vector mechanical impedance $Z$ of the resonator by Eq. (8) which is of the same form as the current equation in a series electric circuit.

[12] Dye, p. 453 of paper to be later cited, finds experimentally such a reduction in the capacity of the parallel condenser, and points out that his observed decrease is more than here accounted for. The present decrease should enter every a-c. determination of this capacity whether near crystal resonance or not.

[13] Cf. Crandall, "Theory of Vibrating Systems and Sound," p. 9.

$$\dot{x}=\frac{F_0 \cos \omega t}{Z}=\frac{F_0 \cos \omega t}{N+j(\omega M-g/\omega)} \tag{8}$$[14]

Substitution of (8) in (7) gives the Eq. (9) for current to the crystal.

$$I=I_1+I_2=C_1\dot{V}-\frac{2\epsilon A F_0}{l'Z}\cos \omega t \tag{9}$$

When the equivalent driving force on the vibrating face (area $A'$) of the crystal is expressed in terms of the maximum applied field $V_0/e$ by Eq. (4), $F_0=2A'p_0=2A'V_0\epsilon/e$ Eq. (9) takes the form

$$I=I_1+I_2=C_1\dot{V}-\frac{V_0 \cos \omega t}{\dfrac{el'N}{4\epsilon^2 AA'}+j\left(\dfrac{Mel'\omega}{4\epsilon^2 AA'}-\dfrac{el'g}{4\epsilon^2 AA'\omega}\right)} \tag{10}$$

$I_2$ is the current to the vibrating crystal or to whatever group of circuit elements may be imagined to take its place. The

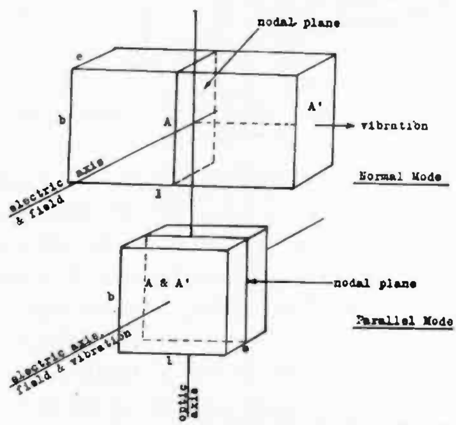


Fig. 2—Crystal Dimensions in Normal and Parallel Fundamental Modes

denominator is, by definition of electrical impedance, (ratio of P.D. to current) the electrical impedance of this imagined branch. This denominator has the form of the impedance of a series chain of electrical elements—inductance $L$, resistance $R$, and capacitance $C$—for which the expression is $R+j(\omega L-1/\omega C)$. Hence the elements are

$$R=\frac{el'N}{4\epsilon^2 AA'};\ L=\frac{el'M}{4\epsilon^2 AA'};\ C=\frac{4\epsilon^2 AA'}{el'g};C_1=\frac{A}{e}\left(\frac{k'}{4\pi}-\epsilon\delta\right)=\frac{Ak}{4\pi e} \tag{11}$$

[14] This is also the derivative of Cady's equation (20). *Phys. Rev.*, **19**, 6, 1922.

Thus the entire piezo-electric resonator may be replaced by the equivalent network of Fig. 1 where the condenser $C_1$ represents the quartz dielectric and the series chain represents the piezo-vibrator.

The values of the elements of the series chain depend upon the dimensions of the crystal, upon its characteristics as a mechanical vibrator, and upon the value of the piezo-electric constant of the quartz. The factor $el'/4\epsilon^2AA'$ is a sort of conversion factor which converts impedance when measured in mechanical units for the vibrator into impedance in electrical units for the current. The $R$, $L$, and $1/C$ would all have smaller values and thus have more effect on the current in the external circuit if a piezo-electric material were used which was more strongly piezo-electric, i.e., had a larger piezo-electric constant.

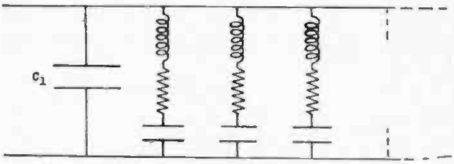


Fig. 3—Single Network Equivalent to Resonator at a Number of Its Natural Frequencies.

The expressions (11) may be simplified for either mode of vibration by substituting for area $A'$ of the moving face, the crystal dimension $l'$ in the direction of the vibration (i.e., the half-wavelength in the crystal) and the area $A$ of the faces which the electrodes cover, their respective values in terms of the length $l$, breadth $b$, and thickness $e$ of the crystal. These are measured in directions which are respectively normal to the electric and optic axes, along the optic axis, and along the electric axis. For the *normal mode* the series elements become[15]

$$R=\frac{N}{4\epsilon^2b^2};\ L=\frac{M}{4\epsilon^2b^2};\ C=\frac{4\epsilon^2b^2}{g} \tag{12}$$

[15] The author outlined the derivation of this equivalent network before the American Physical Society in Washington in April, 1925, the abstract appearing in *Phys. Rev.*, **25**, 895, 1925. Since then similar equivalent networks have been published. That of Dye (*Proc.* Phys. Soc. of London, **38**, 399, 1926) was derived in form from a general dynamical theorem published by Butterworth, loc. cit. and the values of the elements obtained from resonance curves taken for the resonator. When deriving the network the present author was aware of Butterworth's general theorem but preferred to base his derivation on first principles, with fundamental piezo-electric theory as the starting point. The network was first derived from equation (18) of Cady's I.R.E. paper by a few simple steps,

and for the parallel mode

$$R=\frac{e^2N}{4\epsilon^2A^2}=\frac{e^2N}{4\epsilon^2l^2b^2}; \quad L=\frac{e^2M}{4\epsilon^2A^2}=\frac{e^2M}{4\epsilon^2l^2b^2}; \quad C=\frac{4\epsilon^2A^2}{e^2g}=\frac{4\epsilon^2l^2b^2}{e^2g} \tag{13}$$

The quartz-condenser $C_1$ has the same value for either mode, for plates of the same area and thickness.

A single network may be drawn (Fig. 3) to represent the crystal in all of its various modes of vibration provided the frequency of each mode is remote enough from other resonance frequencies to insure that the other series chains have very high impedance —a frequency separation of but a tenth of a per cent or so with quartz resonators.

Illustrative values for the elements of the network will now be given. The plate $N_2$ which was described by Cady[16] and used in his early experiments, is a rod about 30×4×1.5 mm. vibrating at its *normal fundamental* (lengthwise vibration) with a resonant frequency of about 90 kc. per second. Substituting in the formulas (12) and converting from e.s.u. to practical units, the following values are obtained:

$L=137$ henrys $\qquad C=0.0235\ \mu\mu f.$
$R=$ about 15,000 ohms $\qquad C_1=3.54\ \mu\mu f.$

For a plate of a size commonly used in oscillators (25×25×2.5 mm.) vibrating at its *parallel fundamental* (thickness vibration) near to a resonant frequency of about 1100 kc. per second, on substitution in Eqs. (7) the following values are obtained:

$L=0.33$ henrys $\qquad C=0.065\ \mu\mu f.$
$R=$ about 5,500 ohms $\qquad C_1=1.0\ \mu\mu f.$

The numerical values for $R$ in the above networks were computed not from the known viscosity characteristics of quartz, but rather from an assumed apparent viscosity ($Q=100$) which is intended to include in a rough way the effects of air damping and mounting friction. If the crystal were mounted in a vacuum and free from mounting friction the viscosity would probably be

—a method which avoids the "modified strain" used in the present development.

Other published papers in which an equivalent network has been shown include:

Balth. van der Pol.—"Gendenboch Uedenlausche Vereeniging voor Radiotelegrafic" (1926, p. 293–8). Abstract appeared in *Jahrb. der. drahtl. tel.*, **28**, 194, 1926.

F. Bedeau, *QST* Francais, **8**, p. 22, 1927.

Y. Watanabe.—*Elektr. Nachr. Techn.* **5**, 45, 1928.

Y. Watanabe.—*Jr.* I.E.E. of Japan, No. 466, 506, 1927.

[16] Proc. I.R.E., loc. cit.

less than half of this value, and the values of $R$ thus only half as large. The value $Q=100$ is that which corresponds to the decrement which Cady obtained for the resonator $N_2$ vibrating in air with electrodes as close as possible to the crystal.

The assumption of this same value of the viscosity from which to compute $R$ for the longitudinal mode is even less justified because of the entirely different shape of the crystal. Also Cady's theory of longitudinal vibrations of viscous rods is not rigorously applicable to the "thickness" vibrations of a plate, since in the rod the cross-section is supposedly small compared to the length. However, for the fundamental compressional mode the procedure should be valid as a first approximation when the proper elastic modulus is used.

Our application of piezo-electric theory in the above development has involved the assumption that the electrodes were in contact with the quartz. This is usually not the case in the practical application of the piezo-electric resonator. Even when one supposes that the electrodes are in contact, the vibrations of the crystal push them periodically away so that there is always at least a small air gap and perhaps a varying one. The complications which this air gap introduces are treated experimentally by Dye who represents the capacity of the gap by a small capacity in series with a network of the type here derived.

The reactions caused by stationary compressional waves in the air gap should be represented, as Dye[17] points out, by a secondary resonant circuit to be equivalent to their system of vibration as coupled to the piezo-vibrator.

The derived values for the network hold rigorously only in the vicinity of the mode of resonance for which they were derived. This limitation is inherent, as Cady makes clear, in the substitution of the equivalent damped mechanical resonator for the actual viscous rod. But as the series chain has a large positive reactance above resonance, and a large negative reactance below resonance, the network reduces, so far as practical considerations go, to the capacity $C_1$ except in the vicinity of resonance.

The magnitude of the elements of the network are at first sight rather striking. When the radio engineer considers an inductance coil of 100 henrys or more and tries to picture such a coil having so little distributed capacity as to remain inductive at 90 kc., serious constructional difficulties arise in his mind. To

[1] Loc. cit. p. 428.

tune such a coil to resonance at 90 kc. with a condenser having a capacity of but 1/40 of a $\mu\mu f$. requires that the coil shall have ridiculously small distributed capacity. Fortunately, we are not concerned with the construction of such a coil, for we already have in effect the coil and the condenser tuned to the desired frequency in the properly cut quartz crystal. Furthermore, the coil and condenser whose places the quartz crystal takes are put into so small a space in quartz that we have little worry from stray inductive and capacitive fields. Though these values of $L$, $R$, and $C$ seem absurd from the constructional standpoint, nevertheless, to the engineer who wishes values of electrical elements to use on paper in design problems, their magnitude presents no such difficulties.

Marked resonance phenomena seem oddly associated with resistances of the order of 15,000 ohms. It is to be recalled, how-

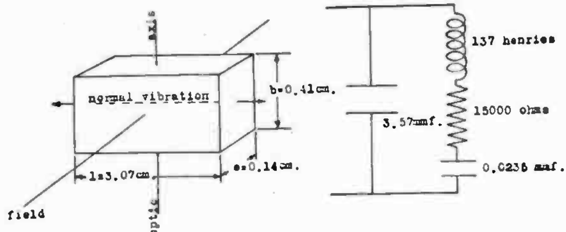


Fig. 4—Resonator $N_2$ and Its Equivalent Network for Normal Fundamental Mode

ever, that it is the ratio $L/R$ which determines the time constant of a coil and the decrement of a circuit. Here, though $R$ is large, $L$ is even larger in proportion than with most tuned circuits, and the decrement is thus extraordinarily small.

The sharpness of the resonance obtained with the piezo-vibrator may be seen from the ratio of the reactance of $L$ at resonance to $R$. Consider the plate $N_2$: 137 henrys at 90 kc. has a reactance of nearly 80 megohms, while the resistance is but about 15,000 ohms; their ratio is about 5,000.[18] At a frequency 1 per cent different from the resonance frequency the reactances of $C$ and $L$ which were equal and opposite at resonance, now

[18] In this connection the "mechanical resonator" described by Moore and Curtis, Bell System Technical *Journal*, 6, 222, 1927, shows a reactance-resistance ratio ten times as large as in the present case. Their resonator was a carefully mounted steel rod. From more recent estimates of the decrement of quartz resonators the reactance-resistance ratio should have a value several times the value of 5000 used.

differ by one hundred times the value of $R$. Even at 1/10 per cent from resonance their difference is nearly 10 times $R$. To shift the angle of the impedance of the piezo-vibrator by 45 deg. it it necessary to vary the driving frequency but about 1/100 of one per cent from resonance,—about 9 cycles at 90 kc.

When the frequency is varied about 0.3 per cent above resonance, or less than 300 cycles, the positive reactance of the piezo-resonator is equal in magnitude to the negative reactance of the quartz condenser. At this frequency the mounted resonator has the resistive impedance which is characteristic of parallel resonance (anti-resonance) and is greater than the impedance of either branch. If we call this frequency of parallel resonance $f_3$ and the frequency of series resonance of the piezo-vibrator $f_1$, there is still another frequency of resonance in between these.

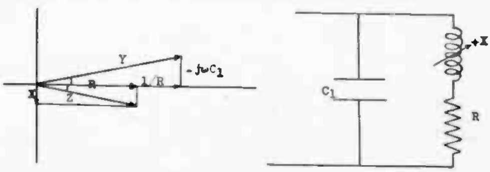


Fig. 5—Impedance and Admittance at $f_2$

The parallel network, a resistance and a capacity in parallel at $f_1$, offers there an impedance with a slight negative angle. At a frequency slightly below $f_3$, however, the positive angled impedance of the $L$, $C$, $R$ chain has a lower value than the condenser $C_1$ by definition of $f_3$. Hence the parallel network has an impedance whose angle is positive near $f_3$, and therefore must have changed sign between $f_1$ and $f_3$,—say at $f_2$ a second frequency where the device acts as a pure resistance. Though this frequency may be determined by setting up the equations for the admittance of the parallel network and solving for the condition that its angle shall be zero, it is more easily visualized by vector methods.[19]

The frequency $f_2$ in question may be shown to be extremely close to $f_1$, for the network of the crystal $N_2$ within about 1/30 of one cycle. $f_2$ may be seen from the vector diagram (Fig. 5) to be that frequency for which the admittance $Y$ for the piezo-

[19] The vector admittance circle as applied by Cady is, of course, best suited to the analysis of the resonator behavior, as it is also for the analysis of the network. Cady's circle is applicable to the network on substitution of circuit elements for crystal elements.

vibrator has such an angle as to result with the admittance $-j\omega C_1$ of the parallel condenser in a combined admittance of zero angle. (For $N_2$ at 90 kc., $1/\omega C_1$ is about 1/2 a megohm.) The angle of $Y$ is equal, but of opposite sign, to that of the impedance $Z$ of the piezo-vibrator and has a value $X/R$. (Here $X$ is

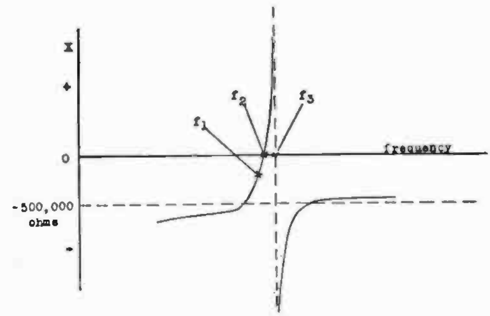


Fig. 6—Reactance of Resonator $N_2$

the reactance of $LCR$.) If the angle is small, $Z$ is practically equal in magnitude to $R$ and hence its reciprocal $Y$ to $1/R$. Thus in the present case where $R$ = about 15,000 ohms, $Y = 1/15000$ and the angle of $Y$ (or of $Z$) is (by $\omega C_1/Y$) 0.3 (as $1/\omega c_1 = 500{,}000$ ohms) from which by equal angles $X/R = 0.03$. From the resistance and reactance data given above it may be seen that this ratio of $X/R$ occurs at a frequency only 3/10,000 of one per cent

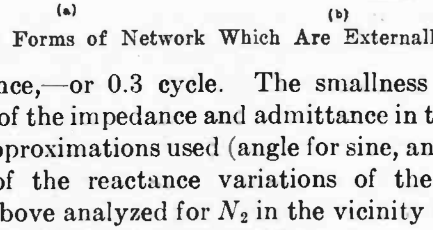


Fig. 7—Two Forms of Network Which Are Externally Equivalent

above resonance,—or 0.3 cycle. The smallness of the angles (1/30 radian) of the impedance and admittance in this illustration justifies the approximations used (angle for sine, and for tangent). The nature of the reactance variations of the piezo-electric resonator as above analyzed for $N_2$ in the vicinity of resonance is shown qualitatively by the curve of Fig. 6.

Johnson and Shea[20] have demonstrated the rigorous equivalence between networks of the two forms shown in Fig. 7, pro-

[20] Bell System Technical *Jr.*, 4, 60, 1925.

vided the ratio of the resistance to reactance is the same for both coils, and the ratio of resistance to reactance is the same for all condensers. Hence the present network equivalent to the resonator may be converted to the form (*a*) Fig. 7. For the conversion formulas the reader is referred to the paper cited. The equivalence of these two forms is of particular interest in connection with any allowance which should be made for the gap effect as separate from the reaction of the resonant air column. The present network has assumed the electrodes to be in contact with the crystal. Dye[21] shows experimentally that the effect of the gap may be represented by a capacity (which he calls $K_2$) in series with a network (our Fig. 1). If such a condenser can be placed in series with the network (*b*) Fig. 7 to include the gap, it can also be placed in series with (*a*) Fig. 7 and can then be allowed for by a revision of the value of $C_2$. From this revised network (*b*) Fig. 7 a new network of type (*a*) Fig. 7 can be drawn, but with values of the constants which are now corrected for the gap. Thus it would seem to be unnecessary in picturing the circuit behavior of the resonator to introduce a series condenser for the gap effect. The air column effect, however, as was pointed out above must be considered to involve a separate tuned circuit coupled in some way to the present series *LCR* branch.

The author has used the cathode-ray oscillograph to demonstrate experimentally the nature of the variations in the crystal impedance as the frequency is varied through resonance. Using as the oscillograph the Western Electric[22] vacuum-tube No. 224-A, one pair of deflecting plates of the oscillograph is arranged to show the P. D. across the resonator while the other pair of plates shows the current to the resonator. Since the oscillograph has a high impedance, resistances are placed across each deflecting system. Thus the deflections and their phases are not characteristic of the mounted crystal alone but depend also upon the magnitude of the resistances or condensers used. Bridge arrangements of the types shown as (*a*) and (*b*) (Fig. 8) have been used. Fig. 9 shows a typical set of ellipses sketched for the series of frequencies shown in the vicinity of resonance (normal mode) about 90 kc. for a plate ($30\times10\times1$ mm.). The elements of the bridge were all resistances; those across the deflecting plates being

[21] Dye, loc. cit.

[22] The circuit used is analogous to that described for studying dielectric losses in the "Instructions for Operating the No. 224-A Vacuum-Tube."

25,000 ohms each, and the third element 100,000 ohms. As the crystal frequency is approached from the low-frequency side, the ellipse shifts through the phases shown, all in a narrow frequency range in the vicinity of resonance. For the figures shown the deflections up and down the paper are proportional to currents through the crystal, and those across the page to P.D. across crystal and resistance in parallel.[23]

One difficulty in following such phase changes continuously through resonance in early experiments was the reaction of the crystal back on the source,—its stabilizing action. The sudden changes in phase of the load react in such a way on the source as

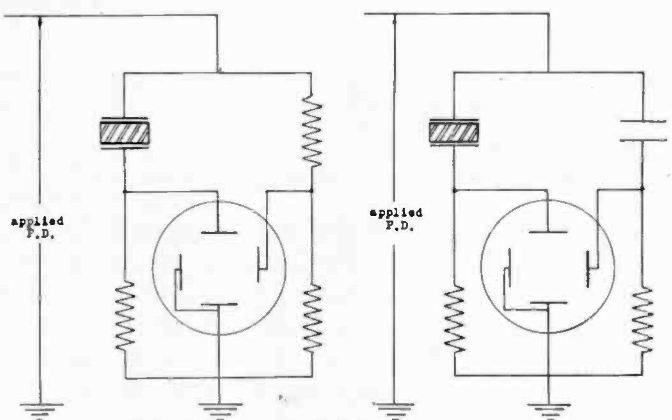


Fig. 8—Piezo-Electric Resonator in Oscillograph Bridge

to prevent continuous tuning through the resonance frequency of the load unless the crystal forms a very insignificant part of that load. In the present experiments the driving oscillator is of such power and so loosely coupled to the crystal and oscillograph bridge that no reactions of this sort are detected. As a result the ellipses vary continuously through the sorts of changes indicated as the frequency is slowly varied through resonance.

A series of figures of ellipses of this type has been made for various sorts of quartz resonators varying in frequency from 20 kc. to 1000 kc. It is hoped that careful measurements on these ellipses taken at carefully determined frequencies in the vicinity of resonance can be solved to yield reliable estimates of the magnitude of the elements which make up the equivalent network of

[23] Watanabe, *Jr.* I.E.E. Japan, May 1927, shows similar oscillograph figures for the piezo-resonator.

the resonator, though no estimate of the precision to be expected has yet been made.

The use of the cathode-ray oscillograph is found of further value in distinguishing between true and false response frequencies of the crystal. When a crystal is placed across a vacuum-tube oscillator and the oscillator is tuned continuously through a range of frequencies, crystal reactions result in clicks heard in phones connected to the oscillator or in a secondary circuit. Such clicks are heard not only for oscillator frequencies which match crystal vibration frequencies, but occur as well for those oscillator frequencies which have harmonics of appreciable magnitude at a crystal frequency. The crystal is thus driven by

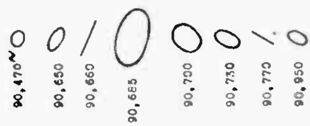


Fig. 9—Oscillograph Patterns Near Resonance

the harmonic, and a click resulting from its reaction on the oscillator circuit may be heard. These clicks are usually not so loud in this latter case as when caused by the fundamental frequency of the oscillator, and a basis for their recognition is had in this difference in loudness. On the other hand there are marked differences in the loudness of clicks for different true crystal resonance frequencies. With the cathode-ray oscillograph, however, when the crystal is responding to the driving oscillator, the ellipses shown above appear if the response is to the fundamental of the oscillator, but if to some over-tone in the oscillator output, the patterns show a two-ring, or a three-ring structure instead of the simple ellipse. This method has been used to locate the resonance vibration frequencies of resonators of various shapes. It has the added advantage of permitting one to explore carefully the region in the immediate vicinity of a resonance frequency. Thus it has been found in some cases that a crystal has a pair of resonance frequencies so close together that by the click method they have appeared as one.

The author cannot close this paper without acknowledging Professor Cady's patient counsel and helpful assistance in its preparation.

# SOME CORRELATIONS OF RADIO RECEPTION WITH ATMOSPHERIC TEMPERATURE AND PRESSURE*

By

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***Summary***—*Night reception and temperature at the receiver are found directly related, maximum reception being associated with maximum temperatures and vice versa. This is the reverse of the relation previously found by Austin for day reception, where falling temperature improved reception, and is therefore another case of the already established inverse relation of night to day reception. The temperature effect appears to be local to the receiver, for no definite relation was found between temperature at the transmitter and reception. A correlation between night reception and pressure was also found, signal strength increasing as areas of low pressure passed over the receiver, and decreasing with the passages of high pressures.*

THE first definite correlation between radio reception and any measured atmospheric state was made by Austin in 1924[1]. Comparing air temperatures with day reception at Washington from low-frequency stations in New Jersey he found an inverse relation, which was particularly marked with cold waves. Early in my measurements of night reception in the broadcast band I examined pressure gradients between transmitter and receiver without finding a definite relation to reception.[2] But when a homogeneous series of reception measurements extending over a period of two years had accumulated, meteorological relations began to appear, particularly those with temperature and pressure which form the subject matter of this paper.

Two series of reception measurements were available for this comparison; my own of station WBBM at Chicago as received in Newton Centre, Massachusetts, and those of Mr. Howell C. Brown at Pasadena, California, of transmission from station KWFI at San Francisco. Mr. Brown's measurements are of peculiar value in this work, as his transmission path (perhaps because it has a considerable component parallel to the Earth's magnetic field) appears but little affected by solar activity and magnetic disturbances. Unfortunately, although my measure-

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* Presented before the International Union of Scientific Radiotelegraphy, Washington, D. C., April 19, 1928.

[1] Proc. I.R.E., 12, 681, December, 1924; 14, 781, December, 1926.

[2] Proc. I.R.E., 15, 95, February, 1927.

ments over a West-East path now cover a period of over two years, the California series began in July, 1927, and is as yet too short for a really satisfactory comparison.

In order to eliminate the effect of seasonal and other long period changes in reception the individual night fields were divided by their moving 27-day average, thus giving a series of index numbers representing the percentage deviation of each night from the 27-day average centering on that night. Selecting a sufficient number—about thirty—of temperature or pressure changes as epoch dates, a mean of the reception index numbers was made for the day of the change and also for each of a number of days before and after.

Daily mean temperatures taken by the Weather Bureau offices nearest the transmitter and receiver are used throughout.

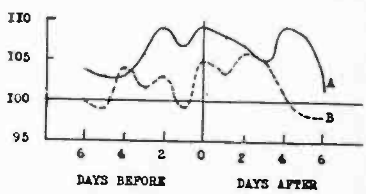


Fig. 1—Newton Reception from Chicago Against Chicago Temperature Changes.

For the Chicago-Newton path, the temperatures are those of Chicago and Boston, while for the San Francisco-Pasadena transmission, those of San Francisco and Los Angeles are employed. To eliminate the effect of small and purely local temperature changes, only changes of 10 deg. F. or over were taken for the Chicago-Newton transmission, and because of the disturbing effect of the East wind on Boston summer temperatures only the fall, winter, and spring months were used. For the San Francisco-Pasadena path, the more uniform California climate made it necessary to take smaller temperature changes, those of 5 deg. F or over being used.

In Fig. 1 is shown a comparison of Newton reception with temperature changes at the Chicago transmitter over a period of thirteen days centering on the days of the temperature change, the ordinates being reception percentages. The full-line curve *A* is of reception accompanying falling temperatures, while the dotted curve *B* is for rising temperatures. No definite relation appears, and the amplitude is small, about five per cent in each case.

But when daily mean temperatures at Boston, 11 kilometers from the receiver, are taken, a more definite relation is found. In Fig. 2 the full-line curve $C$ is for cold waves, while the dotted

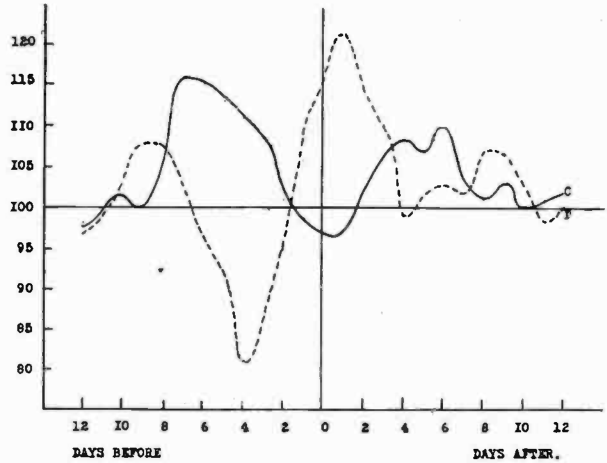


Fig. 2—Newton Reception from Chicago Against Boston Temperature Changes.

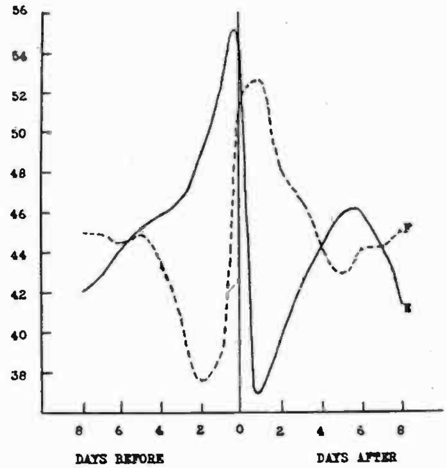


Fig. 3—Character of Boston Temperature Changes.

curve $D$ is for abrupt temperature rise. Considering for the moment only the central part of the figure it will be seen that whereas reception shows a distinct minimum centering on the day of the cold wave, it also shows a marked maximum nearly coinciding with the day of the temperature rise.

To interpret Fig. 2 fully it is first necessary to determine the character of the temperature changes for several days before and after the abrupt rise or fall. In Fig. 3 is shown the result of

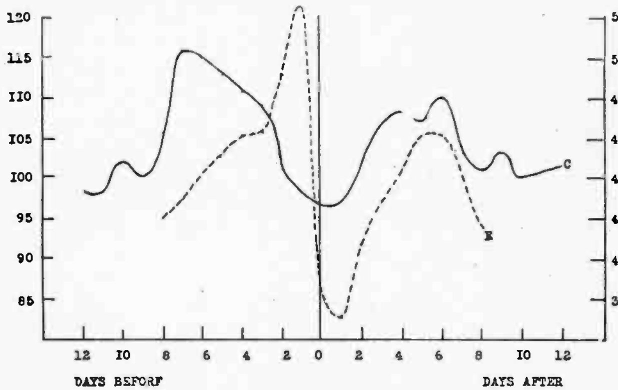


Fig. 4—Comparison of Newton Reception and Boston Temperature. Changes.

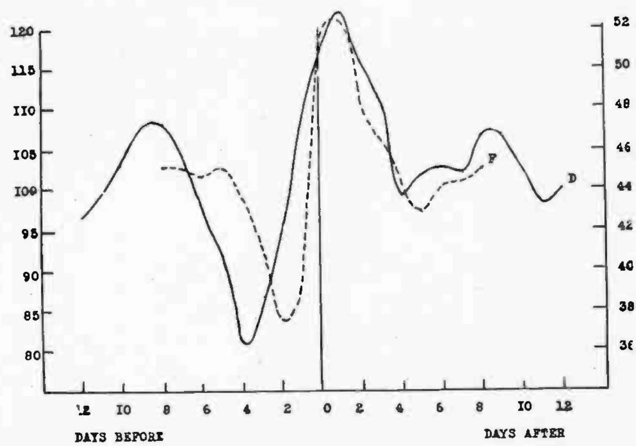


Fig. 5—Comparison of Newton Reception and Boston Temperatures. Temperature Rise.

averaging daily mean temperatures over a period of seventeen days centering on the day of the change, full-line curve *E* being for cold waves, while dotted curve *F* is for sudden rises. The curves are very nearly inverse, showing in each case a slow rise or fall, a sudden fall or rise on the day of the change, and then a slow recovery to normal.

We will first compare reception and temperature before, during, and after a cold wave. In Fig. 4 this is done by superposing reception and temperature curves $C$ and $E$, the ordinate scale at

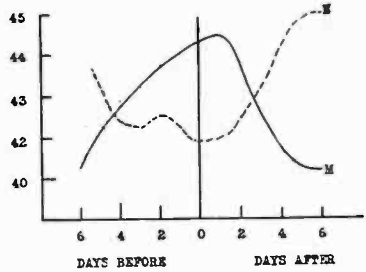


Fig. 6—Boston Temperatures Accompanying Newton Reception Maxima and Minima.

the left being reception percentage, while that at the right is temperature. Clearly temperature and reception are here directly related, although the curves are somewhat displaced, reception changes coming in advance of temperature.

An even clearer relation is shown for temperature rise and reception in Fig. 5. Here curve $D$ is for reception, while curve $F$

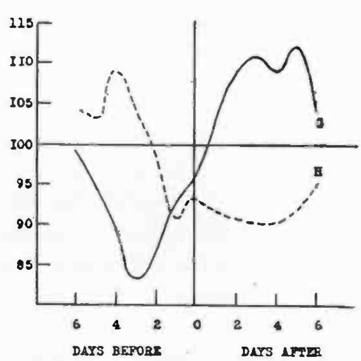


Fig. 7—Newton Reception Against Passage of Cyclones and Anticyclones over Massachusetts.

is for temperature. And in the left-hand portion of this figure the same tendency of reception to lead temperature is shown, which suggests that reception may not be directly related to temperature, but perhaps instead to a temperature-controlling cause.

As a check upon the foregoing, I have reversed the process so far employed, by taking the maxima and minima of my night reception series as epoch dates, and finding the associated Boston temperatures. The result is shown in Fig. 6, where full-line curve *M* represents the temperatures centering on reception maxima, and dotted curve *N* those accompanying reception minima. Again maximum temperatures are found associated with maximum reception, with minimum temperatures accompanying minimum reception.

There is another meteorological event which repeats at sufficiently frequent intervals for correlation with my night reception data; the passage over Massachusetts of the centers of cyclones and anticyclones, which are the familiar "Lows" and

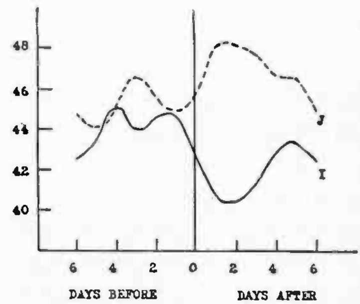


Fig. 8—Boston Temperature Changes Accompanying the Passage of Cyclones and Anticyclones.

"Highs" of the weather map. In Fig. 7 is shown in full-line curve *G* the relation between reception and the passage of Lows, while curve *H* is the relation to anticyclones or Highs. A very distinct inverse relation exists between these two curves, reception showing a decrease before and a rise after the passage of an area of low pressure, while reception is better before and worse after the passage of an area of high pressure.

But cyclones and anticylcones are related to temperature changes, which makes it possible that the relation shown in Fig. 7 is with temperature rather than with pressure. In Fig. 8 is shown in full-line curve *I* the temperature changes at Boston accompanying Lows, while dotted curve *J* represents the temperatures associated with the passage of Highs. A comparison of this figure with the preceding one makes it evident that the temperatures accompanying cyclones and anticyclones are not responsible for the reception changes. The passage of a Low, for example, is associated with a temperature drop a day or two later, whereas reception, which is directly related to temperature, shows a maximum several days after the passage. Similarly, the passage of an area of high pressure is associated with a temperature rise reaching a maximum a day or two after, but reception is at a minimum during and after the passage.

Although the San Francisco-Pasadena series is as yet too short for satisfactory correlation with meteorological elements, I have compared San Francisco temperatures with Pasadena reception and have found no relation, whereas temperatures taken at Los Angeles, only 18 kilometers from Pasadena, show a distinct direct relation to reception. In Fig. 9 full-line curve $K$ is for

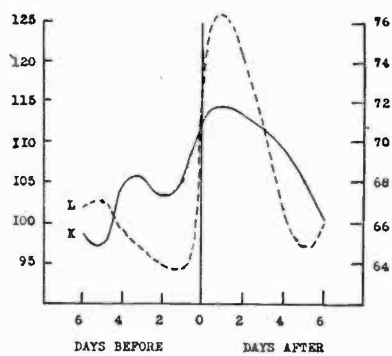


Fig. 9—Pasadena Reception from San Francisco Against Los Angeles Temperature Changes. Rise in Temperature.

Pasadena reception, while dotted curve $L$ is for the accompanying temperatures; a clear direct relation is shown.

Through the kindness of Mr. William K. Aughenbaugh, of Altoona, Pennsylvania, I have received a nightly estimate of signal strength in the 3.75 megacycle amateur band since the middle of October, 1927. Although this series is still far too short for comparison purposes, I find in it a distinct direct relation between reception and Altoona temperatures.

These correlations form an interesting supplement to Austin's original discovery, in that they are the reverse of his resnlts. Just as in the past we have found that solar activity and magnetic storms are associated with improved day and depressed night reception, it now appears that cold waves are related to improved day and lowered night reception; it is another instance of the inverse relation between day and night transmission.

It must be understood that these relations between night reception, temperature and pressure do not hold over long periods, as for example between monthly means of reception and temperature. At least in years of sunspot minima night reception is dictinctly better in winter than summer, but this is not a temperature effect; rather it is because of the longer hours of sunlight and increased absorption by vegetation during the summer months.

Inasmuch as both temperature and pressure are related to solar activity[3] it is not safe to assume that the correlations given above are purely those of cause and effect. And it is possible that changes in the troposphere are associated with events above the isothermal layer, either directly or as common effects of a solar cause. Certainly it is difficult to see how changes of temperature, pressure or humidity can in themselves affect radio transmission.

[3] "Solar Radiation and Weather or Forecasting Weather from Observations of the Sun", H. H. Clayton, Smithsonian Miscellaneous Collection, **77**, No. 6, June 20, 1925; "Solar Activity and Long Period Weather Changes", H. H. Clayton, Smithsonian Miscellaneous Collection, **78**, No. 4, September 30, 1926.

# TECHNICAL CONSIDERATIONS INVOLVED IN THE ALLOCATION OF SHORT WAVES; FREQUENCIES BETWEEN 1.5 AND 30 MEGACYCLES *

By

Lloyd Espenschied

(American Telephone and Telegraph Company, New York City)

THE information given on the accompanying chart is based on the experience of engineers of the Bell System, combined with information secured by engineers of other organizations, and represents what is believed to be the consensus of present knowledge on this subject. The chart was ori-

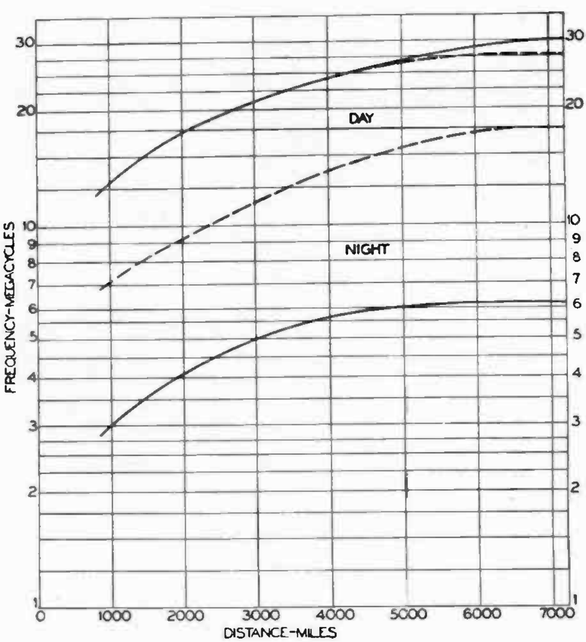


Fig. 1—Approximate Relation of Optimum Frequency to Distance in Short Wave Radio Transmission. 1½ to 30 megacycles, 0 to 7000 miles.

ginally drawn up as something which would be of interest to the Federal Radio Commission and was submitted to the Commission at the hearing upon the allocation of short waves which was held in Washington, January 17 and 18, 1928.

* Original Manuscript Received by the Institute, March 28, 1928.

## Distance-Frequency Relations

The controlling factor in the allocation of short waves is, of course, the relation which exists between frequency or wavelength and the distance over which transmission may take place. The frequency best adapted for a given distance of transmission varies with time of day, with season of the year and with

905-2493

| International Allocation Washington 1927: Frequency Band (Megacycles) | Services | Available Frequency Channels: Theoretical, Telegraph 1000 ~ | Theoretical, Telephone 10000 ~ | Present Practice, Telegraph | Present Practice, Telephone | Group |
|---|---|---|---|---|---|---|
| 28-30 | AMATEURS AND EXPERIMENTAL | 2000 | 200 | 20 | 20 | BAND C LONG DISTANCE-DAY WORLD WIDE IN EFFECT |
| 23-28 | NOT RESERVED | 5000 | 500 | 60 | 60 | |
| [illegible] | [illegible] | [illegible] | [illegible] | [illegible] | [illegible] | |
| 17.8-21.4 | FIXED (BROADCASTING) | 3650 | 365 | 58 | 58 | |
| [illegible] | MOBILE AND FIXED | [illegible] | [illegible] | [illegible] | [illegible] | |
| [illegible] | MOBILE | 700 | 70 | [illegible] | [illegible] | |
| 15.3-16.4 | FIXED (BROADCASTING) | 1050 | 105 | 16 | 16 | |
| [illegible] | FIXED | [illegible] | 70 | [illegible] | [illegible] | BAND B LONG DISTANCE-NIGHT MODERATE DISTANCE-DAY WORLD WIDE AND REGIONAL IN EFFECT |
| [illegible] | [illegible] | [illegible] | [illegible] | [illegible] | [illegible] | |
| [illegible] | FIXED | [illegible] | [illegible] | [illegible] | [illegible] | |
| [illegible] | MOBILE AND FIXED | [illegible] | [illegible] | [illegible] | [illegible] | |
| [illegible] | MOBILE (FIXED) | [illegible] | [illegible] | [illegible] | [illegible] | |
| [illegible] | FIXED (BROADCASTING) | [illegible] | [illegible] | [illegible] | [illegible] | |
| [illegible] | MOBILE | 600 | [illegible] | [illegible] | [illegible] | |
| 9.6-11. | FIXED (BROADCASTING) | 1400 | 140 | 40 | 40 | |
| 8.9-9.5 | FIXED | 600 | 60 | 20 | 20 | |
| [illegible] | MOBILE AND FIXED | [illegible] | [illegible] | 12 | 12 | |
| [illegible] | MOBILE | [illegible] | [illegible] | 13 | 13 | |
| 7.3-8.2 | FIXED | 900 | 90 | 35 | 35 | |
| [illegible] | AMATEURS | [illegible] | [illegible] | 13 | 13 | |
| 6.6-7.0 | FIXED | 325 | 32 | 14 | 14 | |
| 6.1-6.6 | MOBILE (BROADCASTING) | 525 | 52 | 25 | 25 | |
| [illegible] | FIXED | [illegible] | [illegible] | [illegible] | [illegible] | |
| [illegible] | MOBILE | [illegible] | [illegible] | [illegible] | [illegible] | |
| 4.0-5.5 | MOBILE AND FIXED | 1500 | 150 | 98 | 98 | BAND A MODERATE DISTANCE NIGHT SHORT DISTANCE-DAY REGIONAL IN EFFECT |
| 3.5-4.0 | MOBILE FIXED & AMATEURS | 500 | 50 | 40 | 40 | |
| 2.8-3.5 | MOBILE AND FIXED | 650 | 65 | 62 | 62 | |
| [illegible] | FIXED | [illegible] | [illegible] | [illegible] | [illegible] | |
| 2.2-2.7 | MOBILE | 500 | 50 | 62 | 50 | |
| 2.0-2.2 | MOBILE AND FIXED | 250 | 25 | 35 | 25 | |
| 1.7-2.0 | MOBILE FIXED & AM. | 285 | 28 | 45 | 28 | |
| 1.5-1.7 | MOBILE | 215 | 21 | 40 | 21 | |
| | BAND-C | 15000 | 1500 | 206 | 206 | |
| | BAND-B | 9000 | 898 | 272 | 272 | |
| | BAND-A | 4500 | 449 | 417 | 359 | |
| | TOTAL | 28500 | 2847 | 895 | 837 | A. T. AND T. CO. DEP'T. OF DEV. & RES. 1-18-28 |
| SUMMARY | MOBILE | 4165 | 415 | 196 | 165 | |
| | FIXED | 10375 | 1037 | 256 | 256 | |
| | MOBILE SHARED | 5410 | 540 | 320 | 283 | |
| | FIXED SHARED | 5410 | 540 | 320 | 283 | |
| | BROADCASTING | 850 | 85 | 22 | 22 | |
| | AMATEUR | 700 | 70 | 21 | 21 | |
| | AMATEUR SHARED | 2785 | 278 | 105 | 88 | |

Fig. 2

other conditions, such as solar radiation, and is, therefore, a difficult one for definite determination, the more so because these relationships have not yet been very completely explored. Taking the best information we have for average conditions, we find there exists the general relationship between frequency and distance delineated by the curves in Fig. 1.

These curves illustrate phenomena well-known to those familiar with short-wave transmission,—the fact that frequencies which may be suitable for transmission to a given distance at night may be quite unsuitable for daytime communication over the same distance, and vice versa; and that, in general, there are definite limitations in the distance for which the various parts of the short-wave radio spectrum are best adapted at a given time. The curves on this chart should be regarded as general boundaries outside of which it is not usually advisable to choose frequencies. For example, to communicate over a distance of 4000 miles during the day time, it is desirable to select a frequency between about 13.5 and 24 megacycles, preferably one near the middle of this range.

## National and International Aspects

It is convenient to divide the short-wave spectrum into three bands as indicated at the extreme right of Fig. 2. Of course, these bands are not discrete but merge into one another.

Band A—1500 to 6000 kc. (200 to 50 meters). This band is best adapted to communication over moderate distances in the world-wide sense, distances up to, perhaps, 1000 miles at night. The band may, therefore, be considered as regional in its service range. The higher frequencies may cause interference over intercontinental distances at night.

Band B—6000 to 15000 kc. (50 to 20 meters). In its range for communication this band is more or less regional for that portion of the globe which is in daylight, but may include practically the entire hemisphere which is in darkness. During the winter months of the year the daylight hours for the northern hemisphere of the globe are relatively short and the higher of these frequencies are very widespread in their effect during the winter season.

Band C—15000 to 30000 kc. (20 to 10 meters). This band (the higher limiting frequency being somewhat uncertain) appears to be world-wide in its communication range, extreme distances being reached especially over the hemisphere which is in daylight.

For all three bands world-wide coordination is, of course, necessary in respect to the *services* for which the frequencies are used. Futhermore, world-wide coordination in the individual channel assignments to stations will be requried for Bands *B* and *C*, and probably also for the higher frequencey end of Band *A*. For the lower frequency end of Band *A* it may prove to be practical in large isolated areas, as in the North American Continent, to make individual station assignments without requiring coordination with similar assignments in other continents.

### International Frequency Allocation

In the center of the chart are two columns which give the allocation of frequency bands in this range adopted by the International Radiotelegraph Conference, held in Washington in 1927. This allocation becomes effective, for the countries which ratify the Washington Convention, on January 1, 1929. It is, therefore, to be expected that the various national agencies will immediately begin to use this allocation as a guide in making frequency assignments to stations in this range.

### Available Frequency Channels

The number of channels which can be simultaneously occupied by radio stations is definitely limited. In the chart are given figures showing the limiting number of channels for radiotelegraph and radiotelephone communication, assuming channel spacings of 1,000 cycles and 10,000 cycles, respectively. This number is far in excess of the number which can be used practically in the present art.

An estimate based upon present general practice is also given on the chart. Such an estimate will vary considerably, depending upon the engineering assumptions. The present estimate shows something less than 1000 channels, either telegraph or telephone, the limitation being not in the band widths of the channels themselves, but in the separation at present required between them to avoid interference. For services which undertake to be continuous and reliable in operation it must be recognized that two or three or four frequencies per station will be required to cover the various parts of the day and seasons of the year.

The more important of the factors which necessitates a substantial separation between channels are outlined below. The separation may be expected to be diminished in time, as the results of further development work are embodied in practical operation.

(1)—Variation in the frequency of the transmitting station. Current practice at the best stations is to use piezo-electric crystal control with temperature regulation. Without such control and regulation, wide variation in the transmitted frequency results.

(2)—Lack of selectivity of receiving equipment. Any radical improvement in the selectivity of short wave radio receiving equipment is likely to involve complicated apparatus and will, therefore, be somewhat expensive. Development along this line must, however, be expected if the most adequate use is to be made of space in this range.

(3)—Practical factors concerned with the use of these channels, such as the geographical relation between sending and receiving stations, the different types of services required and the divers practices of the operating organizations involved.

# THE NAVY'S PRIMARY FREQUENCY STANDARD*

By

Robert H. Worrall and Raymond B. Owens

(Naval Research Laboratory, Washington, D. C.)

***Summary***—*In this paper is described a constant frequency source which is used as the Navy's primary working standard of frequency. This frequency source is a crystal-controlled oscillator of special design which was developed at the U. S. Naval Research Laboratory, Washington, D. C. This oscillator employs a type of circuit which is rich in harmonics, over 200 of which may be used as fixed standard frequencies.*

*It also describes in detail the method employed to determine the fundamental frequency of this standard in terms of Naval Observatory time to an accuracy of about one part in 100,000. The ultimate standard of frequency is therefore the mean solar day.*

## History

Several years ago when the Navy began to experiment with high-frequency continuous-wave radio communication, the Navy standard of frequency was an absorption type of wavemeter. It consisted of a precision condenser and a number of coils designed to have a constant inductance and a low radio-frequency resistance. A resonance indicator, if desired, could be loosely coupled to this circuit. This standard wavemeter was calibrated by means of the multi-vibrator or other oscillator the output of which was rich in harmonics. The multi-vibrator in turn was adjusted to resonance with a standard 1000-cycle tuning fork. By this means the wavemeter calibration could be maintained by frequent checking to an accuracy of one-tenth of one per cent. For spark and low-frequency continuous-wave transmission, precision of this order was satisfactory.

One of the first results of the advent of high-frequency transmission was to indicate the inadequacy of a frequency standard the calibration of which might be in error by 0.1 per cent in either direction. For at a frequency of 10,000 kilocycles, this

* Original Manuscript Received by the Institute, January 9, 1928; Revised Manuscript Received by the Institute, March 30, 1928.

* Read by Dr. L. P. Wheeler for the authors at meeting of the International Union of Scientific Radiotelegraphy in Washington, D. C., October 13, 1927.

would mean a possible error of 10 kilocycles, plus or minus. Thus at this frequency a transmitter would occupy a band 20 kilocycles wide, which would increase interference between stations, or greatly decrease the number of communication channels by requiring a greater separation in kilocycles between assigned frequencies. Naturally a service wavemeter calibrated from this standard might conceivably be in error by a much larger per cent than the standard itself.

Therefore there was presented to the Naval Research Laboratory for solution a three-fold problem: (1) to develop for the Naval service a frequency standard whose constancy should be as nearly absolute as possible; (2) to provide a means of obtaining direct from this standard a large number of fixed frequencies for calibration purposes; (3) to develop a method for determining the frequency of this standard in terms of observatory time to better than 0.01 per cent, and as near 0.001 per cent as possible.

## Crystal-Controlled Oscillators

The application of the piezo-electric properties of quartz crystals to control the frequency of vacuum-tube oscillating circuits, regardless of small changes in capacity or inductance, suggested a crystal-controlled oscillator as the solution to our first problem. After experimenting with different types of circuits and making a number of refinements in crystal holders a crystal-controlled oscillator was developed which with automatic temperature control was found to give an output of practically absolute constancy.

Tests have shown that the combined effect on the frequency of the standard oscillator caused by five per cent change in plate voltage is less than 0.001 per cent. A ten per cent change in filament voltage produces no noticeable change in frequency. Changing from one tube to another of the same type produces a change in frequency varying from zero to a maximum of about 0.002 per cent. On account of the lack of uniformity in tube constants, several tubes that give the same frequency as the oscillator tube used in the calibration are kept as spares, which practically eliminates the error due to tubes. The temperature coefficient of the standard crystal in its holder is less than one part in one hundred thousand per degree centigrade.

## Types of Circuits

One type of circuit used in the preliminary work was that shown in Fig. 1. This tuned plate circuit,[1] using a small inductance and a relatively large capacity, gives maximum output on the fundamental frequency and a rapidly diminishing amount of energy on the harmonics. In other words, this is a transmitter type of circuit, and only a few harmonics could be

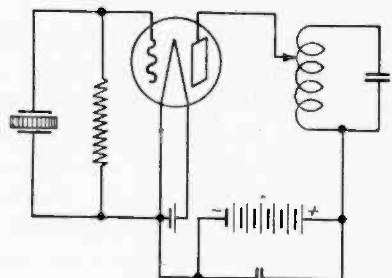


Fig. 1—Tuned Plate Type of Crystal Oscillator Circuit, Having a Low $L/C$ Ratio.

used for calibration purposes. This circuit was therefore modified with the purpose of reducing the strength of the fundamental and intensifying the harmonics, so that each multiple of the crystal frequency could be used as a standard frequency. This

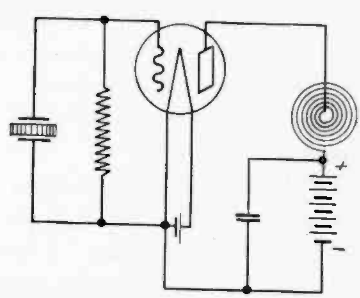


Fig. 2—Circuit Used in Navy Standard Oscillator to Develop Harmonics of Crystal Frequency

modification consisted in replacing the tuned plate circuit, with a very large inductance having a small distributed capacity and no condenser across the coil. See Fig. 2. This untuned coil must have a natural frequency higher than the crystal fundamental; that is, it must be an inductive reactance at the crystal

[1] Developed by Dr. J. M. Miller.

frequency to give the feedback in the right phase relation to the grid potential to sustain oscillation.

To bring out more clearly the difference in the strength of harmonics in these two types of plate circuits, it may be stated that the amount of radio-frequency current in the plate circuit of an oscillator on a given harmonic frequency depends greatly on the ratio of the external impedance at this frequncy to the internal plate impedance of the tube. When these impedances are about equal for a given frequency, we get the maximum energy on this frequency. If the external circuit impedance is small compared to the tube impedance at a given frequency the energy radiated at this frequency will be correspondingly small.

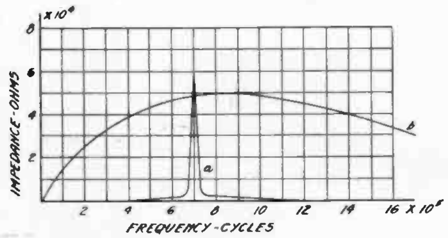


Fig. 3—Impedance Curves (calculated) for Circuits in Figs. 1 and 2.

Now the tuned circuit with a low $L/C$ ratio gives an external impedance of about the order of the tube plate impedance, or greater, over a narrow band of frequencies. See Fig. 3 (a). For instance, the external impedance is relatively very small at a frequency five times the crystal fundamental. Therefore this harmonic would be very weak in comparison with the energy radiated on the fundamental frequency.

On the other hand, a coil with a large $L/C$ ratio gives an impedance curve that is very broad, indicating that the strength of the fundamental and of a large number of successive harmonics will be of the same general order, as indicated in Fig. 3 (b).

This oscillator may be considered as a very low-power crystal-controlled transmitter radiating on perhaps two hundred frequencies each one of which is an integral multiple of the crystal fundamental frequency.

To make use of all these harmonics, as in the calibration of heterodyne frequency meters, it is necessary to couple this oscillator, as well as any oscillator to be calibrated from it, to a detector and amplifier. The crystal oscillator and the detector and amplifier are built into a single unit, which with the automatic temperature control device is called the *crystal-controlled standard oscillator.* See Fig. 4. The input of the detector is untuned and the coil is designed to respond to all frequencies with-

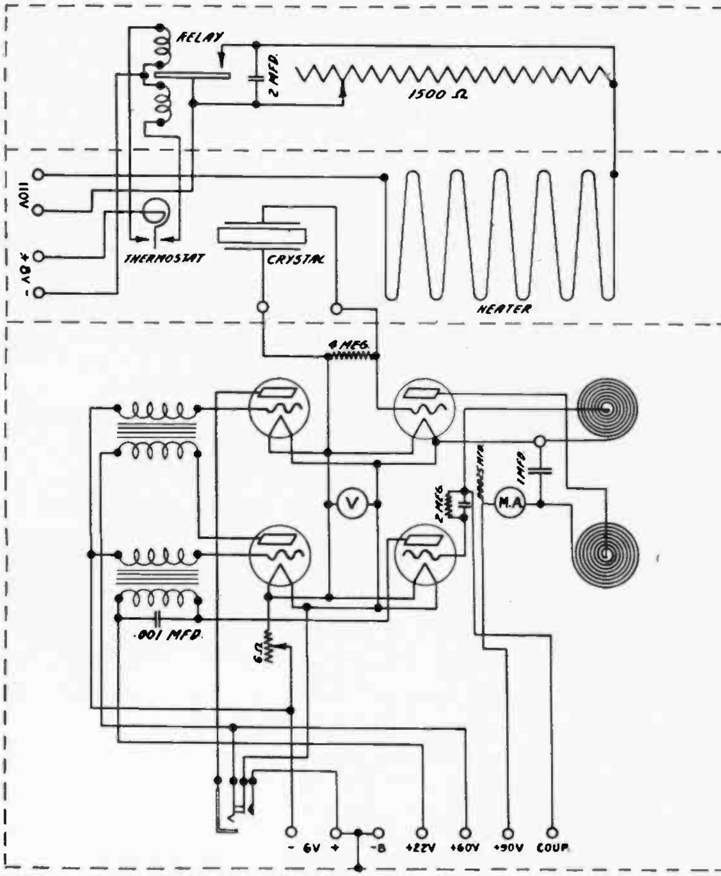


Fig. 4—Circuit of Crystal-Controlled Standard Oscillator Showing Heater Circuit.

out any adjustments. In practice, the coupling is fixed between the crystal oscillator circuit and the detector coil, and a coupling wire is run from the detector sufficiently close to the oscillator under calibration for coupling purposes. When the heterodyne frequency meter (or other type of continuously variable oscillator) is adjusted to within an audible frequency of any harmonic of the cystal oscillator, the beat note between these is rectified by the detector and amplified by one or two stages of amplification.

The decrease in beat note strength between the successive harmonics of the crystal fundamental frequency and another oscillator is too slight to notice. The signal produced by the hundredth harmonic is of the same general order of audibility as that caused by the fiftieth, and the two hundredth harmonic gives a note of easily readable strength. This device is portable as illustrated in Figs. 5 and 6. At the Naval Research Laboratory the frequency of the standard crystal is 25 kc. From this one crystal and the associated circuits as just described, standard

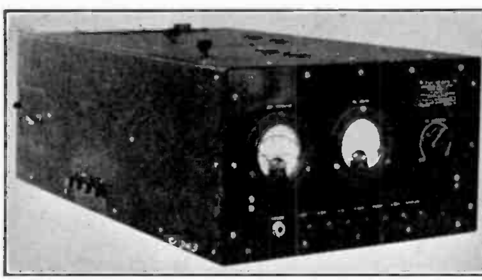

Fig. 5—Outside View of Navy Standard Oscillator.

frequencies are obtained every 25 kc. up to 5000 kc. For a high-frequency range, another standard is provided with a crystal of a frequency of 250 kc. These two frequency standards enable us to cover the entire radio-frequency scale.

## Calibrating the Standard Oscillator

### General Description of the Method Used.

Now that a constant frequency source had been developed, capable of giving approximately two hundred points for calibration purposes, the final problem was to provide a means of determining the frequency of this standard to as near 0.001 per cent as possible.

The method that was finally decided upon lends itself particularly well to the calibration of low-frequency sources, such as the 25 kc. standard, but it is not limited to any such frequency. In brief, the method employed makes use of a motor-driven generator the frequency of which is varied until some harmonic of the generator output gives "zero beat" with the fundamental of the crystal oscillator. Then the generator is operated at this exact frequency for some minutes during which time a record consisting of a dot per second is imprinted upon a sheet of paper on a revolving chronograph drum through the medium of a striker actuated by an electrical circuit operated by a standard chronometer. From the slope of the lines of dots thus made on the chronograph drum, the frequency of the generator is very accurately determined. This frequency, multiplied by the number of the harmonic of the generator which was adjusted to the

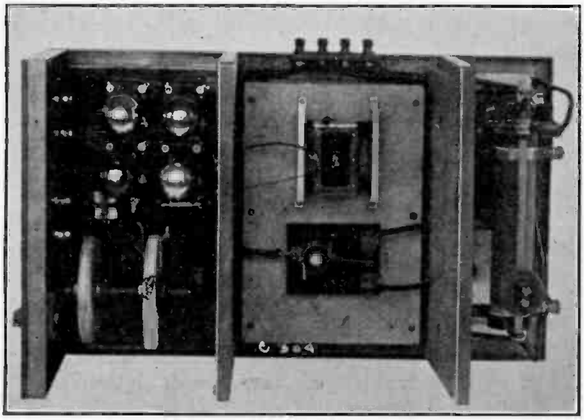

Fig. 6—Standard Oscillator with Cover Removed.

"zero beat" with the crystal oscillator, gives the frequency of the oscillator itself.

To make clear how the method works out in practice, a more detailed description of the apparatus and procedure follows:

**Apparatus Used in the Calibration of the Crystal-Controlled Standard Oscillator.**

(1) A 500-cycle motor generator.

(2) A very accurate chronometer.

(3) A chronograph.

(4) A Maxwell bridge curcuit.

(5) A circuit for the suppression of the 500-cycle fundamental of the generator and the accentuation of the harmonics of the 500-cycle fundamental.

(6) A special amplifier.

(1) The motor generator outfit consists of a five horse power d-c. motor connected by a heavy fly wheel to an alternator

type 500-cycle generator, operated on a high capacity 110-volt battery. (See Fig. 7).

(2) The chronometer is a large Navy standard chronometer adjusted and periodically inspected at the Naval Observatory. The chronometer is checked daily at the Laboratory against the standard time signals. The original rate as determined at the observatory was one and one-half seconds a day. Our observations covering several years show a slight seasonal variation in the rate, about one second a day in summer and one and one-half

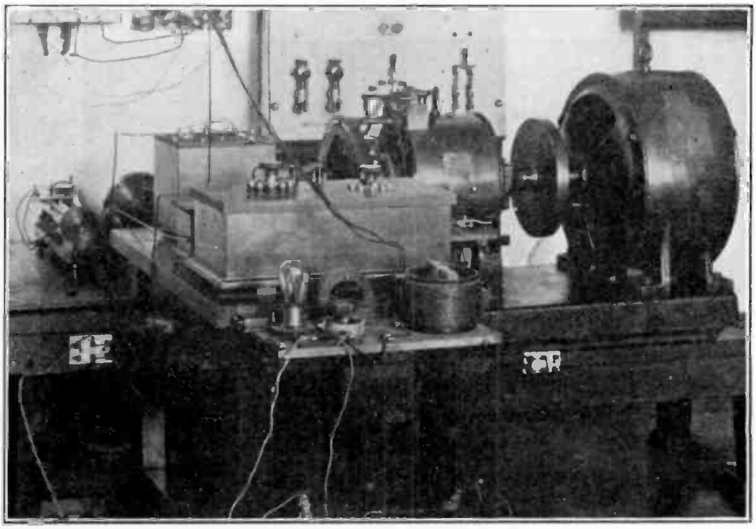

Fig. 7—Apparatus Used in Calibrating Crystal. Motor-Generator at Right, Maxwell Bridge Units in Center, Chronograph Drum at Left Rear.

seconds in the winter. A correction chart has been plotted showing the rate for any time of the year. But when calibrations of the standard crystal are being made, the rate is determined from the daily check against the time signals.

(3) The chronograph consists of the following units; the chronometer, a relay, a striker and a drum. The drum, Fig. 8, is a hollow brass cylinder about 7 in. in diameter and 16 in. long, accurately machined and geared to the generator by means of reduction gears, giving one rotation of the drum to 150 revolutions of the generator shaft. On this cylinder is placed a piece of paper upon which the striker referred to above makes a record as it moves along the face of the drum through the medium

of a worm. The striker records a dot upon the paper once a second (omitting the 59th second every minute) and is actuated by the chronometer. One rotation of the generator shaft corresponds to an alternating potential output of twenty cycles, as can be noted by counting the poles. Now when the drum, operated through the reduction gear just mentioned, rotates once in exactly six seconds, as indicated by six rows of dots exactly parallel to the axis of the drum on the record sheet, the generator frequency is $\frac{150\times 20}{6}$ or 500 cycles per second.

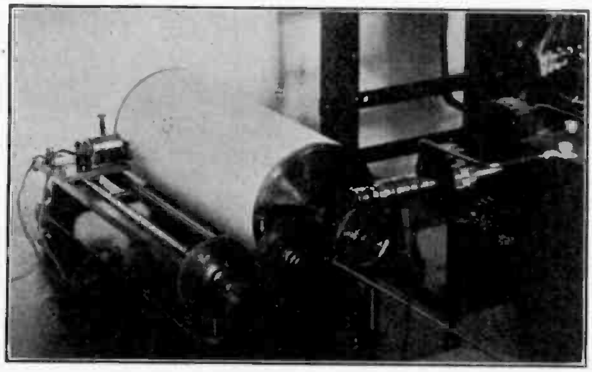

Fig. 8—Chronograph Drum, with Paper Mounted for Calibration Record. Recording Relay at Left, on Carriage.

If the rows of dots slope up the generator output is less than 500 cycles; if they slope down, it is more. From this chronograph record the exact generator frequency during a run is determined, as will be explained in detail further on.

(4) The Maxwell bridge, a diagram of which is shown in Fig. 9, is used in connection with the calibration of the standard crystal as an aid in maintaining the speed of the motor-generator constant at the value desired.

The condenser shown in one arm is charged and discharged from the battery across the bridge by means of a rotating interrupter driven from the generator shaft. The commutator segments of the interrupter operate through three brushes to charge and discharge the condenser.

When the bridge is balanced at a given generator speed, the slightest variation from this speed causes a deflection of the beam of light reflected from the galvanometer mirror on a ground glass scale. The direction of the movement of the light beam indicates whether the generator is running too slow or too fast, thus enabling the operator to vary the pressure on the fly wheel to restore the generator speed to the zero beat condition.

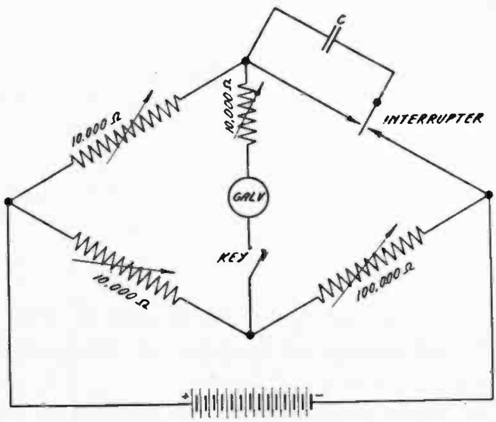


Fig. 9—Maxwell Bridge Circuit.

This use of the generator, chronograph, and Maxwell bridge was patterned after the apparatus used at the Bureau of Standards[2] for the accurate measurements of small capacities.

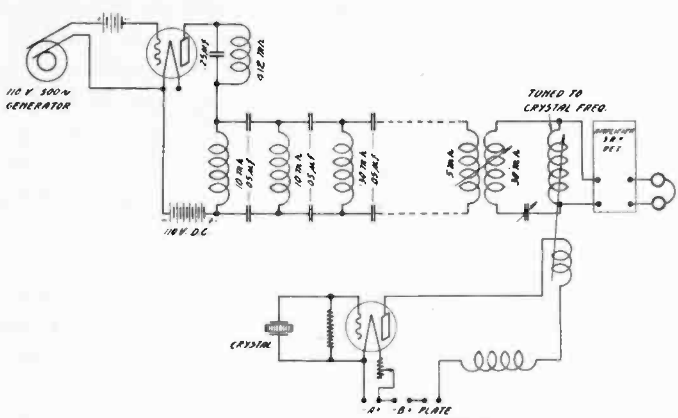


Fig. 10—Crystal Calibrating Circuit.

(5) The method of obtaining harmonics from the output of approximately 500 cycles is to impress this frequency upon the grid circuit of a vacuum tube, the plate circuit of which is

[2] Bulletin of the Bureau of Standards, vol. 3, No. 4.

rich in the required harmonics. As can be seen from Fig. 10 a tuned resonant circuit in the plate circuit partially absorbs the 500-cycle fundamental which, of course, is very strong. The rest of the circuit consists of a filter designed to pass only the high-frequency harmonics of the generator fundamental. This filter circuit terminates in a coupling coil which is coupled to a circuit tuned to the crystal frequency. This tuned circuit is also coupled to an amplifier by means of which the beats between the crystal and the generator harmonics are amplified and transmitted to the phones.

(6) The amplifier consists of three radio stages and a detector.

## Making a Run.

The frequency of the standard crystal is first determined approximately by means of a standard wavemeter. This is accurate to about 0.1 of 1 per cent.

A sheet of paper is placed on the drum and the generator started up. After the generator has warmed up for about 30 minutes the phones from the amplifier shown in Fig. 10 are plugged into the generator room. The operator, with one hand on the motor generator flywheel, varies the generator speed slightly from 500 cycles until some multiple or harmonic of it approaches the crystal frequency when beats are heard in the telephones. The operator now adjusts his speed until slow beats are heard. With the assistance of the other operator the Maxwell bridge is balanced, which brings the light beam to rest on the scale. This is done while the first operator is controlling the slight changes in generator speed by changing the pressure upon the flywheel.

The only function of the bridge is to indicate to the operator by the direction of the deflection of the beam of light whether fast beats in the phones represent an increase or a decrease in generator speed, so that he may quickly increase or decrease pressure on the flywheel to restore the generator speed to its proper value as indicated by very slow beats in the phones. The note in the phones is of the pitch of the generator fundamental frequency, which fact may be made clear as follows: when any harmonic of the generator frequency is at zero beat with the crystal, the harmonics next above and below this one give a beat note with the crystal. If the harmonic at 25,000 cycles is tuned to the crystal, then the one at 24,500 and at 25,500 give a 500-cycle note with the crystal. Pulses or beats in this note indicate how much the 25,000-cycle harmonic differs from the crystal frequency. If a few beats are heard on the fast side, the generator speed is reduced just enough to give the same number of beats on the slow side of the zero beat position. A careful operator may thus compensate for slight changes in beat frequency during a run.

The operator now controls the generator speed to give as slow a beat frequency as possible for a few minutes before starting to record the run. When he notes that the machine is stable enough, he closes the chronometer circuit and the

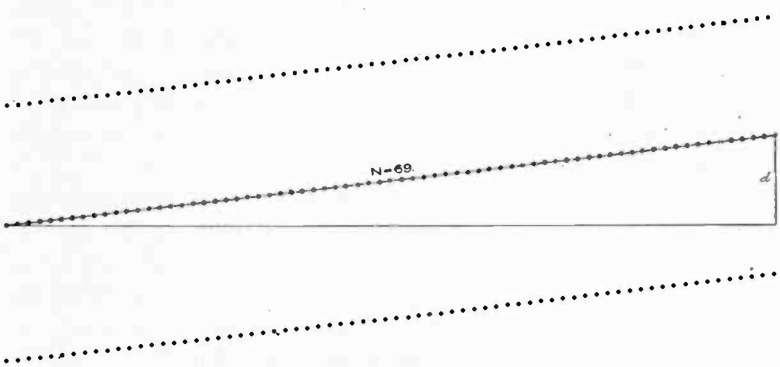


Fig. 11—Representation of Partial Chronograph Record Made in Calibrating a Crystal.

striker commences to put a record upon the paper mounted on the revolving drum. The time required to make a run varies from 8 to 15 minutes.

After the run the second operator may make a duplicate run upon the same harmonic, and then both operators make runs on another harmonic. No error due to the human element is observable in the results.

**Obtaining the Generator Frequency.**

Before removing the paper record from the drum a line is drawn on it parallel to the axis of the drum. The completed record will contain six parallel rows of dots since the drum makes one complete revolution in approximately six seconds, depending upon the frequency of the crystal. The striker puts a dot on the paper every second, while the carriage upon which the striker rests advances approximately 1 millimeter per revolution of the drum. In Fig. 11 a partial representation of a record is shown. This record does not apply to the 25 kc. standard oscillator.

The method for computing the generator frequency is based on the fact that owing to the number of poles in the generator and the ratio of the reduction gear between the generator and the chronograph drum as previously mentioned, a rotation of the drum in exactly six seconds corresponds to a generator frequency of exactly 500 cycles. That is, if the six parallel rows of dots printed on the drum through the chronometer are parallel to the axis of the drum the generator is operating at 500 cycles.

If the slope of these rows of dots is up, the drum has not made a complete rotation in six seconds and therefore the generator speed is less than 500 cycles. A downward slope indicates more than one rotation of the drum in the six second interval or a generator speed greater than 500 cycles.

Assume for example that the rows of dots are parallel to the axis of the drum. Then in 100 rotations a point on the surface of the drum would have traveled 100 times the circumference of the drum, 100 x 51.9 cm. This corresponds to a 500-cycle generator speed. By a simple proportion the generator frequency for any slope of line can be determined thus:

$$CN : CN \pm d = 500 : X$$

where
$C$ = the circumference of drum
$N$ = number of rotations of the drum
$d$ = the perpendicular distance between the first and the last dot in a line
$X$ = the generator frequency.

The $d$ above is the difference between the distance actually traveled by a point on the circumference of the drum during a run, and that corresponding to the nearest number of complete revolutions of the drum. If the slope is up, $d$ is negative; if down, the sign is positive.

Example

$C = 51.9$
$N = 100$
$d = 0.30$ cm. Slope is up. Therefore $d$ is negative.

Generator frequency = 499.971 cycles per second.

From the measurement of the crystal frequency by the standard wavemeter, its value was found to be roughly 25 kc. Therefore, it must have been the fiftieth multiple of the generator fundamental that was held during the run at the crystal fundamental frequency.

The crystal-oscillator frequency is therefore fifty times the generator frequency or 24,998.6 cycles. If the generator speed is increased sufficiently, a point will be found around 510 cycles where the zero beat condition may be again located by means of the phones. This is obviously the forty-ninth harmonic of the generator adjusted to the crystal frequency. ($510 \times 49 = 25{,}000$ approximately).

**Accuracy of the Method.**

The accuracy of a calibration is chiefly dependent on the constancy to which the generator frequency can be held. To assist in this, the motor-generator is operated on a bank of high capacity storage batteries kept in good condition and fully charged. The machine is allowed to warm up thoroughly before a run is attempted. The Maxwell bridge battery is also carefully checked for constant voltage, and the brushes and contactor segments in the bridge circuit are kept clean and properly adjusted. As previously stated, slight variations in the generator frequency can be compensated for by the operator, but if the run is noticeably poor it is so recorded, or discarded.

The error arising from the measurement of the line $d$ is minimized by determining it for all six parallel lines in a record, and using the mean as the correct value for $d$. The resulting error does not average more than one part in 200,000. The combined errors then, on a good run, are well within 0.001 per cent. This degree of accuracy is possible only when the crystal is left undisturbed. The accuracy claimed for the standard crystal oscillator when used as a portable standard is only 0.002 per cent. No appreciable error is observable due to a probable error in the measurement of the drum circumference.

To give an idea of the correspondence of different calibrations a table of eight consecutive runs on a standard crystal is appended. It is, of course, understood that the last figure in the frequency column below is carried along merely to discriminate further between the measured values for different calibrations. It will be noted that the probable error in the mean value is three parts in two and one-half million.

TABLE OF CALIBRATIONS

| Date | Temp. Deg. | Frequency, cycles |
|---|---|---|
| 8-15-27 | 38.45 | 24998.60 |
| 8-15-27 | 38.45 | 24998.75 |
| 8-16-27 | 38.45 | 24998.71 |
| 8-16-27 | 38.45 | 24998.73 |
| 8-29-27 | 38.55 | 24998.99 |
| 8-29-27 | 38.55 | 24998.82 |
| 9- 8-27 | 38.50 | 24998.71 |
| 9- 8-27 | 38.50 | 24998.63 |
| Mean value | = | 24998.74 |

## Uses of the Standard Crystal-Controlled Oscillator

This standard oscillator is a portable device, the size being 8 x 14 x 22 1/2 inches. It is simple to use, there being no variables or tuning controls. It requires a plate potential of only ninety volts, and the heater may be operated on the 110-volt line. By its use, two hundred or more fixed frequencies equally spaced in kilocycles and accurate to about 0.001 per cent are made available. The permanence of the crystal calibration depends only on the permanence of the crystal itself.

This oscillator is used at the Naval Research Laboratory to standardize all quartz crystals manufactured there. Through the medium of a precision heterodyne frequency meter with a straight line frequency characteristic, a crystal of any frequency can be calibrated from the standard oscillator to about 0.002 per cent.

Furthermore, all heterodyne frequency meters for the Naval Service are calibrated directly against the harmonics of a standard oscillator. Readings may be taken as rapidly as the heterodyne frequency meter can be tuned to the successive harmonics of the crystal oscillator. For very accurate work, readings are taken on a beat note of 1000 cycles (as determined by a tuning fork) on both sides of the zero beat point.

The calibration of a heterodyne frequency meter can be rapidly checked to determine the error at any time, by taking readings on a few harmonics of the crystal. For this purpose a special type of crystal oscillator without temperature control has been furnished the Fleet under the name of the "Crystal-Controlled Calibrator," Navy Type No. SE 2907.

The standard oscillator may also be used for the intercomparison of frequency standards. One such comparison has already been made between the Navy's standard crystal-controlled oscillator and the Bell Telephone Laboratories' standard tuning-fork, to determine the relative constancy and the agreement in calibration of the two standards. For this test, the Navy oscillator was transported to New York City. A number of oscillographic records of the beat frequency between harmonics of the two standards taken at intervals over a period of three days indicated that there was no noticeable variation in the relative constancy of the two sources. The determination of frequency by the two standards differed by one part in 80,000 or 0.0012 per cent. This agreement is considered excellent when it is remembered that the Navy standard had been taken from Washington to New York for the test, and that entirely different methods were used to calibrate the two standards.

Since this comparison with the Bell Telephone Laboratories' standard was made, this Laboratory has built four crystal-controlled oscillators similar to the one described in this paper, as a part of a program for the establishment of a national secondary frequency standard decided upon at a conference in Washington in February of last year. This conference was attended by representatives of the various interested government departments, and of a number of the radio manufacturing companies.

It is hoped that, after a determination of the frequency of these oscillators at several other laboratories has been made, and an average result arrived at, they serve as useful secondary standards of frequency for the government and the radio industry of the country.

# A TRANSMITTER MODULATING DEVICE FOR THE STUDY OF THE KENNELLY-HEAVISIDE LAYER BY THE ECHO METHOD*


By

M. A. Tuve and O. Dahl

(Department of Terrestrial Magnetism, Carnegie Institution of Washington, Washington, D. C.)



***Summary***—*The importance is emphasized of sending out "peaks" of very short duration and proper spacing for the study of the radio reflections from the ionized layer in the upper atmosphere by the echo method, and objections to modified alternating-current modulation are pointed out. A method is outlined for modulating the transmitter, based on the sudden pulses of plate-current which occur in an unbalanced multivibrator circuit. The application of this method to the transmitter is described and its effectiveness illustrated by typical recorded signals.*


In the study of the Kennelly-Heaviside layer by the *echo* method[1] a radio transmitter is modulated to emit a set of spaced "peaks" or "humps" of audio frequency, and the resulting "peaks" at the receiver are recorded by an oscillograph. If the transmitted peaks are of sufficiently short duration, single, double, or even multiple peaks are recorded at the receiver, corresponding (under proper conditions) to the ground wave and to zero, one or more "sky waves" arriving at the receiver via the reflecting layer in the upper atmosphere. In the previous work referred to above[2] modulation of the transmitter was accomplished by superposing an alternating e.m.f. of 500 to 1,500 cycles on a high negative bias applied to the grid of the intermediate amplifier in the transmitter. Under these conditions, power is radiated during a part of each positive half cycle of the alternating e.m.f., the time between successive peaks being only slightly greater than their duration or "on-time." Excessively high voltages would be required to make the "on-time" very much less than one half-cycle. The resulting duration of the peaks at 500 cycles was too long to give complete resolution at the receiver for small heights of the layer, and always failed to resolve completely the multiple "reflected peaks" which were


* Original Manuscript Received by the Institute, March 2, 1928.


[1] G. Breit and M. A. Tuve. *Phys. Rev.*, 28, pp. 554–575, 1926. O. Dahl and L. A. Gebhard. Proc. Inst. Radio Engineers, 16, No. 3, March 1928.

[2] Since the work described in this paper was done, a publication by R. A. Heising has appeared (Proc. Inst. Radio Engineers, 16, No. 1, Jan. 1928) containing an account of the work he has done using a similar method. His *peaks* were of duration 0.001 and 0.0016 second, and he did not obtain complete resolution except under unusual conditions.

received at times. To eliminate the possibility that false peaks may be produced by interference effects, it is necessary that the waves shall arrive separately at the receiver, giving peaks which are completely separated on the base-line of the oscillograph record. An additional objection to this method of modulation was that because of the uniform and short spacing between successive peaks there was ambiguity as to whether a given "reflected peak" was due to the "ground peak" immediately preceding it or to an earlier one. Rapid interruption or "keying" of the transmitter eliminated this ambiguity, provided the end of a given signal happened to be recorded on the fast-moving film at the receiver. These objections pointed clearly to the need for a shorter "on-time" and a longer interval between peaks.

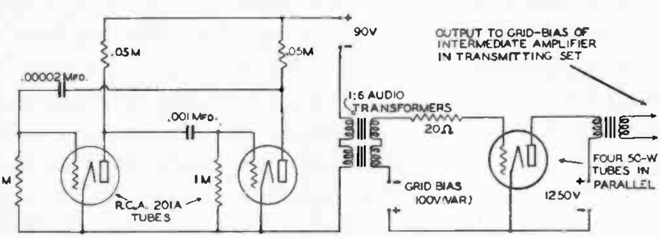


Fig. 1—Diagram of Connections. The Resistance of 20 Ohms in Grid Circuit of 50-Watt Tubes is Put Very Close to the Grids to Prevent Ultra High-Frequency Oscillations.

The use of 1,500-cycle modulation gave peaks of shorter duration, but they were also more closely spaced, and this resulted for the most part in still greater confusion.

When a "multivibrator" circuit as described by Abraham and Bloch[3] is used in a very unbalanced condition, an oscillograph shows the existence of very sharp and widely spaced pulses of current superposed on the otherwise steady total plate-current of the two tubes. The duration and frequency of these pulses may be controlled readily by varying the resistances and capacities interconnected between the plates and grids of the two tubes. Using 201-A tubes with circuit resistances of 0.1 and 1 megohm and capacities of 0.0002 and 0.001 $\mu$fd. (see Fig. 1), the "frequency" of the pulses (their spacing is somewhat irregular) is about 300 per second and their duration perhaps 0.0002 second. This immediately suggests a means of modulating a transmitter for the radio "echo" experiments. A transformer in the plate-supply circuit of these tubes converts a current pulse into two voltage impulses (plus and minus), each of shorter duration than the current pulse. Applied to the grid of an amplifier with a suitably high negative bias, only the voltage impulse in the positive direction affects the tube, resulting in similar pulses of even shorter duration (due to the grid bias) in its plate current. Again transforming these, a sufficient voltage may be obtained to modulate a radio transmitter (see Fig. 1).

Using 201-A tubes for the multivibrator and one 50-watt tube as an amplifier, a transformer in the plate circuit of the 50-watt tube gave peaks of 1,200 volts and of extremely short duration (less than 1/4,000 second). The amplifier gives "clean" pulses of short duration in its plate current only if the input transformer to its grid is so connected that the initial rise of the multivibrator current for one peak superposes a negative e.m.f. on the negative amplifier grid bias, and not vice versa. Thus the amplifier pulses are due to the rate of change of the current at the peak of the pulse, and not at its beginning. This multivibrator and amplifier set was applied to the 20-kilowatt, 4,015-kilocycle transmitter at the Naval Research Laboratory, Anacostia, D. C., which has been used previously in the reflection experiments.[1] In this transmitter two 250-watt tubes in parallel are used as the intermediate amplifier between the crystal-controlled master oscillator and the 20-kilowatt tube, and a change of several hundred volts on the grids of these tubes is sufficient to modulate the transmitter from zero to full power. It was consequently expected that the above multivibrator set would serve very well for direct grid modulation on these tubes. However, when it was connected, the transmitter was modulated by each peak to only a fraction of its full-power emission, as shown by comparison of the received signals using 500-cycle modulation.

When the transmitter is radiating full power, the bias on the grids of the 250-watt tubes is about 100 volts negative. However, the radio-frequency grid voltage superposed on this bias is sufficient to make the grids positive during a fraction of each radio-frequency cycle, resulting in instantaneous values of grid current which may be of considerable magnitude, and it is clear that the device (the modulator) supplying the grid bias must maintain its voltage when this current is drawn, i.e., must supply power to the grids of these tubes. When four 50-watt tubes in parallel were substituted for the single 50-watt amplifier tube,

[3] H. Abraham and E. Bloch. *Annales de Physique* 9, 12, 1919, p. 237.

the resulting output was amply sufficient to modulate the transmitter to full power. Actually the peak value of the power using this modulation is somewhat greater than the usual full power for continuous output.

The resulting modulation of the transmitter is almost ideal for the study of the Kennelly-Heaviside layer by the echo

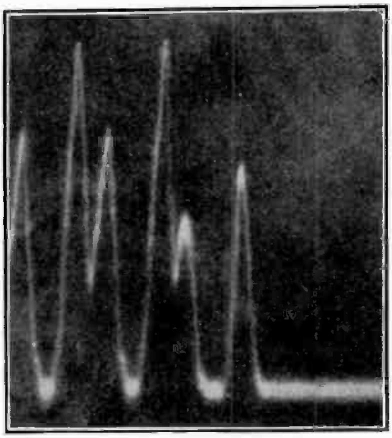

Fig. 2a—Reflections from a 146-mile Layer using 500-cycle Modulation.

method. The "frequency" of the peaks is readily varied, their duration is extremely short (less than 1/4,000 second), and their spacing is somewhat non-uniform. The latter fact is of great usefulness in the identification of the separate peaks due to the


Fig. 2b—Reflections from a 137-mile Layer Using Multivibrator Modulation.

"ground" and "sky" waves. The peak due to the "ground wave" is completely separated from the "reflected peaks" even for very low heights of the layer, eliminating any possibility of interference effects. By reason of the same possibility of interference effects, the brief "on-time" is of great importance in the study of the apparent multiple reflections which occur at times. With this modulation, these multiple peaks are usually completely resolved. Fig. 2a shows reflection from a layer 146 miles high using 500-cycle modulation. Fig. 2b shows reflection from a layer 137 miles high using the multivibrator modulation. The measured apparent width of the peaks is about 1/4,000 second. However, records made with incomplete damping of the oscillograph indicate that this is only an upper limit set by the period of the oscillograph itself. The actual duration of the peaks may be considerably less. The peaks emitted with the multivibrator modulation are not of uniform amplitude, as may be noted. This provides a useful check on the identification of the various peaks, and also on possible distortion in the receiving apparatus.

The writers are indebted to Dr. G. Breit for suggesting the method, and to Messrs. G. Breit, M. H. Schrenk, and L. A. Gebhard for assistance and cooperation in its application.

# A COMPENSATED ELECTRON-TUBE VOLTMETER

By

H. M. Turner

(Yale University, New Haven, Connecticut)

A SOURCE of error sometimes encountered in the use of electron-tube voltmeters is that due to unavoidable changes in filament current especially when the filament is purposely operated well below the saturation value in order to stabilize the calibration over long periods of time. In Fig. 1 is

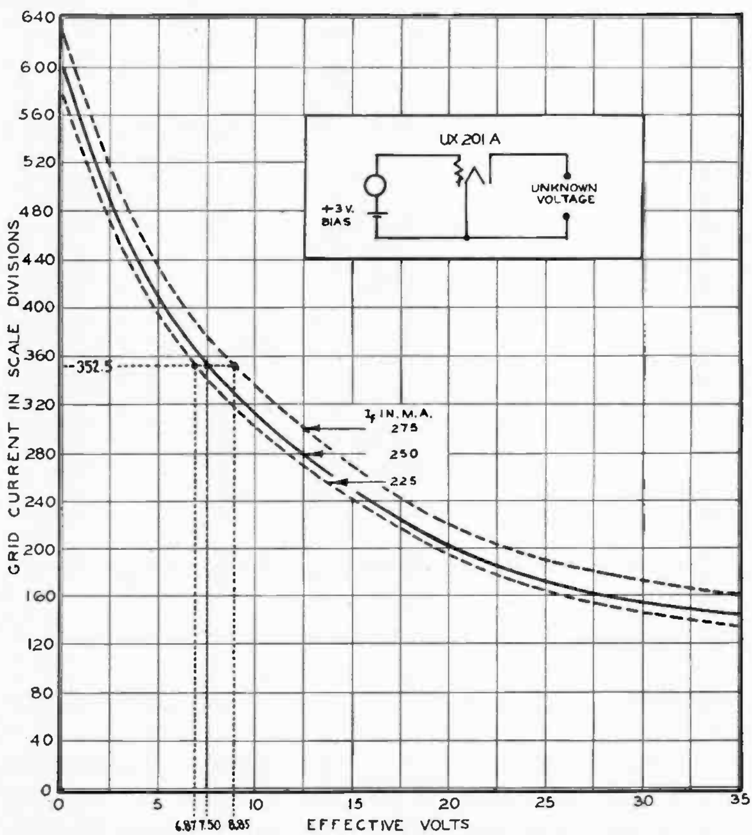


Fig. 1—Uncompensated Electron Tube Voltmeter; Grid Current Method.

shown a circuit arrangement for the "Grid Current Method" together with the calibration for normal filament current of 0.25 amperes. Should the normal calibration be used when the filament current was ten per cent below or above normal the error introduced would be 8.2 and 18 per cent respectively for the plate current used in the illustration, Fig. 1. Of course, such large changes are not likely to occur without the knowledge of

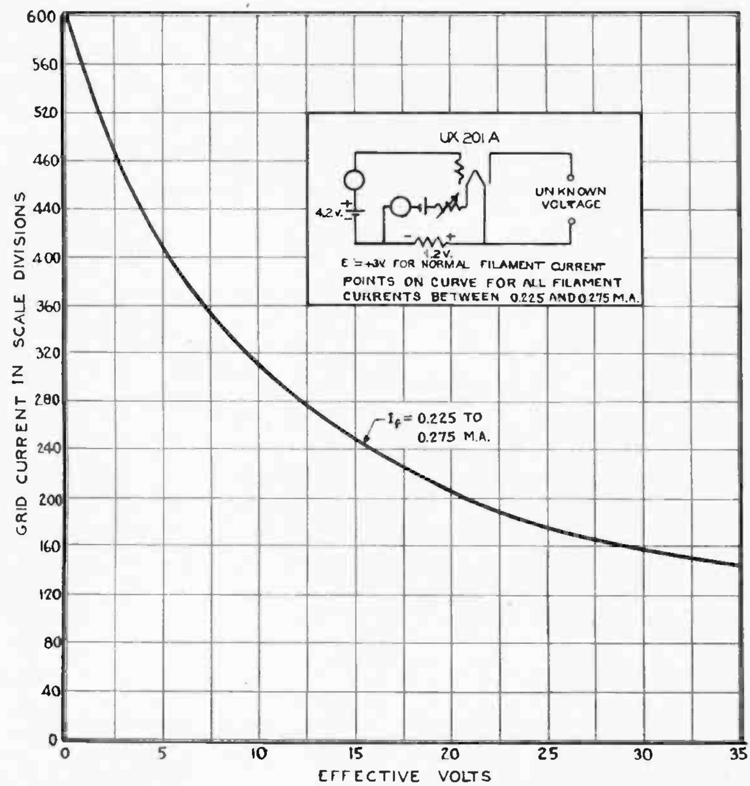


Fig. 2—Compensated Electron Tube Voltmeter; Frid Current Method.

the operator, but these values are used for the purpose of comparison.

A method has been developed which largely eliminates this source of error by causing the grid bias to change with filament current so that the plate current is practically independent of the variation of filament current. Fig. 2 shows the circuit arrangement for the "Compensated Grid Current Method" together with a calibration curve that is good between the limits of 0.225 and 0.275 amperes, and Fig. 3 shows the "Compensated Plate

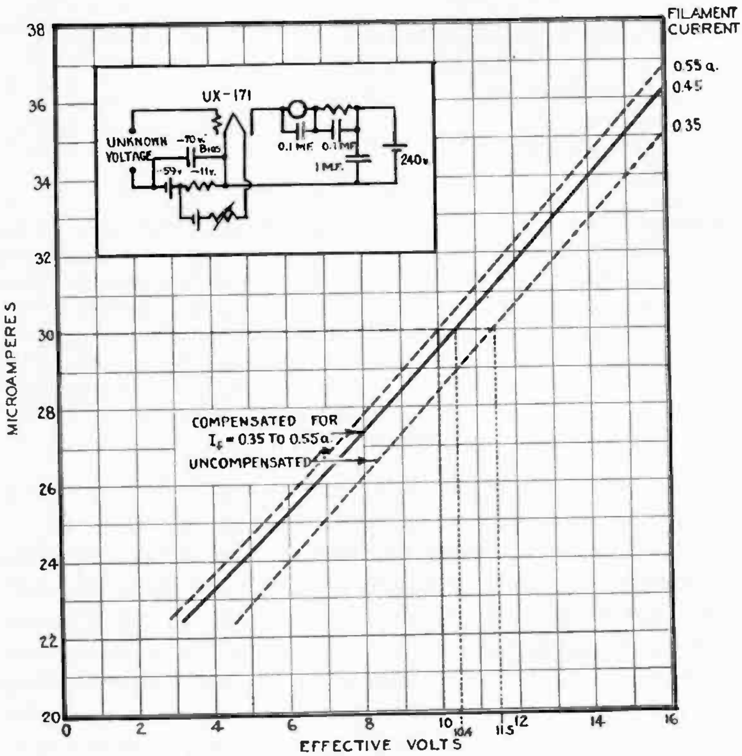


Fig. 3—Compensated Electron Tube Voltmeter; Plate Current Method.

Current Method;" the calibration curve (solid) is good between the limits 0.35 and 0.5 amperes. The dotted curves show that for the same filament current limits the error for the uncompensated method would be 10.6 and 4 per cent respectively.

# NOTE ON THE EFFECTIVE HEATING OF CODE TRANSMITTERS*

By

Frederick Emmons Terman

(Department of Electrical Engineering, Stanford University, California)

***Summary***—*By considering the frequency of letters and words in the English language it is found that in slow speed transmission of code using the theoretical spacing, the transmitter is in operation only 46.5 per cent of the time. A transmitter can accordingly be given a power rating for code that is 2.15 times the rating of the same equipment when operating continuously, on the basis of the same operating temperature. At high speeds of transmission the heating is slightly less than at slow speeds. The desirability of designing large transmitters to take advantage of the full code rating possible during periods of weak signals or heavy static is brought out.*

THE power rating of radio transmitters and radio transmitting apparatus is determined either partially or entirely by the allowable temperature rise during operation. In most types of code transmitters the full power is on only during the part of the time the key is depressed. In rating transmitters and transmitting equipment to be used for code purposes it is accordingly important to know the average rate of heating of the set over a long period of code operation compared with the rate heating that would take place with full power continuously on.

The average rate of heating depends somewhat upon the speed of transmission. At slow speeds the transmitted code impulses are substantially rectangular, as shown in Fig. 1a, but at high speeds the time required for the signal characters to reach approximately full value is an appreciable part of the dot time interval, and the current takes a corresponding time to die out after the key is opened.

The rise and fall of transmitted current follows an exponential law, giving high speed dot and dash characters the shape shown in Fig. 1b.

Taking into account the number of dots and dashes required for each letter, the frequency with which each letter is used in the English language, etc., it is found that a transmitter in the act of sending code with the theoretical spacing is in operation only 46.5 per cent of the time. With slow speed transmission

* Original Manuscript Received by the Institute, April 19, 1928.

the average rate of heating is therefore 46.5 per cent of the rate of heating with full power on. In high speed transmission the average rate of heating is a little less. With an exponential law for the rise and fall of transmitted current, *assuming that the instantaneous heating is proportioned to the square of transmitted instantaneous current,* and neglecting the heating of the small current that remains one dot period after the key has been opened, the following table holds:

TABLE I

HEATING OF CODE TRANSMITTERS

| Completeness of current rise in one dot interval | Average rate of heating with code | Power rating for code / Rating for continuous use |
|---|---|---|
| 1.00 | 0.465 | 2.15 |
| 0.98 | 0.364 | 2.75 |
| 0.95 | 0.347 | 2.88 |
| 0.90 | 0.312 | 3.20 |
| 0.80 | 0.278 | 3.60 |

The first column introduces the transmission speed in terms of the nearness with which the current reaches the steady state

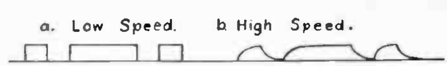


Fig.1—Shape of Transmitted Impulses at High and Low Speeds of Transmission.

value during the sending of a dot. A value of 1.00 represents slow speed transmission shown in Fig. 1a. The second column gives the average rate of heating in terms of the rate of heating when full power is continuously on. The third column is the reciprocal of the second column, and gives the ratio of power rating for code to power rating for continuous transmission for equal average heating (assuming heating is proportional to current squared).

When the thermal capacity of the equipment is large, as in the case of high power transmitters, the operating temperature depends upon the average rate of heating over an appreciable time interval. The third column of Table I shows that such transmitters can safely transmit code at over twice the power output permissible when energy is continuously radiated.

A code transmitter adjusted to give the maximum permissible output when sending code would be seriously overheated if allowed to radiate this energy continuously for a short while Protective relays to prevent such overheating are essential, as continuous transmission is desirable for tuning-in and testing, and forgetful operators will hold the key down. Thermally-controlled relays could be used to lower the power output to the value safe for continuous operation in the event danger of overheating was imminent.

The extent to which the full possibilities of the power rating for code transmission can be taken advantage of with safety depends upon the circumstances. In a large, well-supervised land station, with protective thermally-operated relays, it should be possible to realize all the benefits of the full code rating. Where it is not thought advisable to go to the limit, the service rendered by the station could be very greatly improved by designing the power supply and other equipment to make possible the use of the code power rating during periods of heavy static or weak signals.

The writer is indebted to Dr. J. E. Coover of the Stanford University Psychology Department for information giving the frequency with which different letters are used in English. Computations based on Dr. Coover's count show that 229 words, 1517 dots, and 1008 dashes are transmitted when sending 1000 letters.

# FOUR-ELEMENT TUBE CHARACTERISTICS AS AFFECTING EFFICIENCY

By

David C. Prince

(Research Laboratory, General Electric Company, Schenectady, New York)

***Summary***—*The variation in the ratio of grid and plate current of three-electrode vacuum tubes aroused considerable curiosity and the tests reported in this paper were undertaken in an endeavor to ascertain the laws of current division. It was found that in a tube having symmetrical electrodes, that is, straight wire filament, concentric cylindrical anode and cylindrical grid, made up of wires parallel to the axis, the ratio of grid and plate current was a function of the tube geometry and quite different from that usually found in commercial design. The ratio of grid to plate current in such a tube exceeds the ratio of projected grid area by only a small amount, easily accounted for by variations in the electric field around the grid wires. The considerable departures from this ratio in commercial tubes appear to be due to a combination of secondary emission from the tube anode and unsymmetrical arrangements of grid wires and supports.*

## Object

THIS work was undertaken because of the apparent lack of any logical proved explanation for the amount of current collected by the grid of a three-element tube of standard design when the grid and plate are at nearly the same potential.

At first glance it might appear that a tube having two electrodes, a cathode and a grid shaped anode should have nearly the same impedance characteristics as one having a solid plate. The following qualitative argument should show that this is not so. The current for any potential difference less than temperature saturation values is limited by space charge, that is the cumulative electrostatic field due to the electrons in the space between the electrodes. Individual electrons, when emitted, are acted upon by the electrostatic field of the anode and move in that direction since their initial velocities are relatively small. As the electrons approach the anode their distribution is more or less uniform. Due to their mass they cannot follow the lines of force to the anode wires and, in departing from those lines, some will fail to strike the anode. Those electrons which pass through accumulate in the outside space until their space charge in that zone is sufficient to drive some of them back toward the anode when a

* Original Manuscript Received by the Institute, February 1, 1928.

portion will again pass through. Those electrons which pass a second time add to the charge and therefore reduce the number which can be emitted in unit time. This is equivalent to an increase in impedance.

For a cylindrical tube as long as the motion of the electrons is radial the number of electrons striking the anode on the outward trip should be proportional to the projected area of the anode wires. On the return trip the same fraction in addition should strike the wires. The large additional numbers which do strike must then be due to turbulence.

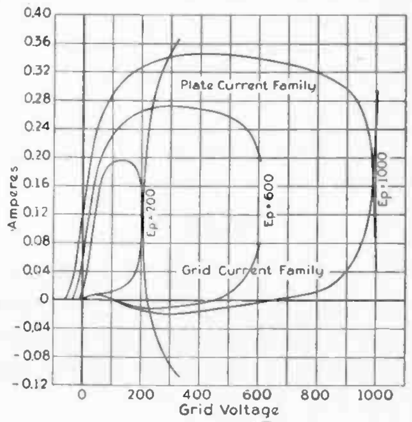


Fig. 1—Characteristics of a 250-Watt Pliotron.

Electrons which have been drawn into space by a positively charged grid structure may be collected by any electrode which lies in their path and is more positive than the cathode. If a large percentage of all the emitted electrons could be collected by an anode at low potential after having been drawn away from the cathode by a highly charged electrode, the total loss of energy in the tube due to space charge could be made very low.

This report covers the making of tests and calculations in a search for the requirements of such a low loss tube.

## Work Performed

In a three-electrode tube a large number of electrons should pass through a positively charged grid even with the plate at cathode or ground potential. With the plate at positive potentials considerable current should flow. That this is not the case is strikingly shown by Fig. 1 which is the characteristic of a 250-watt pliotron. At 300 volts grid and 200 volts plate, the plate not only does not collect electrons, but actually emits them. This is unmistakable evidence that secondary emission from the plate contributes to the supply of electrons which may reach the grid through turbulence. The problem in hand can, therefore, not be studied effectively unless this phenomenon can be excluded. To eliminate secondary emission, recourse was had to a fourth electrode. This electrode was in the form of a second grid placed between the usual grid and the plate. By maintaining this grid below plate potential, a negative gradient is main-

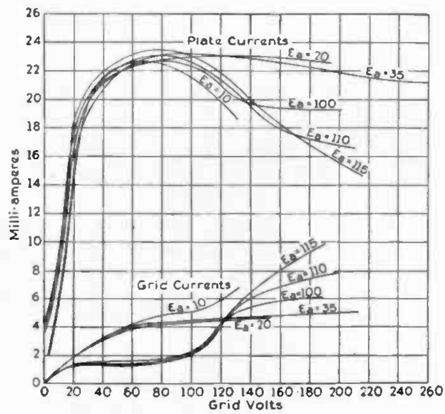


Fig. 2—Characteristics of Tube A with 125 Volts on Plate. $E_a$ is the Anode or Outer Grid Potential.

tained at the surface of the plate, tending to drive secondary electrons back to the plate even though emitted with considerable velocity.

The first of these tubes to be tested was a small receiving type with oval elements designated Tube *A*. The observed characteristics are shown in Fig. 2. It will be observed that the tendency of the plate to lose electrons to the grid is checked by lowering the potentials on the anode or outer grid. With anode volts $E_a$ equal to 10, the plate again loses electrons due to some cause not obvious but probably connected with turbulence effects.

Since the tests of Tube *A* showed that the loss of electrons by the plate and gain by the grid could be positively reduced by the fourth electrode, a larger test sample, Tube *B*, was made up from parts of a 1 kw. pliotron. The inner grid was, however, made with coarser mesh than the standard tube in order that the combined effect of the two grids might be nearly the same as

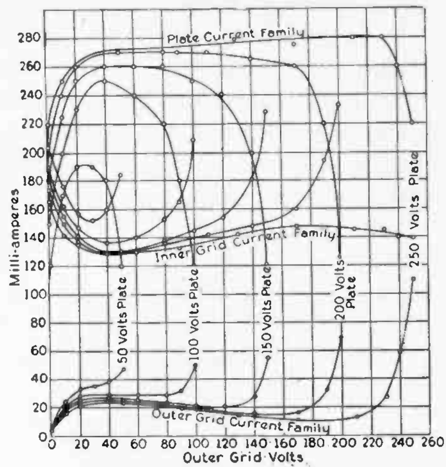


Fig. 3—Characteristics of Tube B with 250 Volts on Inner Grid and 15 Amperes Filament. Principal Dimensions: plate, 3¼ in. long by 1½ in. in diam.; filament, 6⅛ in. of 18 mil. wire (V-shaped); outer grid 1¼ in. in diam. 13 turns per inch of 10 mil. wire; inner grid, ¾ in. in diam. 13 turns per inch of 10 mil. wire.

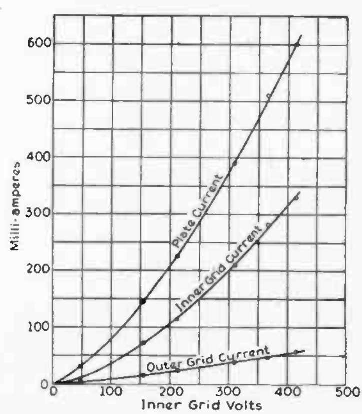


Fig. 4—Characteristics of Tube B with 150 Volts on Plate, 50 Volts on Outer Grid, and 16.5 Amperes Filament.

that of the single grid in the standard tube. The characteristics of this tube are shown in Figs. 3 and 4.

From Fig. 3 it appears that about 150 volts on the plate are required to get substantially full current and that, for this potential, 50 volts on the outer grid give minimum inner grid current. Maintaining these conditions, Fig. 4 gives the variation of current with inner grid voltage. It is apparent that throughout the range of this test the currents to all three positively charged electrodes are substantially proportional and that the plate current is highest even though the inner grid is most positive of any of the elements. Nevertheless the inner grid current shown is out of all proportion to the grid area.

It seemed a fair assumption that the plate secondary emission in Tube $B$ is very small under the conditions used. The large grid current was therefore attributed to lack of symmetry and resulting turbulence. It was decided to construct a tube having the greatest possible symmetry, especially in the inner grid. A sketch of this tube is shown in Fig. 5. The filament, inner grid and plate of this tube are very perfect geometrically. The outer grid is made with longitudinal supports and helical wires, so that its symmetry is much less perfect, but the nature of this structure seemed relatively less important.

Characteristics of this tube, which has been designated Tube $C$, are shown in Figs. 6, 7, 8, and 9. A discussion of these observations follows the discussion of the theory.

## Theory

Assume a cylindrical tube having a cross section such as shown in Fig. 10. Referring to this figure and Fig. 11, which is an enlarged portion of Fig. 10, let:

$r$ = radius of grid

$r_1$ = radius of cylinder which, if substituted for the grid, would produce the same field at the filament

$r_2 = r - c$ = radius of an imaginary polygon

$s$ = distance between grid wires

$\rho$ = radius of grid wires

$p$ = velocity acquired by an electron in passing from the filament to the polygon

$v$ = tangential component of $p$ referred to nearest grid wire (Fig. 11)

$u$ = radial component of $p$ referred to nearest grid wire

$u_0^2 = p_1^2 - p^2$ = increase in square of velocity between polygon and grid wire

$p_1$ = velocity of an electron reaching the grid

$$v = p \sin \alpha$$
$$u = p \cos \alpha$$
$$u^2 + v^2 = p^2$$

$d$ = distance from center line joining filament and grid wire at which an electron can pass the boundaries of the polygon and just come tangent to the grid wire

An examination of the static field produced by a grid shows the irregularities produced by the individual grid wires practically

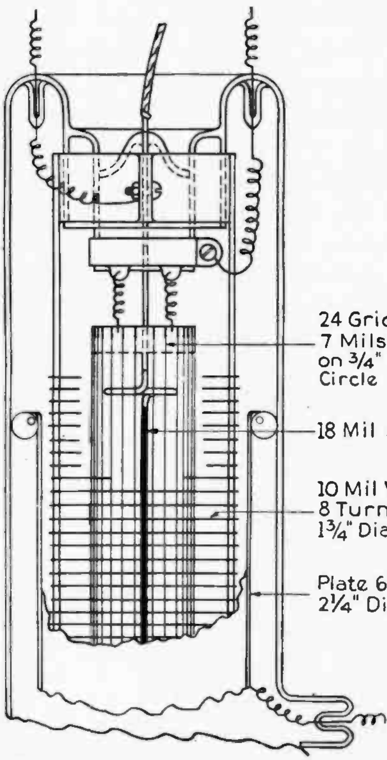


Fig. 5—Tube C—Four-Element Vacuum Tube.

disappear at a certain distance from the grid. Suppose an imaginary polygon to be drawn inside the grid at this distance. It seems reasonable to assume that the motion of an electron may be approximated closely by assuming that it moves inside the polygon in response to a field radial with respect to the filament, and outside the polygon in response to a field radial with respect to the grid wires. Since the electrons which might strike one grid wire are diverging very slowly, the approximation is very little affected by assuming that the electron paths are normal to the sides of the polygon as they cross it.

Under these assumptions

$$\alpha = \tan^{-1}\frac{d}{c}$$

The radius $r_1$ is determined by the formula

$$\log_e\frac{r_1}{r} = \frac{s}{2\pi r}\log_e\frac{s}{2\pi\rho} \qquad (1)$$

This is derived in the conventional way, assuming the grid wires small compared with their spacing and neglecting the presence of the plate.

Inside the polygon the electrons acquire a velocity such that

$$p^2 = Kr_2^{2/3} = K(r-c)^{2/3} \qquad (2)$$

The potential of the grid is the same as that of the equivalent cylinder of radius $r_1$ so that the velocity of an electron reaching the grid is

$$p_1^2 = Kr_1^{2/3} \qquad (3)$$

The increase in the square of the velocity between the polygon and the grid wires is then

$$u_0^2 = p_1^2 - p^2 = K\left[r_1^{2/3} - (r-c)^{2/3}\right] \qquad (4)$$

Dr. Hull[1] has developed the following expression for the condition under which an electron emitted by a cylindrical cathode of radius $R$ with radial velocity $u$ and tangential velocity $v$ will just come tangent to an internal cylindrical anode of radius $\rho$

$$\frac{2e}{m}V = \left(\frac{He}{2\,m}\right)^2(\rho^2 - R^2)\left(1 - \frac{R^2}{\rho^2}\right) + \frac{He}{m}\,\frac{Rv}{\rho^2}(\rho^2 - R^2) - \frac{v^2}{\rho^2}(\rho^2 - R^2) - u^2 \qquad (5)$$

If there is no magnetic field present (5) reduces to

$$\frac{2e}{m}V + u^2 = -\frac{v^2}{\rho^2}(\rho^2 - R^2) \qquad (6)$$

[1] A. W. Hull, "The Effect of a Uniform Magnetic Field on the Motions of Electrons between Coaxial Cylinders," *Physical Review*, **18**, No. 1, July, 1921.

Since $\frac{2e}{m}V$ is the gain in the square of the velocity in passing between cathode and anode and $R^2 = c^2$, (6) may be rewritten

$$\frac{u_0^2 + u^2}{v^2} = \frac{c^2 - \rho^2}{\rho^2} \tag{7}$$

Equation (7) then defines the condition under which an electron will just reach the grid wire. Substituting values

$$\frac{u_0^2 + p^2 \cos^2 \alpha}{p^2 \sin^2 \alpha} = \frac{c^2 - \rho^2}{\rho^2} \tag{8}$$

$$\frac{u_0^2 + p^2}{p^2 \sin^2 \alpha} = 1 + \frac{c^2 - \rho^2}{\rho^2} = \left(\frac{c}{\rho}\right)^2 \tag{9}$$

$$\sin \alpha = \frac{d}{\sqrt{d^2 + c^2}} \text{ and } \frac{1}{\sin^2 \alpha} = \frac{d^2 + c^2}{d^2} = 1 + \left(\frac{c}{d}\right)^2 \tag{10}$$

$$u_0^2 + p^2 = p_1^2 \tag{11}$$

the final velocity of the electron

$$\text{therefore} \left(\frac{c}{d}\right)^2 = \frac{p^2}{p_1^2}\left(\frac{c}{\rho}\right)^2 - 1 \tag{12}$$

From (12) $d$ can be determined, and it represents the distance from the center line joining filament and grid wire at which an electron, crossing the boundary of the polygon, will just come tangent to the grid wire. Electrons passing outside the distance $d$ will not strike the grid wire, while electrons passing inside this distance will strike it.

Since the grid wires are separated by a distance $s$, the length of one side of the polygon is $s\frac{r_2}{r}$. Electrons passing either side of the center line within the distance $d$ will strike the grid. The proportion of electrons which strike the grid is then the ratio of these dimensions, that is

$$\frac{i_g}{i} = \frac{2dr}{sr_2} \tag{13}$$

where $i_g$ is grid current and $i$ total emission.

For the Tube $C$ under investigation

$\rho=0.0035$, $r=0.375$, $s=0.098$, $r_1=0.4$ from (1), $p_1^2/K=0.543$

| | | | | |
|---|---|---|---|---|
| Let $c/s=$ | 0.6 | 0.8 | 1.0 | 1.2 |
| $c$ | 0.0588 | 0.0784 | 0.098 | 0.1176 |
| $r_2=r-c$ | 0.316 | 0.297 | 0.277 | 0.257 |
| $p^2/K=(r_2)^{2/3}$ | 0.464 | 0.445 | 0.425 | 0.405 |
| $(c/\rho)^2=$ | 282 | 500 | 780 | 1120 |
| $(c/d)^2=\dfrac{p^2}{p_1^2}(c/\rho)^2-1$ | 240 | 410 | 610 | 835 |
| $d/c$ | 0.00645 | 0.00493 | 0.00405 | 0.00346 |
| $d$ | 0.00379 | 0.00386 | 0.00397 | 0.00407 |
| $i_g/i=\dfrac{2d}{s}\times\dfrac{r}{r_2}$ | 0.0917 | 0.0995 | 0.110 | 0.121 |

The relation between $c/s$ and $i_g/i$ is shown in Fig. 12.

From Fig. 7, for 250 volts on the inner grid, the emission is made up as follows:

| | |
|---|---|
| Plate current | 1.21 Amperes |
| Outer grid | 0.13 |
| Inner grid | 0.15 |
| Total Emission | 1.49 Amperes |

$$\frac{i_g}{i}=\frac{0.15}{1.49}=0.101$$

Projected area of grid $\dfrac{2\rho}{s}=0.071$

The grid current can thus be determined for this particular tube by assuming $c=0.83s$. The foregoing method of determining grid current is obviously empirical and the results of a test of one tube cannot be considered as quantitatively conclusive. However, the results seem to verify the general theory as conclusively as a single experiment can be expected to do. The main purpose of the investigation has thus been accomplished.

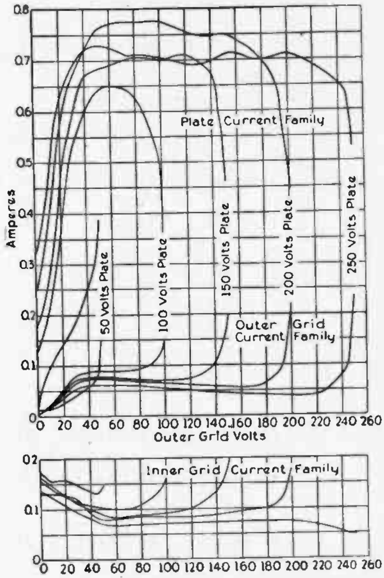


Fig. 6—Characteristics of Tube C with 250 Volts on the Inner Grid and 14.8 Amperes Filament Current.

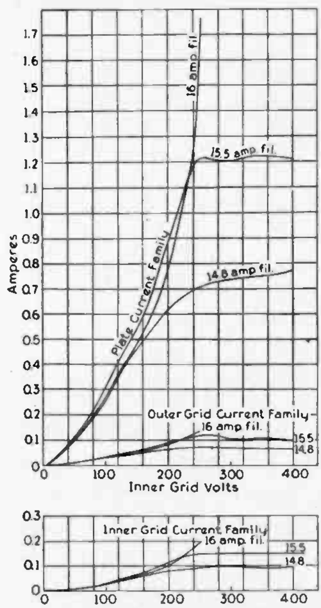


Fig. 7—Characteristics of Tube C with 150 Volts on Plate and 75 Volts on Outer Grid.

It is possible to make other observations of interest from a tube of this arrangement. In the foregoing theory, it has been assumed that all electrons passed in the direction from filament to plate. Suppose the plate to be at zero potential so that it does not receive the electrons. They then return and some strike the grid on the second passage. Let the electrons emitted be $a$, then if $0.9a$ pass the grid and, if none reach the plate, $0.9a$ return and $0.81a$ pass again into the zone between filament and grid. The number striking the grid is $0.19a$. The total electrons

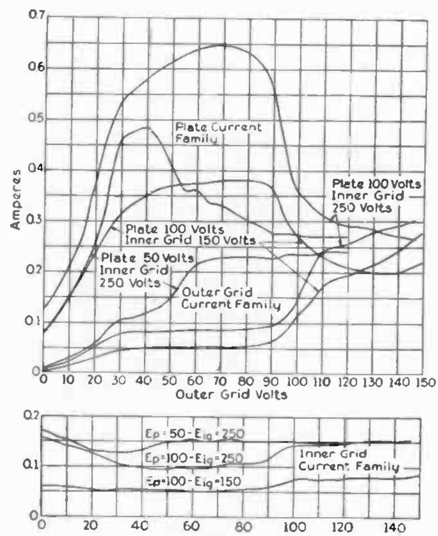


Fig. 8—Characteristics of Tube C with 14.8 Amperes Filament Current.

in the space determine the space charge current so that, if electrons were not passing both ways, the same total number would be emitted. Electrons in filament-grid zone are $1.81a$, and the electrons striking the grid are therefore $\frac{0.19}{1.81} = 0.105$ as compared with 0.100 which would reach the grid were there voltage on the plate. The grid current should, therefore, be 5 per cent higher when there is no voltage on the plate than when there is a positive plate voltage. Fig. 9 shows that this condition probably exists within the limits of experimental error, although the difference might be somewhat greater than 5 per cent due to tangential velocities acquired by the electrons which pass near to grid wires.

Dr. Langmuir[2] has derived relations giving the effect of cathode diameter on space charge. These relations are shown graphically in Fig. 13. These curves apply whether the cathode be inside or outside the anode, providing both are concentric cylinders. When electrons pass from the axial filament through the grid, they immediately begin to produce a space charge outside the grid and they will, therefore, decelerate and lose all

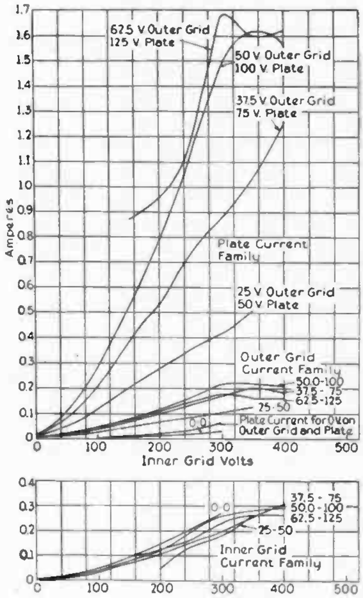


Fig. 9—Characteristics of Tube C with 16.0 Amperes Filament Current.

radial velocity when they have reached a certain distance. Whether the electrons move toward or away from the anode is immaterial, since space charge is a function of the number present and their distribution. The number passing any radius per second is the same and is equal to the current. The velocity distribution is the same since the velocities are zero at the cathode in either case and are equal to a constant times the square root of the voltage at any other radius.

* Irving Langmuir, "The Effect of Space Charge and Residual Gases on Thermionic Currents in High Vacuum," *Physical Review*, **2**, 450-86 1913.

If the electrons are not collected when they come to zero radial velocity, they return again toward the grid and pass through it. The sum of all electrons inside the grid is determined by Fig. 13 for $r<a$; the sum of all electrons outside must be different only by those which strike the grid. A corresponding value of $r/a$ is therefore obtained, giving a zone at which electrons come to rest. This zone then becomes a virtual cathode.

In making an approximate application to the sample tube, assume that the grid has the same effect as a cylinder of the same diameter. For the zone from filament to grid

$$\frac{r}{a}=\frac{0.375}{0.009}=42 \text{ so that } \beta^2=1 \text{ and}$$

$$i=14.65\times 10^{-6}\frac{E^{3/2}}{r}$$

For the outer zone with electrons passing one way only

$$0.9i=14.65\times 10^{-6}\frac{E^{3/2}}{r\beta^2}$$

Dividing
$$\frac{i}{0.9i}=\frac{14.65\times 10^{-6}\dfrac{E^{3/2}}{r}}{14.65\times 10^{-6}\dfrac{E^{3/2}}{r\beta^2}}=\beta^2=1.11$$

for which $\dfrac{a}{r}=2.18$ or $a=0.818$ in. For the case where the electrons pass through the grid but are not collected by the anode and so return, current in the outer zone is

$$\frac{180}{181}i=14.65\times 10^{-6}\frac{E^{3/2}}{r\beta^2}$$

from which $\beta^2=1,\qquad \dfrac{a}{r}=2.1,\qquad a=0.788$ in.

The radius of the outer grid is 0.875 in. We should expect then the equivalent of a three-element tube having a plate of radius 1.125 in., grid of 8 turns per inch 0.01 in. wire with 0.875 in. radius, the cathode varying between 0.788 in. and 0.818 in. radius.

The larger cathode radius would apply to the case where current is being drawn by the plate. However, it seems reasonable to suppose that those electrons which pass close to grid wires will have some of their radial velocity converted into tangential velocity. These electrons would not travel so far from the grid

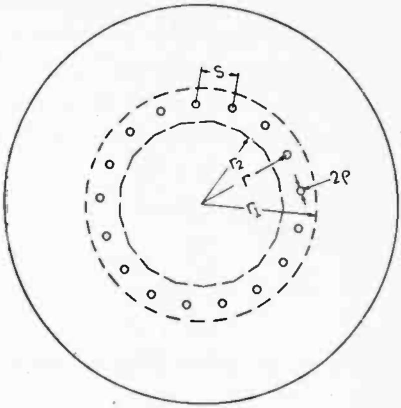


Fig. 10

as those which pass midway between grid wires. This effect would tend to make smaller the effective cathode radius as current draughts increase. Experimental data are not yet sufficient on this point. The apparent effective cathode-grid spacing, from a large number of test points, varies from 0.145 in. to 0.182 in.,

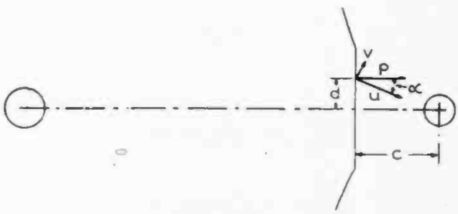


Fig. 11

whereas the theoretical difference is from 0.057 in. to 0.087 in. neglecting the path curvature effect just referred to.

An idea can be obtained from Fig. 6 regarding emission velocities of secondary electrons from molybdenum. The amplification constant of the outer grid referred to the inner grid as anode should be of the order of 25[3] so that the field due to the

[3] F. B. Vogdes and F. R. Elder, "Amplification Constant for Three-element Tubes," *Physical Review*, 21, pp. 683-689, No. 6, Dec. 1924.

inner grid at the plate is small compared with that due to the outer grid. Considering the outer grid only, the electrons appear to begin to leave the plate against a potential of about fifty volts negative on the outer grid.

## Theoretical Conclusions

(1) With a symmetrical grid of the type used, the grid current is always comparatively small. As far as can be told with tests of only one tube, the number of electrons to strike the grid can be determined by assuming that the electrons enter a field which is radial with respect to the grid wires when they

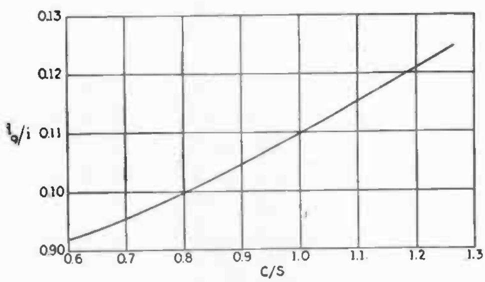


Fig. 12—Variation in Theoretical Grid-Plate Current Ratio with Change in Assumed Equivalent Static Diagram. Projected Grid Area 0.071.

cross a line normal to their paths and removed from the grid wires by a distance equal to 0.83 times the grid wire spacing.

(2) If the plate is protected from emitting secondary electrons, it does not lose current to the grid even though the latter is the most positive element in the tube. The plate voltage current characteristic, under these conditions, corresponds to a tube having a cathode considerably larger than the inner grid. The law covering the exact diameter of the virtual cathode has not yet been established.

(3) To prevent loss of secondary electrons to the grid, a potential on the outer grid, about fifty volts lower than the plate, is required with the tube under test.

## Comparative Conclusions

The practical value of the four-element tube with symmetrical grid lies in the possibility of largely reducing space charge losses without recourse to gas effects, which ordinarily cannot be taken advantage of except at low voltage. Fig. 14 shows the watts per

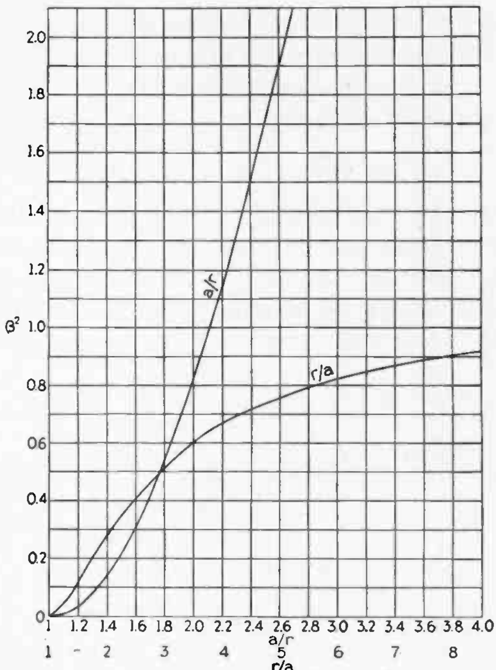


Fig. 13—Space Charge for Concentric Cylindrical Anode and Cathode of Infinite Length.

$$i = 14.65 \frac{E^{3/2}}{r\beta_2} \times 10^{-6}$$

$i$ = current in amperes per cm. $r$ = radius of anode
$E$ = potential difference in volts $a$ = radius of cathode

For large values of $r/a$ the value of $\beta^2$ approaches unity.

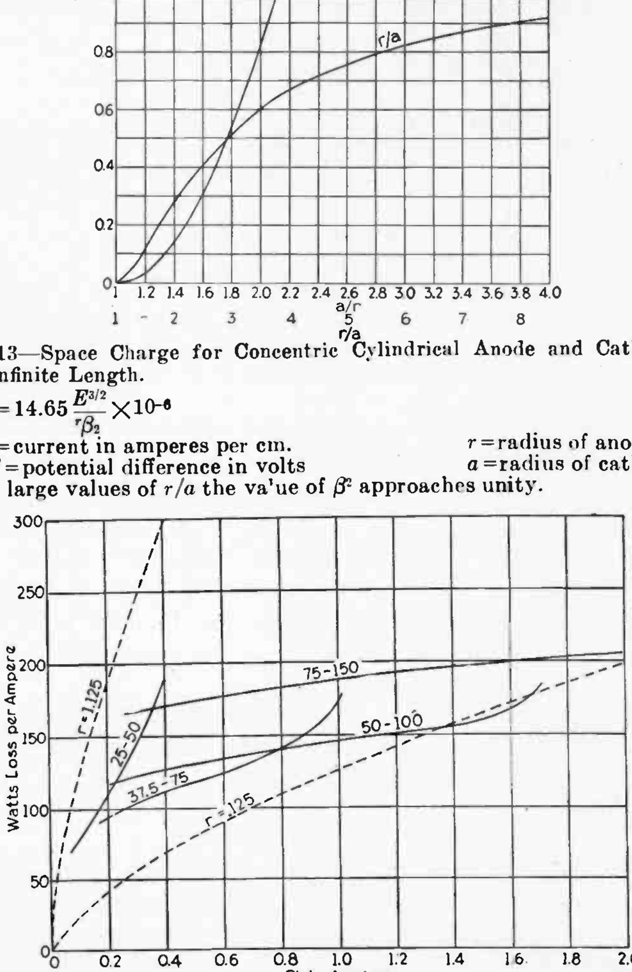


Fig. 14—Loss in Tube C in Watts per Ampere. Solid Curves Are for Tube C. The Figures Indicate Outer Grid and Plate Potentials. Dotted Curves Are for Kenotrons of Radii Indicated.

ampere or equivalent volts drop for the various values of outer grid and plate drop. The figures include watts loss in plate and both grids. For comparative purposes, the losses are plotted for a kenotron having the same plate radius (1.125 in.) and for a kenotron having approximately the same loss. The radius of such a kenotron would be 0.125 in. Such a kenotron would not only be extremely hard to build with large current capacity, but the internal static stresses would probably be prohibitive. The loss per square inch of anode area at 1.6 amperes is 6.22 watts as compared with 58 for the 0.125 in. radius kenotron, not including filament energy.

This four-element tube appears to combine an impedance several fold lower than is obtainable with an ordinary high-vacuum kenotron, and the grid-control and filament shielding of a high-voltage pliotron.

# DETECTION WITH THE FOUR-ELECTRODE TUBE*

By

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***Summary***—*A mathematical analysis of plate rectification is presented. Results of this analysis are applied to the screen-grid tube of the type Cunningham CX-322. Experimental data verifying the mathematical analysis as applied to the CX-322 are presented.*

*The results show that the screen-grid detector tube under proper conditions will efficiently utilize the high radio-frequency voltage obtained with the screen-grid tubes used as radio-frequency amplifiers. The square law holds for large input voltages making it practical for the detector to work power tubes of the type Cunningham CX-371.*

In an investigation of the various applications of the four-element tube of the screen-grid type in the field of broadcast reception, it was evident that the voltage amplification is far above that obtained with an equal number of three-element tubes, and usually higher than that required. When an attempt is made to reduce the number of tubes in the receiver several practical difficulties arise. It is convenient to summarize first the performance obtained with the tubes used, Cunningham type CX-322, before discussing this point in detail.

In service as a radio-frequency amplifier using a tuned output circuit of average good quality a voltage amplification of 25 per stage at 550 kc. and of 40 per stage at 15C0 kc. is readily obtained. This amplification is very much higher than that obtainable with three-element tubes, with which an amplification of ten per stage is perhaps as high a value as is practical to use because of selectivity requirements. In the above case the selectivity using the CX-322 is slightly better than the three-element tube. The simplest and most practical method of coupling radio-frequency amplifiers is the usual method of using a single tuned curcuit between tubes. With the conventional circuits at least three tuned stages will be required to obtain sufficient selectivity.

A high audio-frequency amplification is obtainable, about 40 to 75 per stage, using the CX-322 as a detector and audio-frequency amplifier either as a screen-grid tube or as a space charge tube. This high audio-frequency amplification is not

* Original Manuscript Received by the Institute, March 28, 1928.

always desirable for several reasons. First, the coupling between tubes, due to common voltage supply circuits, becomes more serious and may lead to frequency distortion. Second, microphonic disturbances in the detector are also more serious than usual because of the high audio-frequency amplification.

To take full advantage of the CX-322 amplification it was considered desirable to investigate the detector action of the screen-grid tube with the elimination of the first audio-frequency amplifier in view.

This imposes the following requirements on the detector tube:

First, it must be able to utilize efficiently radio-frequency input voltages of several volts without overloading.

Second, it must be capable of supplying 20 to 30 volts to the grid of the power tube.

A consideration of grid-leak detection was eliminated for two reasons; first, it could not fulfill the first requirement mentioned above, that is, efficiently utilize several volts radio-frequency input without overloading, and second, this method would add damping to the tuned circuit and reduce the selectivity.

Plate rectification using the CX-322 was found to fulfill both of the above requirements imposed on the detector tube provided that a suitable detector output circuit was designed. Either impedance or resistance coupling could be used, but the values required are larger than for a three-element tube because of the high internal impedance of the CX-322.

In this paper the mathematical analysis of the operation of the CX-322 tube used for plate rectification is presented together with data and curves confirming the theoretical results.

The high radio-frequency amplification required to furnish a signal voltage of several volts to the detector can be obtained by taking full advantage of the amplification available with screen-grid tube in the r. f. stages. Hull[1] has obtained an amplification of 40 per stage with an overall amplification of $2 \times 10^6$ at 1000 kc. This amplification is more than ample, under normal receiving conditions, to supply the required voltage to the CX-322 detector. For example, a signal of about 10 microvolts per meter is as small a signal as it is practical to amplify

[1] Hull, "Measurements of High Frequency Amplification with Shielded Grid Pliotrons." *Physical Review*, 27, 4, 439–454; April, 1926.

because of the noise level. As an extreme case, assume that four volts input to the detector is required, (it will be shown later that this value is more than ample). Then with an effective antenna height of four meters, the available signal voltage is 40 microvolts and the r.f. amplification required is 100,000 or only one-twentieth of the total obtained by Hull. Another way of looking at this problem is to assume the first stage of audio has an amplification of about 36. As the response of the CX-322 is proportional to the square of the input voltage, as will be shown later, we will require the square root of 36 or six times the radio-frequency voltage amplification obtained with the average three-element tube set that will satisfactorily operate a power tube.

For a more complete description and theory of the screen grid tube see Hull and Williams.[2] In the theoretical screen-grid tube the mutual conductance is the only factor affecting the output under operating conditions. This is due to the plate current being independent of plate voltage, and also because there is practically no capacity between the control grid and plate, thus reducing all tube capacities to circuit constants. The voltage amplification is equal to the mutual conductance times the external impedance. $\mu$ in this tube varies with the plate, control-grid and screen-grid voltages, but has a definite value as soon as all the voltages are specified. The coefficients will be expressed in turns of the mutual conductance whenever it is possible as this is easily found and it simplifies the analysis.

The basis of attack of the problem is the method given by Llewellyn,[3] in an article on the three-element thermionic tube. The same method is applied here to the case of the screen-grid tube. The coefficients are in a different form from those used by Llewellyn so that it will be necessary to repeat some of his work in the first part in order to complete the analysis.

The impedances offered to the carrier and side bands are different and are worked out completely so that the equations could be applied to intermediate frequencies where the carrier and side bands differ by an appreciable amount. The expression for the input voltage squared will contain all the terms so that the best conditions for the modulated second harmonic of the

[2] Hull and Williams, "Characteristics of Shielded Grid Pliotrons." *Physical Review*, 27, 4, 432–438; April, 1926.

[3] F. B. Llewellyn, "Operation of Thermionic Vacuum-Tube Circuits," Bell System *Technical Journal*, V, 3, 433–463; July, 1926.

carrier may be found if it is desirable to tune to this frequency to increase the selectivity or for any other reason. Approximations are made later and the final results are almost in the same form as the results by Chaffee and Browning,[4] who obtained them by making the approximations first, and neglecting the terms dealing with the second harmonics of the carrier and side bands.

Consider the circuit shown in Fig. 1,

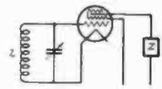

Fig. 1

where $Z$ is a general impedance.

There is no input impedance as there is no mutual capacity between the control grid and plate. The control grid to filament capacity is part of the tuning capacity $C$. This would not be the case for a three-element tube. There is no low-frequency grid impedance; hence, there can be no grid rectification. The control grid is biased negatively so there is no control grid current, making the impedance from the control grid to filament almost infinite.

The plate current is a function of three variables, $E$ plate, $E$ screen grid, and $E$ control grid with $E$ filament constant. The screen grid is at a fixed positive d-c. potential and as there is no impedance in its circuit, it is at zero a-c. potential, making its effect constant. The plate current will be a function of only two variables, $E$ plate and $E$ control grid.

$Eg$ will refer to the control grid and as the screen grid $Ec_2$ is left at a fixed potential the analysis can be made as for a three-element tube.

$$Ip = f(Ep, Eg) \text{ with } Ec_2 \text{ constant.} \tag{1}$$

The following notation will be employed:

$$\begin{aligned} Ip &= {}_0Ip + ip \\ Ep &= {}_0Ep + ep \\ Eg &= {}_0Eg + eg \end{aligned} \tag{2}$$

[4] E. L. Chaffee and G. H. Browning, "A Theoretical and Experimental Investigation for Small Signals," Proc. I.R.E., 15, 2; February, 1927.

where the lower case letters denote variations of the normal or d-c. values of voltages and currents distinguished by the zero subscripts preceding the letters.

$$ {}_0Ip+\delta_0Ip=f({}_0Ep+\delta_0Eg,{}_0Eg+\delta_0Eg). \tag{3}$$

Expanding (3) by the extension of Taylor's Theorem we have,

$$ \begin{aligned} {}_0Ip+\delta_0Ip=f({}_0Ep{}_0Eg)+\frac{\partial_0Ip}{\partial Eg}\delta_0Eg+\frac{\partial_0Ip}{\partial Ep}\delta_0Ep+\frac{1}{2}\,\frac{\partial^2{}_0Ip}{\partial Eg^2}\delta_0Eg^2 \\ +\frac{\partial^2{}_0Ip}{\partial Eg\partial Ep}\delta_0Ep\delta_0Eg+\frac{\partial^2{}_0Ip}{\partial Ep^2}\delta_0Ep^2+\cdots\cdot \end{aligned} \tag{4}$$

Which may be written as

$$ \begin{aligned} ip=\frac{\partial_0Ip}{\partial Eg}eg+\frac{\partial_0Ip}{\partial Ep}ep+\frac{1}{2}\,\frac{\partial_0{}^2Ip}{\partial Eg^2}eg^2+\frac{\partial^2{}_0Ip}{\partial EgdEp}epeg \\ +\frac{1}{2}\,\frac{\partial^2{}_0Ip}{\partial Ep^2}e^2p+\cdots. \end{aligned} \tag{5}$$

Where

$$\frac{\partial_0Ip}{\partial Eg}=\frac{\mu}{rp} \qquad \frac{\partial^2{}_0Ip}{\partial Eg\partial Ep}=\frac{\partial gm}{\partial Ep}$$

$$\frac{\partial_0Ip}{\partial Ep}=\frac{1}{rp} \qquad \frac{\partial^2{}_0Ip}{\partial Ep^2}=-\frac{1}{rp^2}\,\frac{\partial rp}{\partial Ep}$$

$$\frac{\partial^2{}_0Ip}{\partial Eg^2}=\frac{\partial gm}{\partial Eg}$$

and $\mu$ = amplification factor
$rp$ = internal plate resistance
$gm$ = mutual conductance.

Eq. (5) is a power series and may be expressed as follows:

$$ip=a_1eg+a_2eg^2+a_3eg^3+\cdots. \tag{6}$$

Before proceeding further it will be necessary to consider complex quantities. First assume:

$$eg=E\cos\omega t. \tag{7}$$

Now by the well-known theory of the complex variable

$$eg=\frac{E\epsilon^{j\omega t}+E\epsilon^{-j\omega t}}{2}\text{ or }\frac{e}{2}+\frac{\bar{e}}{2}. \tag{8}$$

Where the bar over $e$ indicates that it is the conjugate of $e$.

$$I=\frac{\frac{E}{2}e^{j\omega t}}{Z(j\omega)}+\frac{\frac{E}{2}e^{-j\omega t}}{Z(-j\omega)} \text{ for steady conditions.}^{5} \tag{9}$$

$Z$ for any network may always be expressed as series impedance.

$$Z=r+pL+\frac{1}{pc} \tag{10}$$

For steady conditions $j.\omega.$ may be substituted for $p$ in Eq. (10), this will be done as we are only interested in the steady conditions.

$$Z(j\omega)=r+j\omega L+\frac{1}{j\omega C} \tag{11}$$

$$Z(-j\omega)=r-j\omega L-\frac{1}{j\omega C} \tag{12}$$

$\frac{1}{Z(j\omega)}$ and $\frac{1}{Z(-j\omega)}$ are admittances and may be expressed as:

$$\frac{1}{Z(j\omega)}=a, \text{ and } \frac{1}{Z(-j\omega)}=\bar{a} \tag{13}$$

For detection we will be interested in a modulated radio-frequency voltage of the form

$$eg=A(1+B\cos pt)\cos qt \tag{14}$$

where $q$ is the low frequency
$p$ is the radio frequency
$A$ is the peak value of $eg$
$B$ is the per cent modulation

$$eg=A\cos pt+\frac{AB}{2}\cos(p+q)t+\frac{AB}{2}\cos(p-q)t. \tag{15}$$

$Eg$ from Eqs. (8) and (15) may be expressed as

$$eg=\frac{A}{2}({}_he_1+{}_h\bar{e}_1)+\frac{AB}{4}({}_ke_1+{}_k\bar{e}_1+{}_ne_1+{}_n\bar{e}_1). \tag{16}$$

[5] "Electric Circuit Theory and the Operational Calculus," by John R. Carson, page 9.

where the letter preceding refers to the frequency and the number following to the order

and
$$h=p$$
$$k=p+q$$
$$n=p-q$$

Omitting conjugates, $eg^2$ with its coefficients becomes

$$
eg^2=\left.\begin{array}{ll}
\frac{A^2B^2}{4}\,{}_{k+n}e_2+\frac{A^2B^2}{8}\,{}_{2k}e_2+\frac{A^2B^2}{8}\,{}_{2n}e_2 & (1)\\
\frac{A^2B}{2}\,{}_{h+k}e_2+\frac{A^2B}{2}\,{}_{h+n}e_2+\frac{A^2}{2}\,{}_{2h}e_2 & (2)\\
\frac{A^2}{2}\,{}_{0h}e_2+\frac{A^2B^2}{8}\,{}_{0k}e_2+\frac{A^2B^2}{8}\,{}_{0n}e_2 & (3)\\
\frac{A^2B}{2}\,{}_{h-k}e_2+\frac{A^2B}{2}\,{}_{h-n}e_2+\frac{A^2B^2}{4}\,{}_{k-n}e_2 & (4)
\end{array}\right\} \quad (17)
$$

Line 2 represents the second harmonic of the carrier modulated with the low frequency $q$. Line 1 represents the second harmonics of the carrier and side bands. These second harmonics of the side bands beating against the second harmonics of the carrier would give a frequency $2q$ causing distortion. Thus, if it were desirable to tune to the second harmonic and then to detect again there would be a distortional second harmonic introduced by the first detection.

Line 3 contains the terms causing a change in the direct current of the plate circuit. Line 4 contains the low-frequency modulating frequency $q$ with its second harmonic $2q$, the distortional component.

Eq. (6) becomes, after substituting for $eg$ and $eg^2$ the values found in Eqs. (16) and (17) and neglecting the coefficients $A$ and $B$,

$$
\begin{aligned}
ip= {}& {}_h a_{1\,h} e_1+{}_h \bar{a}_{1\,h}\bar{e}_1+{}_k a_{1k} e_1+{}_k \bar{a}_{1k}\bar{e}_1+{}_n a_{1n} e_1+{}_n \bar{a}_{1n}\bar{e}_1 \\
& +\sum_{hkn} {}_h a_{2\,h} e_2+\sum_{hkn} {}_h \bar{a}_{2\,h}\bar{e}_2+\sum_{hkn} {}_{0h} a_{20h} e_2 \\
& +\sum_{h+k,\,h+n,\,k+n} {}_{h+k} a_{2\,h+k} e_2+\sum_{h+k,\,h+n,\,k+n} {}_{h+k}\bar{a}_2+{}_{h+k}\bar{e}_2 \qquad (18)\\
& +\sum_{h-k,\,h-n,\,k-n} {}_{h-k} a_{2\,h-k} e_2+\sum_{h-k,\,h-n,\,k-n} {}_{h-k}\bar{a}_{2\,h-k}\bar{e}_2
\end{aligned}
$$

$$\text{also } ep = -\sum Zip \tag{19}$$

where $Z$ is the impedance for the currents of different frequencies.

We can write Eq. (5) as

$$ip = \frac{\mu}{rp} eg + \frac{1}{rp} ep + \frac{1}{2} eg^2 \frac{\partial gm}{\partial Eg} + epeg \frac{\partial gm}{\partial Ep} - \frac{1}{2} \frac{1}{rp^2} ep^2 \frac{\partial rp}{\partial Ep}. \tag{20}$$

Substitute the value of $eg$ from (16) and the value of $eg^2$ from (17), neglect the coefficients $A$ and $B$ and equate values of $ip$ from (18) and (20).

Equating the coefficients of $e$ we obtain for the first order effects,

$$\sum_{hkn} {}_h a_1 {}_h e_1 = \sum_{hkn} \frac{\mu}{rp} {}_h e_1 - \sum_{hkn} \frac{1}{rp} {}_h a_1 {}_h e_1$$

or

$${}_h a_1 {}_h e_1 \left[ 1 + \frac{{}_h Z_1}{rp} \right] = \frac{\mu}{rp} {}_h e_1$$

hence

$${}_h a_1 = \frac{\mu}{rp + {}_h Z_1} \tag{21}$$

${}_k a_1$ and ${}_n a_1$ will be of the same form.

This is the usual amplifier equation and will not be further considered here.

Solving for ${}_{h-k} a_2$

$$\begin{aligned} {}_{h-k} a_2 {}_{h-k} e_2 = {} & -\frac{1}{rp} {}_{h-k} a_2 {}_{h-k} e_2 {}_{h-k} z + \frac{1}{2} {}_{h-k} e_2 \frac{\partial gm}{\partial Eg} - \frac{\partial gm}{\partial Ep} {}_k \bar{z}_{1k} \bar{a}_1 {}_{h-k} e_2 \\ & + \frac{1}{2} \cdot \frac{1}{rp^2} \frac{\partial rp}{\partial Ep} {}_h z_1 {}_h a_{1k} \bar{z}_{1k} \bar{a}_1 {}_{h-k} e_2 \end{aligned} \tag{22}$$

or

$$\begin{aligned} 2(h-k) a_2 \frac{rp + {}_{h-k} z}{rp} = {} & \frac{\partial gm}{\partial Eg} - \frac{2\mu_k \bar{z}_1}{rp + {}_k \bar{z}_1} \\ & + \frac{\left(\frac{\mu}{rp}\right)^2 {}_h z_{1k} \bar{z}_1}{(rp + {}_h z_1)(rp + {}_k \bar{z}_1)} \frac{\partial rp}{\partial Ep} \end{aligned} \tag{23}$$

hence

$${}_{h-k} a_2 = \frac{1}{2} \frac{rp}{rp + {}_{h-k} z} \left\{ \frac{\partial gm}{\partial Eg} - \frac{2\mu_k \bar{z}_1}{rp + {}_k \bar{z}_1} \frac{\partial gm}{\partial Ep} \right.$$

$$+\frac{\left(\frac{\mu}{rp}\right)^2 {}_hz_1\,{}_k\bar{z}_1}{(rp+{}_hz_1)(rp+{}_k\bar{z}_1)}\frac{\partial rp}{\partial Ep}\Bigg\}$$

similarly

$$\begin{aligned}{}_{h-n}a_2=&\frac{1}{2}\,\frac{rp}{rp+{}_{h-n}z}\Bigg\{\frac{\partial gm}{\partial Eg}-\frac{2\mu_n\bar{z}_1}{rp+{}_n\bar{z}_1}\,\frac{\partial gm}{\partial Ep}\\&+\frac{\left(\frac{\mu}{rp}\right)^2 {}_kz_1\,{}_n\bar{z}_1}{(rp+{}_hz_1)(rp+{}_n\bar{z}_1)}\,\frac{\partial rp}{\partial Ep}\end{aligned}\tag{25}$$

$$\begin{aligned}{}_{k-n}a_2=&\frac{1}{2}\,\frac{rp}{rp+{}_{k-n}z}\Bigg\{\frac{\partial gm}{\partial Eg}-\frac{2\mu_n\bar{z}_1}{rp+{}_n\bar{z}_1}\,\frac{\partial gm}{\partial Ep}\\&+\frac{\left(\frac{\mu}{rp}\right)^2 {}_kz_1\,{}_nz_1}{(rp+{}_nz_1)(rp+{}_n\bar{z}_1)}\,\frac{\partial rp}{\partial Ep}\end{aligned}\tag{26}$$

Solving for $a_2$ of the form, ${}_{0h}a_2$

$$\begin{aligned}{}_{0h}e_2\,{}_{0h}a_2=&-\frac{1}{rp}\,{}_{0h}a_2\,{}_{0h}e_2\,{}_{0h}z+\frac{{}_{0h}e_2}{2}\,\frac{\partial gm}{\partial Eg}\\&-\frac{\partial gm}{\partial Ep}\Bigg\{{}_h\bar{z}_1\,{}_ha_1\,{}_{0h}e_2+\frac{1}{2}\,\frac{1}{rp^2}\,{}_hz_1\,{}_ha_1\,{}_h\bar{z}_1\,{}_h\bar{a}_1\,\frac{\partial rp}{\partial Ep}\Bigg\}\end{aligned}\tag{27}$$

$$\begin{aligned}{}_{0h}a_2=&\frac{1}{2}\,\frac{rp}{rp+{}_{0h}z}\Bigg\{\frac{\partial gm}{\partial Eg}-\frac{2\mu_h\bar{z}_1}{rp+{}_hz_1}\,\frac{\partial gm}{\partial Ep}\\&+\frac{\left(\frac{\mu}{rp}\right)^2 {}_hz_1\,{}_h\bar{z}_1}{(rp+{}_hz_1)(rp+{}_h\bar{z}_1)}\,\frac{\partial rp}{\partial Ep}\Bigg\}\end{aligned}\tag{28}$$

${}_{ok}a_2$ and ${}_{on}a_2$ are of similar form.

The coefficients for the different second harmonics may be found by the same method.

In most cases

$$pz=p+qz=p-qz \text{ (very nearly)}$$

Then

$$
{}_{h-n}a_2=\frac{1}{2}\,\frac{rp}{rp+qz}\left\{\frac{\partial gm}{\partial Eq}-\frac{2\mu pz}{rp+pz}\,\frac{\partial gm}{\partial Ep}\right.
$$

$$
\left.+\frac{\left(\dfrac{\mu}{rp}\right)^2 pzp\bar{z}}{(rp+pz)(rp+pz)}\,\frac{\partial rp}{\partial E}\right\} \tag{29}
$$

and

${}_{h-n}a_2 = {}_{h-k}\,a_2$ and ${}_{n-k}\,a_2 = {}_{h-k}\,a_2$ except that $2qz$ in ${}_{n-k}\,a_2$ replaces $qz$ in ${}_{h-k}\,a_2$. (30)

Also,

$$
{}_{0h}a_2=\frac{1}{2}\,\frac{rp}{rp+R}\left\{\frac{\partial gm}{\partial Eg}-\frac{2\mu pz}{rp+pz}\,\frac{\partial gm}{\partial Ep}\right.
$$

$$
\left.+\frac{\left(\dfrac{\mu}{rp}\right)^2 pzp\bar{z}}{(rp+pz)(rp+pz)}\,\frac{\partial rp}{\partial Ep}\right. \tag{31}
$$

where $R$ is the d-c. resistance of $Z$.

$$
{}_{0k}a_2={}_{0n}a_2={}_{0h}a_2 \tag{32}
$$

$$
{}_{q}i_{p}=\frac{rp}{rp+qz}\left\{\frac{\partial gm}{\partial Eg}-\frac{2\mu pz}{rp+pz}\,\frac{\partial gm}{\partial Ep}\right.
$$

$$
\left.+\frac{\left(\dfrac{\mu}{rp}\right)^2 pzq\bar{z}}{(rp+pz)(rp+p\bar{z})}\,\frac{\partial rp}{\partial Ep}\right\}\frac{A^2B}{2} \tag{33}
$$

and

$$
{}_{2q}i_{p}=\frac{1}{2}\,\frac{rp}{rp+2qz}\left\{\frac{\partial gm}{\partial Eg}-\frac{2\mu pz}{rp+pz}\,\frac{\partial gm}{\partial Ep}\right.
$$

$$
\left.+\frac{\left(\dfrac{\mu}{rp}\right)^2 pzp\bar{z}}{(rp+pz)(rp+p\bar{z})}\,\frac{\partial rp}{\partial Ep}\right\}\frac{A^2B^2}{4} \tag{34}
$$

$$
{}_0i_p=\frac{rp}{rp+R}\left\{\frac{\partial gm}{\partial Eg}-\frac{2\mu pz}{rp+pz}\,\frac{\partial gm}{\partial Ep}+\frac{\left(\frac{\mu}{rp}\right)^2 pzp\bar{z}}{(rp+pz)(rp+p\bar{z})}\,\frac{\partial rp}{\partial Ep}\right\}\left\{1+\frac{B^2}{2}\right\}\frac{A^2}{4}. \tag{35}
$$

The factor $\frac{\partial gm}{\partial Eg}$ may be found graphically by plotting $gm$ against $Ec_1$ and drawing tangents. Typical static curves for $\frac{\partial gm}{\partial Eg}$ are shown in Figs. 2 and 3. The dynamic curves will be dis-

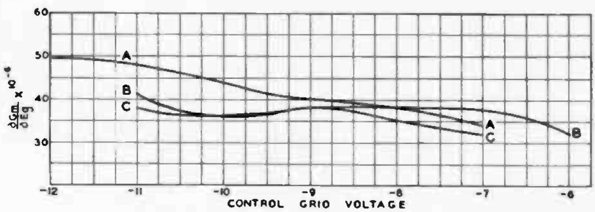


Fig. 2—CX-322; $\frac{\partial Gm}{\partial Eg}$ Control Grid Voltage Characteristics.
A—Static Value $Eb=135$, $Ec_2=67.5$
B—Dynamic Value $Eb=225$, $Ec_2=67.5$
C—Dynamic Value $Eb=135$, $Ec_2=67.5$

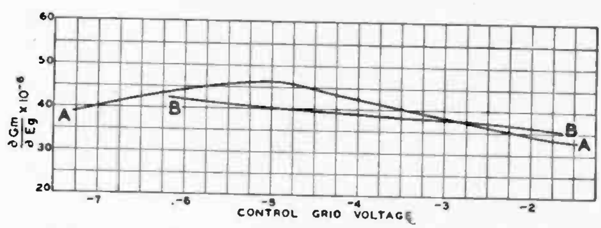


Fig. 3—CX-322. $\frac{\partial Gm}{\partial Eg}$—Control Grid Voltage Characteristics.
A—Static Value $Ec_2=+45$, $Eb=135$
B—Dynamic Value $Ec_2=+45$, $Eb=135$

cussed later. The noticeable feature about these curves is that under the d-c. conditions given $\frac{\partial gm}{\partial Eg}$ remains practically constant. It was found that if $Ec_2$ was varied between about 30 and 67.5 and $Ec_1$ adjusted to give a plate current of approximately 100 microamperes with no signal voltage, the values of $\frac{\partial gm}{\partial Eg}$

were practically constant but increased slightly as the screen-grid voltage was lowered. The region where $Ec_2$ was under 30 volts was not investigated because it was desired to keep the control grid bias at least 7 volts negative.

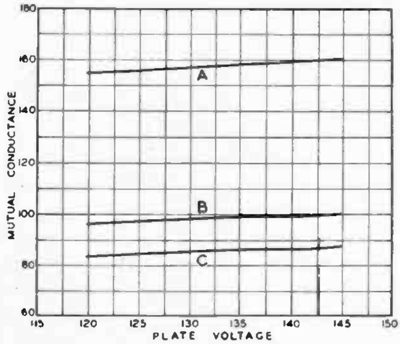


Fig. 4—Mutual Conductance—Plate Voltage Characteristics of CX-322.
A—$Ec_1 = -1.5$, $Ec_2 = 22.5$
B—$Ec_1 = -7.5$, $Ec_2 = 45.0$
C—$Ec_1 = -12.0$, $Ec_2 = 67.5$

The next two terms may be neglected as their values are small. Fig. 4. shows $gm$ plotted against $Eb$. The resultant slope is very small so $\dfrac{\partial gm}{\partial Eg}$ is very nearly zero. The factor $\mu$ is large but

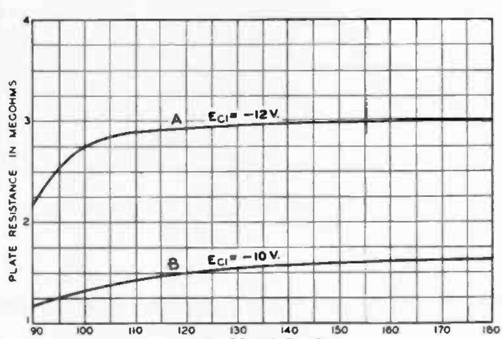


Fig. 5—Plate Resistance—Plate Voltage Characteristics of CX-322.
A—$Ec_1 = -12.0$, $Ec_2 = +67.5$
B—$Ec_1 = -10.0$, $Ec_2 = +67.5$

$rp$ is very large so that the factor $\dfrac{\partial gm}{\partial Ep}$ is small under detection conditions making the effect of the first term negligible. The value of $rp$ plotted against $Eb$ is shown by Fig. 5. This variation is rather large but $rp^2$ occurs in the denominator and $gm^2$ in the numerator of the coefficient multiplying $\dfrac{\partial rp}{\partial Ep}$ so that the second term is also small.

As $\dfrac{\partial qm}{\partial Eg}$ is practically constant and the other two terms

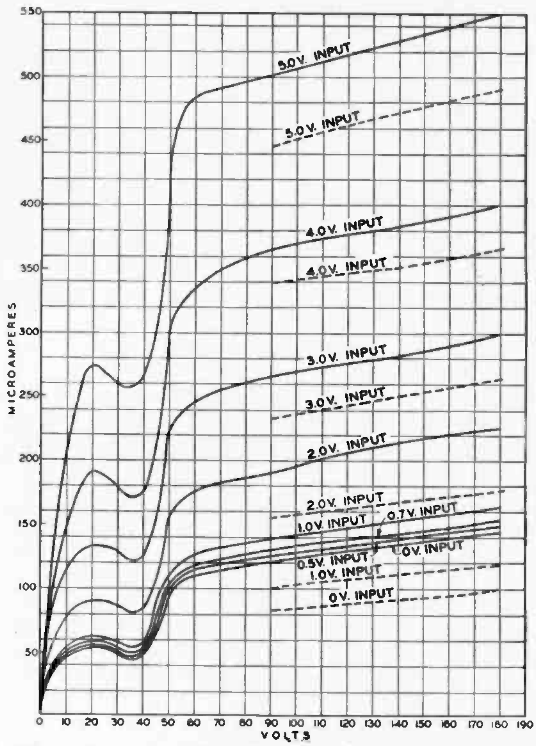


Fig. 6—Plate Current—Plate Voltage Characteristics of CX-322. Variable Input Voltages.
$Ec_1 = -7.5$; $Ec_2 = +4.5$
60 cycles. 1,000 kc.
Note: Different tubes used to show curves are nearly parallel.

can be neglected the theoretical analysis holds for input voltages where the peak values are almost as large as $Ec_1$. If the last two terms could not be neglected it would be almost impossible to evaluate these two terms for large input voltages. Experimental proof will be presented showing that these last two terms may be neglected and that the analysis holds for large input voltages.

Refer to Eq. (35). If $R$ is made zero ${}_pZ$ and its conjugate ${}_p\bar{z}$ are also zero and Eq. 35 becomes

$$ {}_0ip = \frac{\partial gm}{\partial Eg}\left(1+\frac{B^2}{2}\right)\frac{A^2}{4}. \qquad (36)$$

If a known input voltage of any frequency is introduced on the grid the value of $\frac{\partial gm}{\partial Eg}$ can be determined from the direct current change if $B$ is zero or if its value is known. The dynamic values of $\frac{\partial gm}{\partial Eg}$ were found by this method. The agreement between the static and dynamic values is fair considering the difficulties of drawing accurate tangents to curves and the number of times experimental errors enter into the results. As far as the author knows this method of obtaining $\frac{\partial gm}{\partial Eg}$ is a new one.

Fig. 6 shows the plate-current change plotted against plate voltage. The full lines are for 60 cycles and the dotted lines are for 1000 kc. The reason the lines are displaced is that two different tubes were used. The lines are approximately parallel. A condenser was placed across the control grid to filament for the 60 cycles input so as to offer the same impedance as the tube capacity would have at 1000 kc. These curves prove that the value of $\frac{\partial gm}{\partial Eg}$ is very nearly constant for any input voltage nearly as great as the control grid bias under proper conditions, but falls off slightly as the input voltage is increased.

Fig. 7 shows the plate current plotted against plate voltage with different load lines drawn in from 270 volts. The output voltage for a given input voltage and load may be obtained from these curves. For example, with 400,000 ohms load and 2 volts r.m.s. input modulated 50 per cent the output voltage would be 52 volts. This is obtained by taking point $A$ as a reference. The a-c. voltage across the resistor for the above conditions would vary between $B$ and $C$ so the output voltage would be the difference, approximately 52 volts. This is the total swing so the peak value is one half of this, or 26 volts. For 40 per cent modulation the peak value would be eighty per cent of this, or 20.8 volts. This agrees with the results found analytically by another method and also with that found experimentally. For a given input voltage it is only necessary to draw two other curves as, for example, with the curve drawn

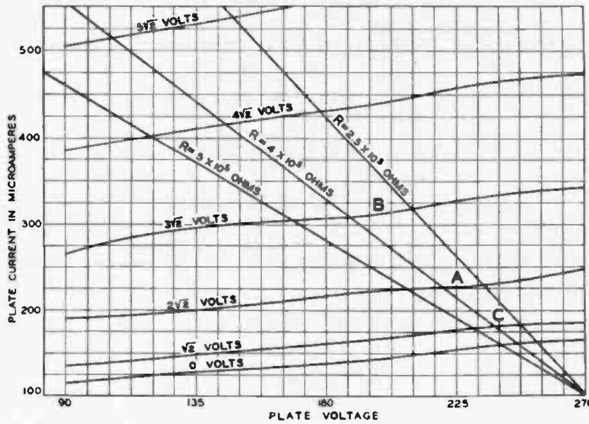


Fig. 7—Plate-Current, Plate-Voltage Characteristics of CX-322. Variable Input Voltages.
$Ec_1 = -12.3$
$Ec_2 - +67.5$
Input at 1,000 kc.

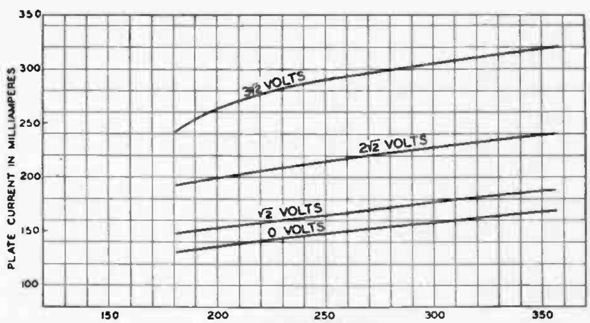


Fig. 8—Plate-Current, Plate-Voltage Characteristics of CX-322.
R=400,000 ohms
$Ec_1 = -12.0$ v.
$Ec_2 = +67.5$ v.

for $2\sqrt{2}$ volts input draw curves for $\sqrt{2}$ and $3\sqrt{2}$ volts input, find the swing at 50 per cent modulation, then as the output voltage is proportional to the modulation its value for any other value of modulation is easily found. The voltage found includes the second harmonic also.

Fig. 8 shows the d-c. plate current plotted against $Eb$ with 400,000 ohms in the plate circuit. The values of $pZ$ and $p\bar{Z}$ are not zero here so if the last two terms of the coefficients of Eq. 35 are not small the d-c. should change when the external impedance is changed. The resistor was shorted for r.f. voltages by a condenser, but no appreciable change could be seen in the direct current.

Eqs. (33), (34), and (35) become

$$_q\dot{i}_p=\frac{rp}{rp+qz}\,\frac{\partial gm}{\partial Eg}\,\frac{A^2B}{z}. \tag{37}$$

$$_{2q}\dot{i}_p=\frac{rp}{rp+2qz}\,\frac{\partial qm}{\partial Eg}\left\{\frac{A^2B^2}{4}\right\}. \tag{38}$$

$$_0\dot{i}_p=\frac{rp}{rp+R}\,\frac{\partial gm}{\partial Eg}\left\{1+\frac{B^2}{2}\right\}\frac{A^2}{4}. \tag{39}$$

Fig. 8 departs from the square law for two reasons; first, a decrease of $\frac{\partial gm}{\partial Eg}$ with input voltage, and second, $rp$ decreases with input voltage. If $R$ remains constant the value of $\frac{rp}{rp+R}$ will decrease with a decrease of $rp$. Fig. 9 shows value of $rp$ plotted against $Ib$ with $Ec_2$ and $Eb$ remaining constant. If $R$ is 400,000 ohms and $rp$ varies from four megohms to one megohm the value of the fraction $\frac{rp}{rp+R}$ varies from 0.91 to 0.71. These would be extreme variations. Some variation from the square law is expected from the theory taking everything into account.

In order to determine the value of the input voltage required to work a power tube an example will be worked out. Resistance coupling will be used to couple the plate of the detector to the power tube. The circuit used is shown by Fig. 10.

The following values were used:

$$R=qz=2qz=5\times10^5 \text{ ohms}$$

$$R \text{ grid } =2\times10^6 \text{ ohms}$$

$$C \text{ coupling}=0.2\times10^{-6} \text{ farads}$$

The a-c. impedance of the combination is practically $4\times10^5$ ohms. Fig. 8 is plotted with an impedance of $4\times10^5$ ohms.

When $B$ is zero the change in d-c. determines the value of $\frac{rp}{rp+R}\frac{\partial gm}{\partial Eg}\frac{A^2}{4}$ as can be seen by referring to Eq. (30). This is one-half of the value of the coefficient of $B$ in Eq. (36). The current $qip$ will be two times the change in d-c. found from curve

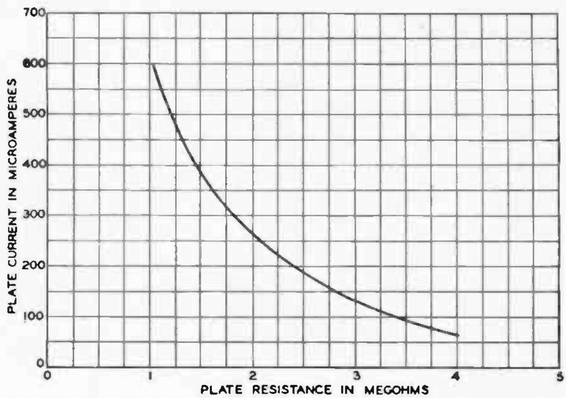


Fig. 9—Plate-Current, Plate-Resistance Characteristics of CX-322.

8 times $B$. If this is multiplied by the external impedance we obtain the voltage $_qe_p$, that is, applied to the grid of the next tube.

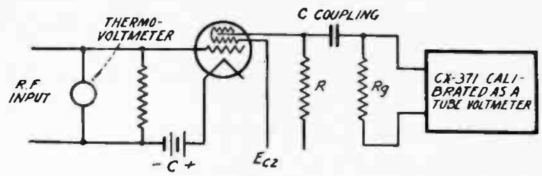


Fig. 10

Assume $B=0.35$ and $A=2\sqrt{2}$.

$$_qe_p=2\times B\times 64\times 10^{-6}\times 4\times 10^5=17.92 \text{ volts}$$

$$_{2q}e_p=B^2\times 64\times 4\times 10^5\times 10^{-6}=3.07 \text{ volts}$$

For $B=0.35$ and $A=\sqrt{2}$

$$_qe_p=2\times B\times 17\times 10^{-6}\times 4\times 10^5=4.26 \text{ volts}$$

$$_{2q}e_p=B^2\times 17\times 10^{-6}\times 4\times 10^5=0.81 \text{ volts}$$

To verify these values a CX-371 was calibrated as a tube voltmeter and the detector was worked directly into it as shown in Fig. 10. The same values of impedances were used as in the calculations. An r.f. choke was placed in series with the coupling condenser so that with no modulation the voltmeter read zero. When the carrier was modulated the tube voltmeter read the peak values of the output voltage which included ${}_qe_p$ and $2_qe_p$. The results for $B$ equal to forty per cent are given in Table I.

TABLE I

INPUT VOLTAGE TO CX-371 TUBE VOLTMETER

B = 40 per cent

| A | $\sqrt{2}$ | $2\sqrt{2}$ |
|---|---|---|
| Tube Voltmeter Reading | 6.0 | 20.8 |
| Calculated Voltage | 5.57 | 21.0 |

The external resistor was shorted with a condenser so as to change ${}_pZ$ and ${}_p\bar{Z}$. In both cases the voltage decreased very slightly. This again proves that the last two terms of the coefficients of $A^2$ in Eqs. (33), (34), and (35) are very small and may be neglected in the analysis. This could be expected from the considerations of the characteristics of the CX-322. The internal impedance is very high. The tube has some plate to filament capacity and this with the wiring capacity has a lower impedance than the internal impedance at radio frequencies. This impedance is in parallel with the load, so the resultant external impedance would be much less than the internal impedance. There would be very little gained by shorting the external impedance for radio frequencies when the tube is used as a detector.

If Eq. (33) is multiplied by $rp+qZ_1$, Eq. (34) by $rp+2qZ_1$, and (35) by $rp+R$ we get the equivalent voltages introduced in the plate circuit. These voltages may be used as $\mu eg$ in amplifier equations. In order to calculate the voltage on the grid of the next tube it is only necessary to solve for the voltage across the grid using the above equivalent voltages as in the regular amplifier equations.

In conclusion the writer wishes to thank Mr. R. M. Wise and Mr. D. F. Schmit for their helpful comments and suggestions in the preparation of this paper.

# THE SCREEN-GRID TUBE*

By

N. H. Williams

(University of Michigan, Ann Arbor, Michigan)

***Summary***—*Radio-frequency amplification by means of the three-electrode tube is usually disappointing. With resistance coupling the feed-back through the tube is in such a phase relation to the input voltage as to reduce the amplification below that given by the simple equation for voltage amplification.*

*With impedance coupling the feed-back causes self-oscillations when the circuit conditions are such as might be expected to give large amplification.*

*In the shielded grid tube the feed-back is reduced to a negligible amount and the current through the tube is very nearly independent of the plate voltage over the working range. Under these conditions, the voltage amplification becomes the product of the mutual conductance and the load impedance. High impedance in the plate circuit is obtained by using a sharply tuned parallel circuit. With proper shielding such a circuit may be used without producing self-oscillation. At a frequency of 700,000 cycles per second, amplifications of 80 fold per stage may be obtained.*

IN the case of resistance-coupled amplifiers, we may write the expression for the voltage amplification of one stage in the form $\mu R/R_0+R$, in which $\mu$ is the amplifying factor of the tube, $R$ the external or load resistance, and $R_0$ the resistance of the tube. In such a system the plate of one tube is connected through a grid condenser to the grid of the next tube. This expression for voltage amplification is based upon the assumption of ideal conditions which are not accurately realized. There are two reasons why this amplification cannot be attained. First: as the plate potential varies through a large amplitude, it induces a potential on the grid which is in phase with itself. But the plate potential differs from the potential of the positive battery terminal by the amount of the $RI$ drop in the load resistance and is therefore at its minimum value when the current through the tube is a maximum. This results in a feed-back through the tube to the grid in opposite phase to the actual grid voltage, thus reducing the effectiveness of the grid in controlling the current. Second: The equation indicates that the amplification may be increased by increasing the load-resistance indefinitely, the limit being the $\mu$ of the tube. As a matter of fact, the capacitance from

* Original Manuscript Received by the Institute, January 20, 1928. Delivered before the Detroit Section of the Institute, December 16, 1927.

plate to filament in the tube is in parallel with the load resistance and hence the impedance of the circuit is increased very little by adding resistance beyond a certain limit. It is for these reasons that resistance coupling for high-frequency amplification is usually disappointing.

With impedance coupling by means of a tuned circuit between plate and battery instead of a pure resistance, the case is very different. The tuned circuit is adjusted to behave as an inductive reactance and then the feed-back through the tube is so changed in phase that regeneration takes place and the amplification is greater than would be expected from the equation that is used in representing it. Furthermore, the capacitance from plate to filament is simply added to that of the condenser in the tuned circuit and is "tuned out"; thus no reduction of impedance results. The feed-back through the tube is so effective in this case as an agent of regeneration that the system breaks into self-oscillations and becomes useless as an amplifier if a sharply tuned circuit is employed to obtain impedance coupling.

The screen grid is a device to prevent the feed-back from plate to grid and thus prevent self-oscillation when the plate circuit is highly tuned. The second grid acts as an electrostatic screen between the plate and the control grid. It is held at a fixed potential and cuts off the control grid from the influence of the fluctuating potential of the plate. This tube for radio-frequency amplification was developed at the Research Laboratory of the General Electric Co. by A. W. Hull and N. H. Williams.[1]

Measurements show that the capacitance from plate to control grid is reduced to 1 per cent of that in the three-electrode tube when the latter is used without its socket. The percentage is much lower when the screen-grid tube is compared with the three-electrode tube as normally used. With these tubes it is possible to use sharply tuned circuits between plate and battery at almost any frequency without causing self oscillations. An amplification of 60 fold per stage is easily possible at a frequency of 700,000 cycles per second, and at lower frequencies still higher amplifications may be reached. An over-all amplification of more than two million-fold has been measured for a five-stage amplifier so built that each stage was in a separate compartment of a metal box.

[1] *Physical Review*, April, 1926.

The tube has a mutual conductance of about 0.4 milliampere per grid volt. Its plate-voltage, plate-current characteristic shows a downward slope where the plate voltage is less than the potential of the screen grid. Thus the tube has negative resistance in this region. This effect is the result of secondary emission from the plate. Each electron that strikes the plate may dislodge several electrons from the atoms of the metal, in which case the current in the plate circuit is reversed. When the plate voltage is above that of the screen grid, the current is positive, the characteristic is nearly straight and nearly parallel to the voltage axis. Here the tube resistance to alternating impulses is very high, which signifies that the plate current is practically independent of the plate voltage. This is an important item in connection with the behavior of the tube, for it accounts for an amplifying factor of over 200 and it eliminates one of the variables which, with other tubes, must be considered in connection with the circuit.

When the three-electrode tube is used in a radio circuit, the high-frequency component of the plate current is a function of seven different properties of the tube. These are:

(1) Rate of change of plate current with grid voltage.
(2) Rate of change of plate current with plate voltage.
(3) Rate of change of grid current with grid voltage.
(4) Rate of change of grid current with plate voltage.
(5) Capacitance between plate and grid.
(6) Capacitance between plate and filament.
(7) Capacitance between grid and filament.

As explained above, the second of this list is eliminated when the screen-grid tube is used because of the flatness of the characteristic curve. When the tube is used as an amplifier, the grid current may be kept at zero value by proper grid bias, and thus the third and fourth items are disposed of. In the screen-grid tube the capacitance from plate to grid and from plate to filament is negligible. The capacitance from grid to filament becomes a circuit constant since it is in parallel with another condenser. There remains, then, only the variation of plate current with grid voltage, i.e., mutual conductance $G$, and it appears that when this tube is used the mathematical solution of circuits is greatly simplified.

The amplifying factor, $\mu$, of the tube expresses how many times more effective an increment of voltage is if applied to the grid than if applied in the plate circuit. In mathematical symbols it is represented by the ratio of two differential coefficients.

$$\mu = \partial I_p/\partial E_g \div \partial I_p/\partial E_p$$

$$\partial I_p/\partial E_g = \quad G \quad = \text{Mutual conductance.}$$

$$\partial I_p/\partial E_p = 1/R_0 = \text{Plate conductance.}$$

$$R_0 = \text{tube resistance.}$$

From this it follows that $\mu = GR_0$.

Substituting $GR_0$ for $\mu$ in the equation with which we started, we have voltage amplification $= GR_0R/R+R_0$.

If $R_0$ is large as compared with $R$, we may neglect $R$ in the denominator and the expression for voltage amplification reduces to $GR$. In the case of impedance coupling, if $Z$ represents the impedance, it is apparent that $GZ$ becomes an approximate expression for the voltage amplification. With a parallel-tuned circuit, the impedance is given by the equation

$$Z = L/CR$$

$L$ being the inductance in the circuit, $C$ the capacitance, and $R$ the resistance. It is not difficult in a carefully tuned circuit to raise the impedance to 150,000 ohms and if $G$ is taken as 0.4 milliampere per grid volt, we have

$$\text{Voltage amplification} = GZ = 60.$$

This emphasizes the simplicity of the computations in the screen-grid amplifier.

The first piece of research that was undertaken making use of the shielded grid amplifier was the measurement of the charge of the electron by the shot effect.[2] This was done with an amplifier that was constructed at the Research Laboratory of the General Electric Co. Because of the fact that electricity is not an infinitely fine-grained fluid, but consists of discrete particles or electrons, the flow of the charge to the plate in a vacuum tube causes minute fluctuations of the plate potential due to the random distribution of the electrons in time. These minute fluctuations of voltage were measured with the screen-grid amplifier and from these measurements the charge of the electron was computed. The results agree within one per cent with those obtained by other and totally independent methods. Electrons produced by thermionic emission, by photoelectric emission, and by ionization of gases were measured in this way and now the measurement of the positive charge of ions is going on at the University of Michigan.

[2] Hull and Williams, *Physical Review*, February, 1925.

## Discussion on
# THE MEASUREMENT OF CHOKE COIL INDUCTANCE*
## (C. A. Wright and F. T. Bowditch)

**W. O. Osbon†**: The authors have given an interesting and valuable discussion of several simple methods of measuring, under operating conditions, the inductances of choke coils carrying direct current. They have made one or two statements, however, which the present writer desires to comment upon.

The inductance of a coil with a closed iron core is given by

$$L=\frac{0.4\pi N^2\mu A}{l}\times 10^{-8}$$

where $N$ = number of turns in winding
$A$ = area of core section, sq. cm.
$l$ = length of iron path, cm.
$\mu$ = effective permeability of the iron corresponding to conditions under which the choke is operating.

It is seen that all the factors entering into the above equation except the permeability are determined by the geometry of the coil, the permeability alone being affected by operating conditions. Hence, an investigation of the variations of $\mu$ under different conditions will lead to an understanding of the variations in inductance for a given coil.

Concerning the effective permeability of a core with an a-c. magnetizing force superposed on a d-c. mmf. the authors have made a serious error. They have stated that the inductance is "determined by the average slope of the saturation curve over the range within which the current varies," and have neglected hysteresis entirely. When an a-c. mmf. of maximum value $1/2\ \Delta H$ is superposed on the d-c. mmf. there will be produced a minor hysteresis loop[1] as at $a$ in the accompanying sketch. The total change in mmf., $\Delta H$, will produce a change in density, $\Delta B$, and the effective or incremental permeability[1] is

* Proc. I.R.E., 16, 3; March, 1928.

† Research Laboratory, Westinghouse Elec. & Mfg. Co., East Pittsburgh, Pa.

[1] T. Spooner, "Permeability" *Jour.* A.I.E.E. Vol. XLII, p. 42 and T. Spooner "Effect of Superposed Alternating Field on Apparent Magnetic Permeability and Hysteresis Loss", *Phys. Rev.* 1925.

$$\mu_\Delta = \frac{\Delta B}{\Delta H},$$

or it is the slope of the line drawn through the tips of the minor loop, which is seen to be quite different from the average slope of the normal curve in the range of variation of $H$. The minor loops at $b$ and $c$ indicate the state of affairs for values of d-c. mmf. which nearly saturate the core. It is seen that the incremental permeability decreases for increasing values of d-c. mmf. The writer has made tests on several kinds of iron which show that this statement holds even for very low values of d-c. magnetization in the range of upward curvature of the magneti-

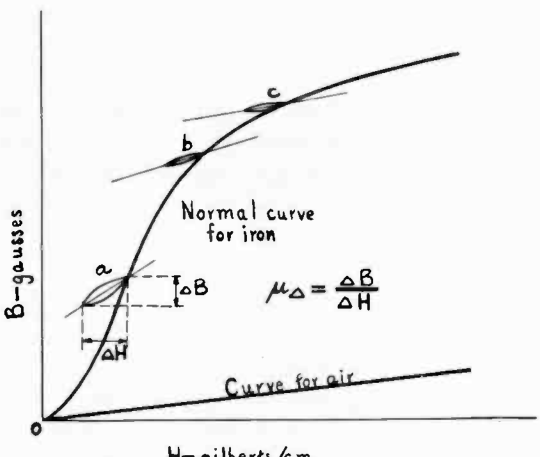


Fig. 1

zation curve. It seems likely to the writer, therfore, that the drop in inductance with no direct current shown in Fig. 6 in the paper was due to the presence of some residual magnetism in the core.

The value $\mu_\Delta$ corresponding to the particular values of a-c. and d-c. magnetization should be substituted for $\mu$ in the formula given above. In the second paper referred to above, Mr. Spooner gives a formula for calculating $\mu_\Delta$. When there is an air gap of length $a$ in the magnetic circuit the formula for inductance takes the form

$$L = \frac{0.4\pi N^2 A}{\dfrac{l}{\mu_\Delta} + a} \times 10^{-8}$$

In a paper by C. R. Hanna,[2] a direct method is developed for calculating the best value of air gap to be used in the magnetic circuit of a choke coil carrying direct current. In a recent paper in QST,[3] curves are given for calculating chokes according to the method developed by Hanna, for several different kinds of iron not considered in Hanna's paper.

In the latter part of their paper, the authors commit a serious error by stating that the slope of the magnetization curve for air is considerably greater than the slope of the magnetization curve for iron in the region of saturation. The truth of the matter is that the slope of the iron curve *approaches* the slope of the curve for air. In other words, the incremental permeability of iron is never less than unity as the statement of the authors would lead one to believe.

**C. A. Wright**†: The authors are familiar with the minor loops referred to by Mr. Osbon and have referred to them in a previous discussion.[1] However, it was decided that in the present case they were of negligible importance and that a consideration of them would unnecessarily complicate the methods of inductance measurement suggested.

The present paper was intended to apply to choke coils used in radio "B" power units, the inductance of which was the subject of much interest at the time the measurements were made.

In the cases of a representative lot of commercial samples measured the inductances were practically the same for increasing and decreasing values of direct current. With the negligible hysteresis thus indicated, the slopes of the minor loops approached closely the slopes of the saturation curve corresponding to the positions of the minor loops, and under these conditions the statements made regarding the value of inductance are correct.

The criticism of Mr. Osbon, that "the slope of the magnetization curve for air is considerably greater than the slope of the curve for iron in the region of saturation," is probably based on the assumption that the abscissas of the curves are measured in

[2] C. R. Hanna, "Design of Reactances and Transformers which Carry Direct Current," *Jour.* A.I.E.E., Feb. 1927, page 128.

[3] D. E. Replogle, "Notes on the Design of Iron-Core Reactances which Carry Direct Current," QST, Apr. 1928, page 23.

† National Carbon Co., Inc., Cleveland, Ohio.

[1] "Telephone Communication," by Wright and Puchstein, Chap. 7, pg. 141–142. Published by McGraw-Hill Book Company, 1925.

the usual magnetizing force per unit length. However, it is to be noted that these abscissas are in "ampere turns" so that the curve for the iron referred to is that of the total length of the iron magnetic circuit, and the curve for air is that of the total length of the air gap. If this air gap is very small, as in the actual case, this curve for air may be more steep, as is stated in the discussion, than that for the curve of the iron portion of the circuit. The "proportional addition" of the curves for iron and air will then have the stated effect on the total curve.

## Discussion on
# ON THE DISTORTIONLESS RECEPTION OF A MODULATED WAVE AND ITS RELATION TO SELECTIVITY
## (F. K. Vreeland)*

V. D. Landon†: The paper by Frederick K. Vreeland, "On the Distortionless Reception of a Modulated Wave and its Relation to Selectivity," was read with interest at the Westinghouse Electric and Manufacturing Company laboratories. A good deal of work has been done here on band-pass filters somewhat similar in design to that described by Dr. Vreeland. It seems desirable to amplify his discussion considerably.

Of course, it will be realized that the scheme outlined is by no means new either with identical or spaced tuning. The circuit shown is merely two conductively coupled tuned circuits with critical coupling or slightly greater than critical coupling. The same principle has been incorporated in commercial receivers for years, one such use being in a superheterodyne provided with intermediate frequency transformers having tuned primary and tuned secondary windings, the coupling between which is quite loose, being only slightly greater than critical coupling. This, of course, gives the well-known response curve characteristic of such an arrangement in accordance with well understood principles. A treatment of coupled circuits that describes the phenomena very well is given in Morecrofts "Principles of Radio Communication" on page 100. The dotted curve of Fig. 99 illustrates the band pass action perfectly. The precise method of coupling employed is of course immaterial.

The idea of tuning both circuits over a considerable frequency band is not novel either. The old "loose couplers" and similar devices were in use before radio broadcasting was known. These receivers employed tuned primary and tuned secondary, and the coupling was adjustable. In general, something approximating critical coupling was striven for, as this gave the best sensitivity retaining good selectivity.

*Presented at the Annual Convention of the Institute of Radio Engineers, January 9, 1928. Proc. I.R.E., 16, 3; March, 1928.

† Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania.

It is of interest to note the behavior of two such coupled circuits when they are tuned to various frequencies throughout the broadcast range. It is well-known that for critical coupling the mutual reactance should equal the square root of the product of the primary and secondary resistances. Or for duplicate circuits the mutual reactance should equal the resistance of one of the circuits. If greater coupling than this is used two peaks will occur in the resonance curve, the greater the coupling the more pronounced the peaks. For coupling only slightly greater than critical a nearly flat top curve, with steep sides, is obtained. For less than critical coupling the curve resembles an ordinary resonance curve, though a more favorable ratio will be obtained of the width near the top to the width near the bottom. A value of coupling just greater than critical gives the best shape to the resonance curve.

It is a well-known fact that the resistance of a coil varies rapidly with frequency. Common types of coils used in broadcast receiver design vary in resistance at a rate lying between the first and second power of the frequency. If a mutual inductive reactance is used to couple the circuits, this reactance will vary in value exactly as the first power of the frequency. If coils could be found whose resistance varied exactly as the first power of the frequency then the value of coupling could be made correct to give the flat top curve all over the range. The fact that the resistance varies faster than the frequency means that the curve will either have a rounded top at high frequencies (corresponding to less than critical coupling) or a fairly pronounced double peak at low frequencies (corresponding to more than critical coupling).

It is to be noticed, however, that even assuming that the resistance varies in the correct way to give the same curve shape at all frequencies, the width of the curve will not remain constant over the range. The width will be directly proportional to the frequency. That is, the band received will be two and one-half times as wide at the high-frequency end of the scale as it is at the low-frequency end.

Also if the resistance varies as suggested, the signal will be amplified a great deal more at the high frequencies than at the low, since the tuned impedance of the circuit will be directly proportional to the frequency.

If the rate of change of resistance lies between the first and second power of the frequency the selectivity variation over the range will be even worse than that indicated. The amplification will be more nearly uniform but will still be greatest at high frequencies.

For these reasons it is highly desirable that the coupling increase at a rate slightly less than as the first power of the frequency. This results in decreasing the amplification at high frequencies to the same value that is obtained at low. At the same time the selectivity is made considerably better. Of course, the desirable square top feature is sacrificed to a certain extent but this does not matter as the curve is already far too broad at this end of the scale.

It was found possible to obtain the desired variation in coupling with frequency by employing combinations of capacitive and inductive coupling in a variety of ways.

If this variation is correctly done the resonance curve will be found only slightly wider at high frequencies than at low. At low frequencies the curve had a very slight double peak and at mid-scale an ideal flat top. At high frequencies the curve was quite rounded on top though not nearly so sharp-pointed and sloping-sided as the curve of the same circuits not coupled, operated in cascade.

It is believed that such an arrangement is more nearly ideal than that discussed by Dr. Vreeland.

## BOOK REVIEW

Experimental Radio, BY R. R. RAMSEY. Ramsey Publishing Company, Bloomington, Indiana. Third Edition. 229 pages. $2.75.

This compendium of radio experiments was first issued by the author for use in his radio course at Indiana University, where he is Professor of Physics. The first edition appeared in 1923 after about five years of collection of experiments from various sources, including textbooks on radio telegraphy and telephony. The author states that the course is intended to be about on a level with a good second-year college course in physics. No attempt is made to cover the more complicated measurements, but some references to these are included. In all there are 117 experiments. Typical ones are as follows: "To Measure the Dielectric Constant of Oil"; "To Calibrate a Wavemeter Using Overtones"; "Measurement of the Amplification Constant of a Tube"; "To Measure the Resistance of a Radio Circuit; Resistance Variation Method"; "Field Intensity Measurements; Radiation from a Coil."

Within its scope, which the author has clearly defined, the book is an excellent manual of radio- and audio-frequency measurements and demonstrations. While all the material in it should be familiar to radio and electrical engineering graduates, experienced engineers will find Ramsey's outline useful for refreshing their memories on specific points with which their work has not brought them into contact for some time. The only general criticism of the book which might be made is to point out the absurd use of the terms "AC current," "DC current," etc., which are employed with a consistency worthy of a more logical nomenclature. CARL DREHER†

† Staff Engineer, National Broadcasting Company, New York City

# GEOGRAPHICAL LOCATION OF MEMBERS ELECTED

## May 2, 1928

**Transferred to the Fellow grade**

| | | |
|---|---|---|
| Minnesota | Minneapolis, University of Minnesota | Jansky, C. M. Jr. |

**Transferred to the Member grade**

| | | |
|---|---|---|
| Illinois | Chicago, 1018 North Hamlin Avenue | Howard, C. M. |
| Michigan | Berrien Springs | Fetzer, John E. |
| New York | New York City, 64 Broad St., Inter. Stand. Elec. Co. | Montcalm, S. R. |
| Pennsylvania | Wilkinsburg, 706 Lamar Street | Brindley, William E. |
| | Wilkinsburg, 519 North Avenue | Horn, C. W. |

**Elected to the Member grade**

| | | |
|---|---|---|
| Dist. of Columbia | Takoma Park, 19 Woodland Avenue | Trogner, Arthur M. |
| Illinois | Oak Park, 551 Bellefonte Avenue | Pearson, Oscar A. |
| New Jersey | East Orange, 30 North Arlington Avenue | Ackerman, Rudolph W. |
| New York | New York City, 40 Rector Street, Room 618 | Chesterton, A. J. |
| | New York City, 66 Broad Street, R. C. A. | Morton, Alfred H. |
| Pennsylvania | East Pittsburgh, Westinghouse Elec. & Mfg. Co. | Hanna, Clinton R. |

**Elected to the Associate grade**

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| | El Segundo, P. O. Box 531 | Hart, William J. |
| | Inglewood, 579 Ellis Avenue | Underwood, Ernest G. |
| | Los Angeles, 580 Crane Blvd. | Power, Ralph L. |
| Colorado | Seibert | Fingado, Francis H. |
| Connecticut | Bridgeport, 246 West Avenue | Dulbis, Geddie E. |
| | Bridgeport, P. O. Box 547 | Dunning, S. |
| | Middletown, Wesleyan Station, Box 283 | Copeland, Herman A. |
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| | Washington, 2016–34th Street, N. W. | Mims, Clinton F. |
| | Washington, 1536 Yon Street, N. W. | Smith, David H. |
| | Washington, c/o National Radio Institute | Sprayberry, Frank L. |
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| | Chicago, 617 N. Lorel Avenue | Denise, Garet W. |
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| | Chicago, 118 South Lincoln Street | Galvez, Manuel E. |
| | Chicago, 427 South La Salle Street, Room 311 | Newlander, Ralph A. |
| | Chicago, 100 W. Monroe St., c/o R. C. A. | Rothenberger, William |
| | Chicago, 300 South Racine Avenue | Van Benschoten, Walter |
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| | Boston, 3 Milford Street | Parnagian, Aram |
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| | Dennis Port | Skinner, Oscar E. |
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| New Jersey | Bloomfield, 230 Grove Street | Hanson, Ernest R. |
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| | Newark, 487 Summer Avenue | O'Dowd, Thomas E. |

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| | Buffalo, 1415 Michigan Avenue | Lewis, John C. P. |
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| | Buffalo, 840 Marine Trust Bldg | Waud, E. C. |
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| | New York City, 177 Brook Avenue | Armus, Harry |
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| | Rochester, 338 East Main Street | Waldorf, E. T. |
| | Rochester, 55 Tacoma Street | Wishart, William Ross |
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| | Cleveland Heights, 2175 Middlefield Road | Williams, Hollie |
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| Texas | Beaumont, 1495 Johns Avenue | Wengel, Raymond W. |
| | Jacksboro, P. O. Box 213 | Isbell, G. Terrill |
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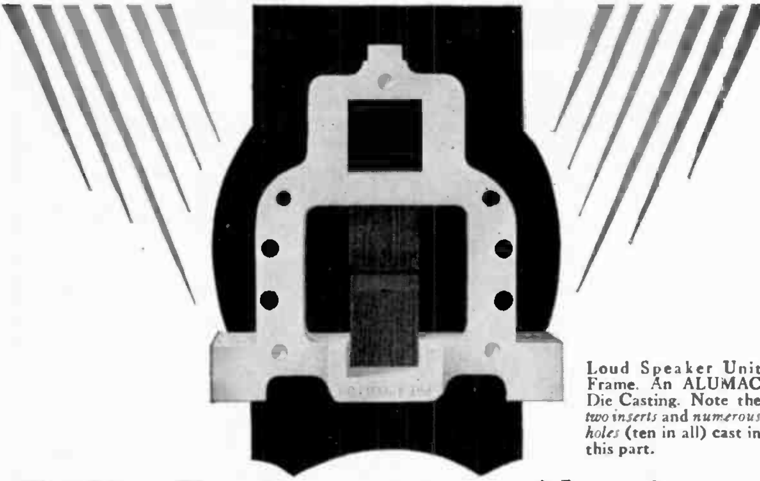

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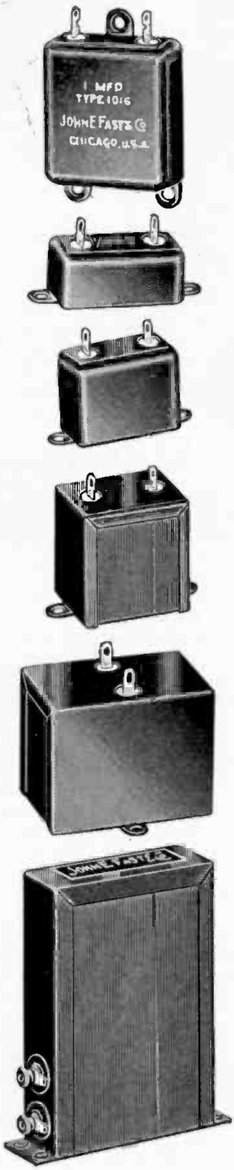


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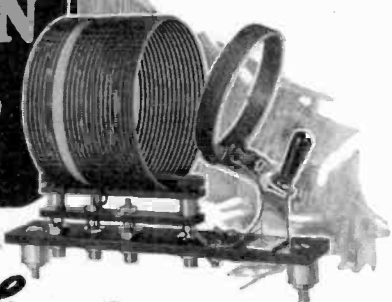

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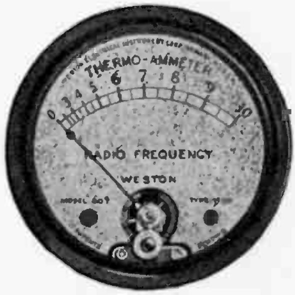


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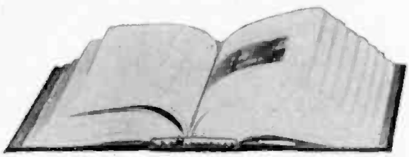

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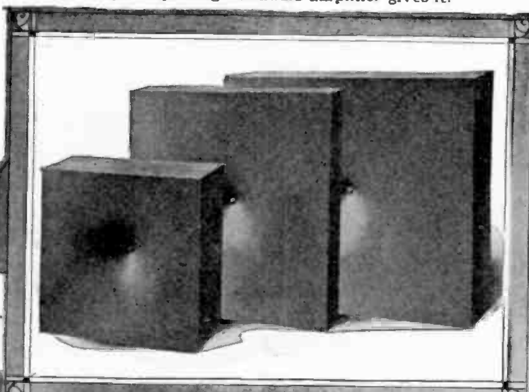

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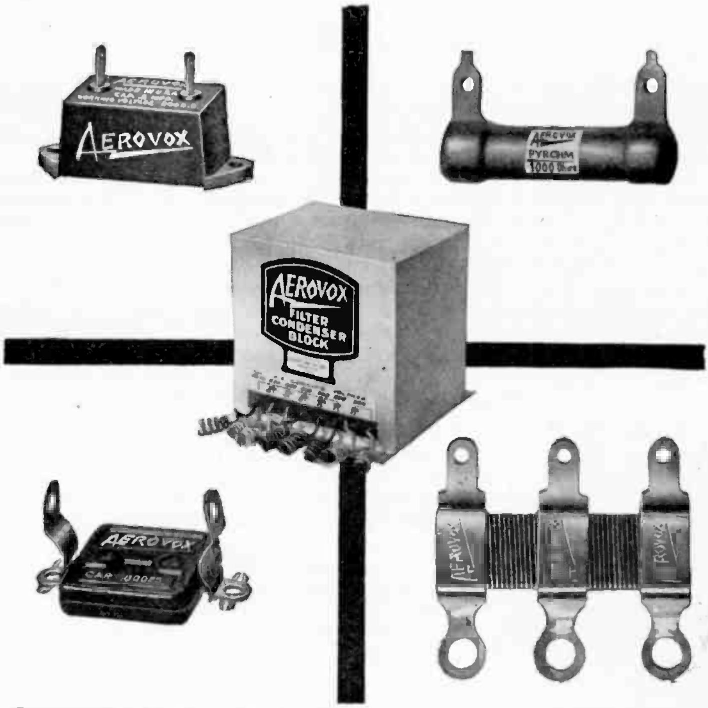


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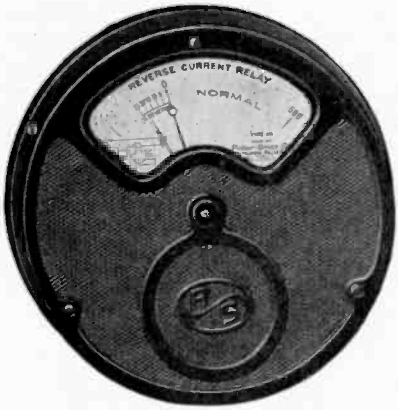

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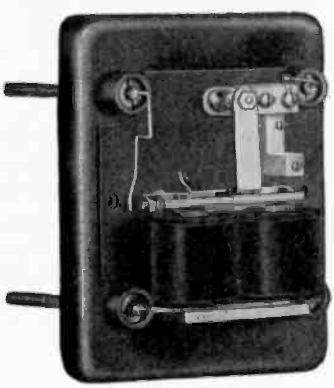
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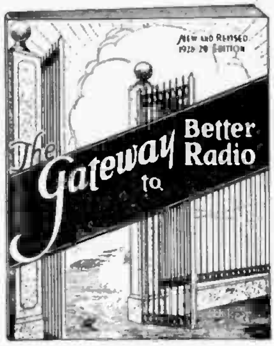


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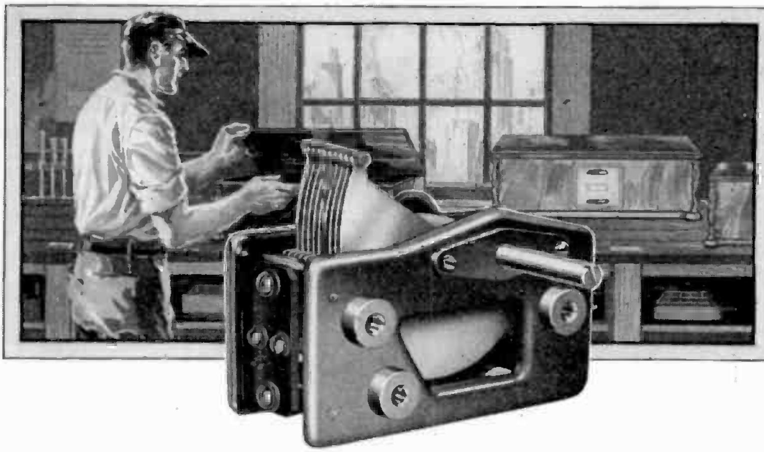

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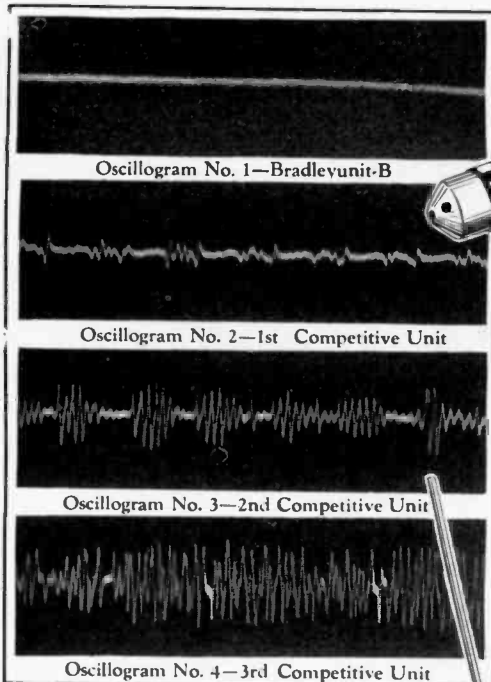

Oscillogram No. 1—Bradleyunit-B

Oscillogram No. 2—1st Competitive Unit

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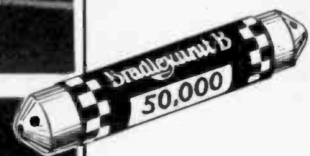



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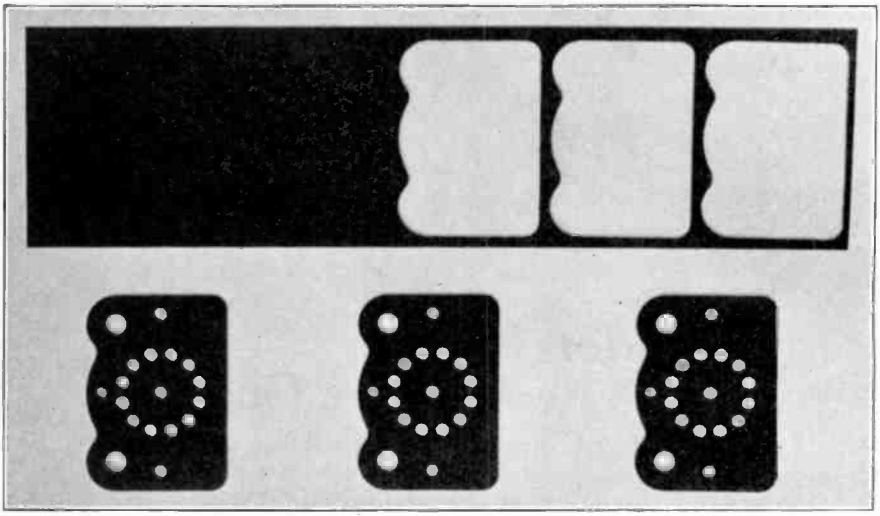


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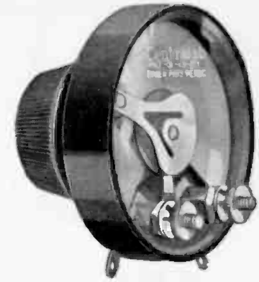

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THE volume control of a radio set is one of the parts most used and subjected to the most wear. Care must be taken to choose the type that will give longest, trouble-free service—a type that will not introduce noise to interfere with the quality of reception after a short period of service.

Centralab Volume Controls have a patented rocking disc contact that eliminates all wear on the resistance material. This feature adds to the smoothness of operation in that a spring pressure arm rides smoothly on the disc and NOT on the resistance. The bushing and shaft are thoroughly insulated from the current carrying parts. This simplifies mounting on metal panel or sub base and eliminates any hand capacity when the volume control is in a critical circuit. Full variation of resistance is obtained in a SINGLE TURN of the knob.

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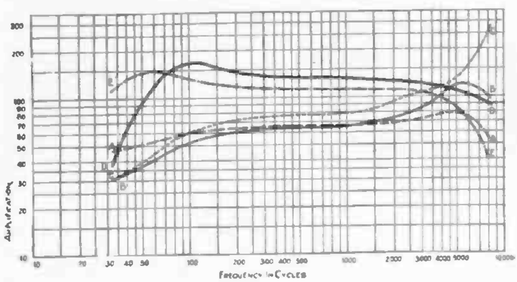
## Public Address Amplifier
## Short Wave Kit

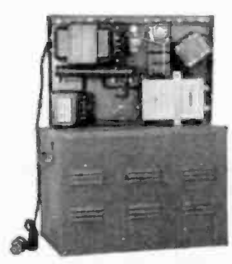


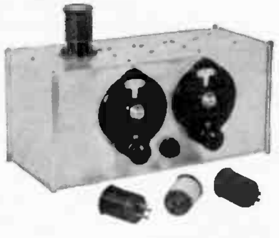
S-M 225 Audio Transformer—See Curve E

SILVER-MARSHALL has ready new audio transformers giving double the amplification of the best existing types, a far better frequency characteristic, and practical elimination of hysteretic distortion. Curve "D" above illustrates the performance of a pair of the new small S-M 255 and 256 transformers listing at $6.00 each, as compared to three pairs of $8.00 to $10.00 transformers on the open market. It tells the story of doubled amplification and improved bass amplification. Curve E is for a pair of S-M 225 and 226 transformers, listing at $9.00 each. These new transformers are available to manufacturers at surprisingly low prices. At the June R. M. A. trade show S-M will demonstrate a comparison amplifier which audibly proves the superiority of the new $6.00 transformers over any standard existing types.

## *Type 685 Public Address Unipac*

For high quality high volume coverage of crowds of 2,000 to 10,000 people with one to twelve speakers, as in theaters, churches, auditoriums, or out-of-doors, the new S-M Unipac is ideal. It is a self-contained light-socket power amplifier for radio, record pick-up or microphone input, using one UY227, one UX226, one UX250, and two UX281 Radiotrons. Type 685 Unipac, factory wired and assembled, is priced at $160.00, less tubes and accessories, or, in kit form, ready to assemble, $124.00.





## *Screen Grid Short Wave Kits*

The new S-M "Round the World Four" and "Round the World Adapter" are two non-radiating short wave receiver kits housed in aluminum shielding cabinets, and tuning from 17.4 to 204 meters with four plug-in coils. They are the first sets to employ R. F. amplification at short waves with but one tuning dial. Price, 730 four tube kit, complete with cabinet $51.00; 731 two tube adapter $36.00; 732 essential kit containing coils, condensers and R. F. chokes, $16.50.

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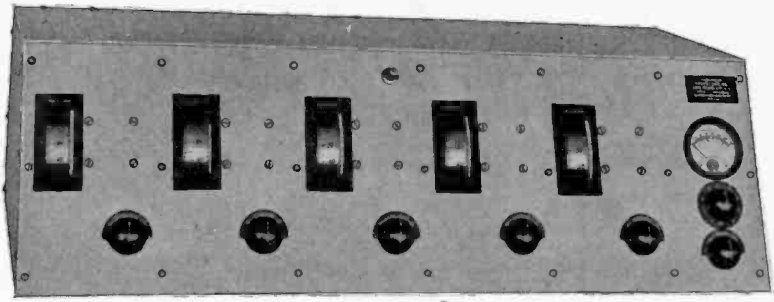
# Universal Transoceanic

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Each Transoceanic is made to special order. The most advanced Transoceanic uses four 222 Screened Grid Tubes in the radio amplifier, 200A detector and a four stage audio amplifier. Two of the four stages are power amplifiers in a push pull system using either 2-210 or 2-250 power tubes. Power units are available to supply 450 volts of B current and the necessary C and A currents, giving complete electric operation.

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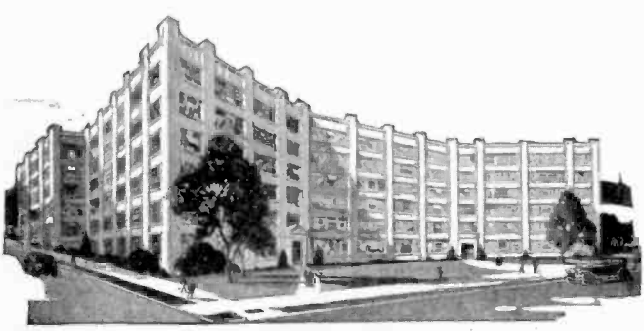

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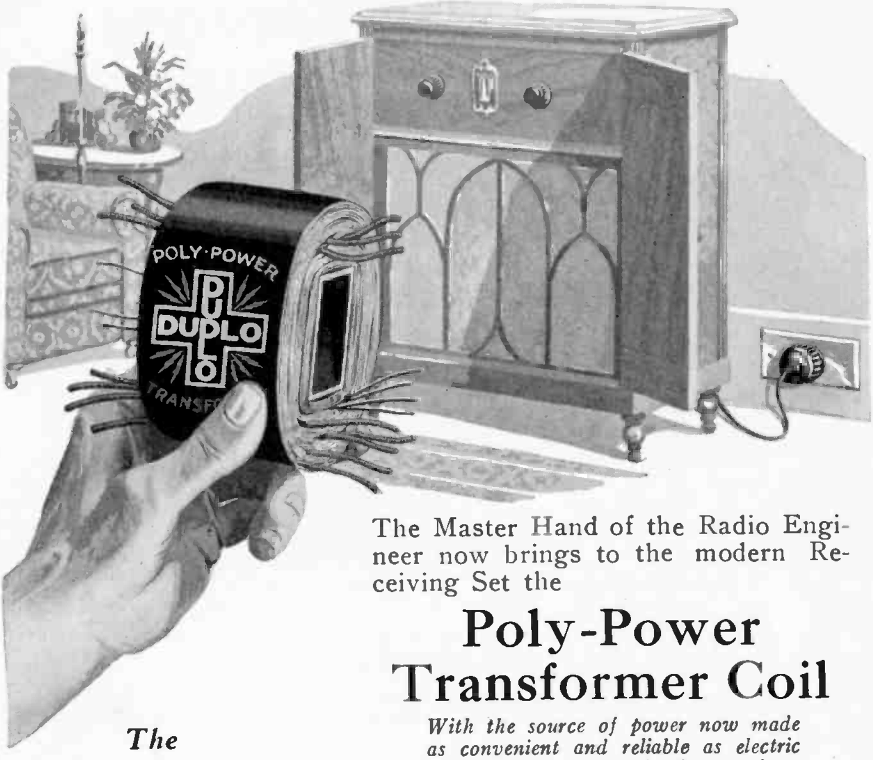


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So again Dudlo meets the needs of the Radio manufacturer by preparing far in advance for volume production of these, the most intricate of all Radio coils. Special machinery has been designed and installed in anticipation of the greatest season the industry has ever known.

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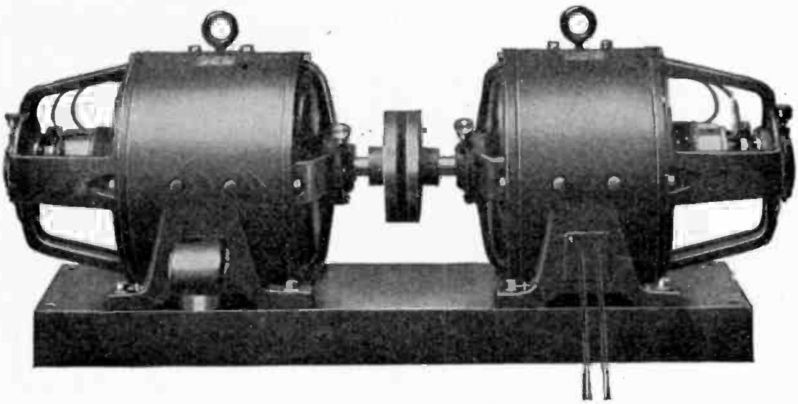

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*No. 199*
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*Analyzer*

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AFTER you have seen the new Jewell Pattern 199 A.C. and D.C. Radio Set Analyzer you will agree that it has desirable features not found in similar instruments now available—features that improve and increase its value in the servicing of radio sets and equipment.

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Some of the features which mark it as distinctly advanced in design are: a new 5-prong plug arrangement, simple push button switches for making tests, provision for an accurate tube test, a new cathode voltage test—all of which are distinct Jewell accomplishments and worthy contributions to the advancement of radio.

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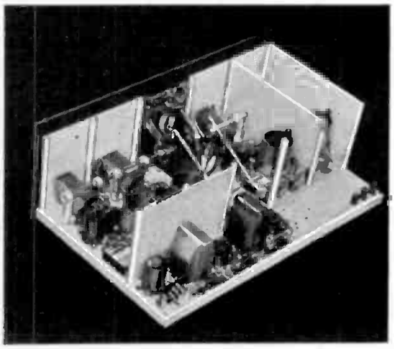

*The famous Hammarlund-Roberts Hi-Q Six employs parts produced by ten of America's finest radio manufacturers, including* ACME PARVOLT *By-Pass Condensers.*

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THE rapidly increasing use of by-pass and filter condensers in modern A.C. operated circuits demands the greatest caution against poor quality, inaccuracy of rating and non-uniformity in condensers.

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It is vital to use condensers whose ratings are based upon CONTINUOUS DUTY.

It is vital to use condensers that are made with AT LEAST the overload factor of safety demanded by the R.M.A. and N.E.M.A.

Condensers which possess these qualities not only aid quality reception, but overcome the possibility of breakdown and heavy losses from ruined tubes, transformers and other parts. . . . .

All ACME PARVOLT Condensers are made and tested to standards of the R.M.A. and N.E.M.A. They are used and recommended by leading radio engineers, designers, service men and custom-builders everywhere. Play safe with PARVOLTS! Made by THE ACME WIRE CO., New Haven, Conn., manufacturers of magnet and enameled wire varnished insulations, coil windings, insulated tubing and radio cables.

ACME PARVOLT FILTER CONDENSERS are supplied in all standard mfd. capacities for 200, 400, 600, 1000, and 1500 Volt D.C. requirements. Uniform height and width for easy stacking. Supplied singly or in complete housed blocks for the important power supply units such as Thordarson, Samson and others.

ACME PARVOLT BY-PASS CONDENSERS are supplied in all standard mfd. capacities and for all required voltages.

# ACME PARVOLT CONDENSERS

*Made by the Manufacturers of*

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For all types of radio wiring. High insulation value; non-inflammable, 10 colors.

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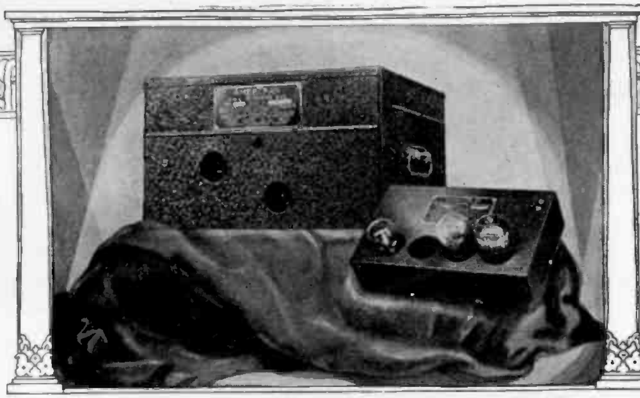
A superior cambric tubing for all practical radio and other electrical requirements. Supplied in 10 colors.



# *Are you Trying to Drive Your Set with a* PIGMY *Motor?*

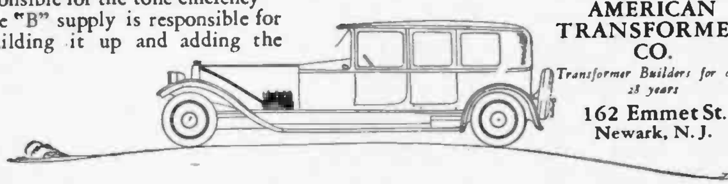
THE cabinet may be large and imposing. It may be made from the finest woods, and yet your set may be compared with a luxurious car with a 156 inch wheelbase and driven by a 20 horsepower motor. The audio amplifier may be compared with the ignition; the "B" supply to the motor in your automobile.

A motor car is judged by its speed; its hill climbing record; its pulling power in sand and mud; its ability to "stand-up" delivering constant smooth power for thousands of miles.

A radio set today is judged by its ability to reproduce the music *as it is played*, not "radioed music" with the inevitable loss of tones and overtones both in the treble and the bass. The audio amplifier is responsible for the tone efficiency—the "B" supply is responsible for building it up and adding the timbre which means the difference between mediocre and natural music.

Your tuner may be the best on the market—give it a chance to deliver the fine music of which it is capable. Replace your audio amplifier with the AmerTran Push-Pull Amplifier. Replace your "B" Batteries or "B" supply with the AmerTran ABC Hi Power Box which supplies 500 volts DC plate current, and sufficient AC filament current to transform any set from DC to AC operation. Of course it may be used with the new AC sets.

Use a power plant worthy of your set. The difference in tone efficiency is so great that words cannot describe it. It is the difference between a dull, cloud filled sky and the brilliance of bright sunlight.



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*Transformer Builders for over 28 years*

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Let your ears tell you the difference. You owe it to yourself to hear what is possible in Radio.

*Both units licensed under patents owned or controlled by RCA, and may be bought with tubes.*

Prices: — Power Amplifier $60.00
A B C Hi Power Box $95.00
Both without tubes
Slightly higher west of the Rockies

AMERICAN TRANSFORMER CO.
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☐ Kindly send complete information on the AmerTran Push-Pull Amplifier and A B C Hi-Power Box.

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Hundreds of invisible beacons planted along the Channel coast from Plymouth to Carlisle suddenly flashed into flame upon theapproach of the Spanish Armada, arousing England to the valiant defense that resulted in the destruction of this supposedly invincible fleet.

*Invisible Value*—is an intangible quality of the Grebe Synchrophase A-C Six. It is an inbuilt quality that cannot be copied. It is the basis upon which for nineteen years all Grebe radios have been made and sold.

The Grebe standards of craftsmanship and materials, the real secret of this *invisible value*, never have been more clearly typified than in the Grebe Synchrophase A-C Six. Its remarkable selectivity, fidelity of tone and distinctive Grebe innovations have brought immediate success. It offers a complete demonstration of the difference between a radio, built just to meet price conditions, and one whose quality must be upheld to maintain reputation.

Also Grebe Synchrophase Seven; Grebe Synchrophase Five; Grebe Natural Speaker; Grebe No. 1750 Speaker.

Hear the Grebe Synchrophase A-C Six today, or send for Booklet I, which fully explains this new receiver.

GREBE SYNCHROPHASE

*A-C Six*

RADIO

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*Makers of quality radio since 1909*

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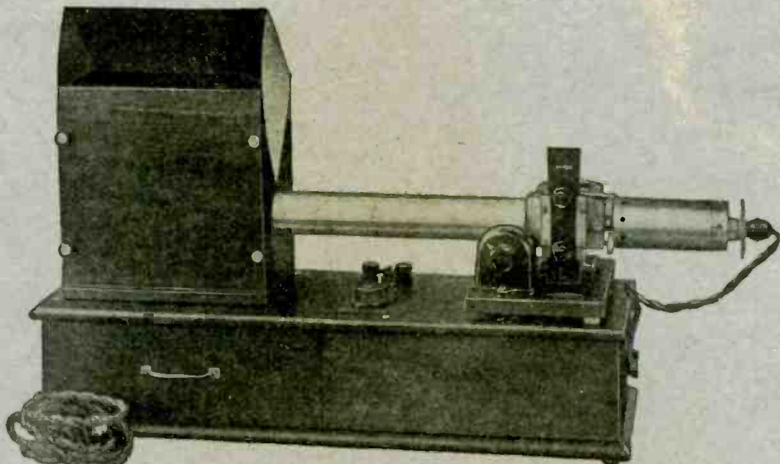
# String Oscillograph



*Type 338 String Oscillograph*

A portable instrument of great sensitivity for visual observation of frequencies up to 3000 cycles.

Operates from 110 volt 60 cycle alternating current.

As a vibration galvanometer the type 338 is very useful in bridge work at low frequencies.

Fully described in Bulletin 8100-I.

Type 338 Oscillograph, with carrying case.......$200.00

Type 338 Galvanometer equipment only,
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