VOLUME 22 JUNE, 1934 NUMBER 6

PROCEEDINGS
of
The Institute of Radio Engineers

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Institute of Radio Engineers Forthcoming Meetings

DETROIT SECTION
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LOS ANGELES SECTION
June 19, 1934

WASHINGTON SECTION
June 11, 1934

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 22 June, 1934 Number 6

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The Institute of Radio Engineers

GENERAL INFORMATION

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

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

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

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	Peoria, Galena Rd.	Karl, R. J.
Indiana	Richmond, 411 Kinsey St.	Timmons, F. E.
Massachusetts	Cambridge, Rm. 1–131, Mass. Inst. of Tech.	King, H. F.
Minnesota	St. Paul, 1946 Portland Ave.	Milinowski, A. S., Jr.
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	West Chester, 202 N. Penn St.	Comfort, H. F.
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Canada	Edmonton, Alta., 400 C.P.R. Bldg.	Mitchell, A. M.
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Elected to the Student Grade

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Kansas	Lawrence, 700 Alabama	Erickson, L. H.
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New Mexico	Albuquerque, 114 S. Maple St.	Remley, H.
Ohio	Cleveland Heights, 3325 Beechwood Ave.	Hill, J. S.
Pennsylvania	Bethlehem, 514 Delaware Ave.	Porter, R. S.
China	Shanghai, c/o Shanghai Radio Central Station, 565 Min Kou Rd.	Hsu, A. L.
Germany	Baden, Studenthaus, Karlsruhe	Keinonen, A. F.
Puerto Rico	Guayanilla, c/o Central Rufina	Valls, R. P.

APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Admissions Committee. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before June 30, 1934. Final action will be taken on these applications July 2, 1934.

For Election to the Associate Grade

Illinois	Chicago, 906 Wrightwood Ave.	Lair, J. R.
Louisiana	New Orleans, 3203 Upperline St.	Young, J. L. H.
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	Cambridge, 5 Haskell St.	Anderson, W. C.
	Chicopee, 68 Moore St.	Brodacki, P. F.
	Chisopee Falls, 243 Grove St.	Atwood, B. E.
Missouri	Kansas City, 305 N. Gladstone Ave.	Loukota, D. H.
New Jersey	Montclair, 314 Linden Ave.	Stanton, S. W.
	Jersey City, 361 Ogden Ave.	Swensen, J.
New York	Brooklyn, 623-80th St.	Danielsen, T.
	Brooklyn, 1069 Halsey St.	Evans, F. M.
	Buffalo, Buffalo Broadcasting Co., Rand Bldg.	Hoffman, K. B.
	Forest Hills, L.I., 6804 Burns St.	Gilbert, J. D.
	New York City, 2825 Webb Ave.	Engelien, J.
	New York City, 310 W. 99th St.	Haeff, A. V.
	New York City, 17 W. 60th St.	Stansfield, J. Q.
Ohio	Cincinnati, Glenwood Apts., College Hill	Crosley, P., III
	Cincinnati, 6822 Bantry Ave.	Stephan, H. W.
Pennsylvania	Philadelphia, 228 Chancellor St.	Gilmer, J. E.
Washington	Bellingham, 408 N. Forest St.	Waters, T. R., Jr.
Wisconsin	Madison, 321 S. Henry St.	Bell, A. L.
England	Hendon N.W. 4, London, 32 Hall Lane, Watford Way	Thompson, J. L.
	London N.W. 10, 23 Phillimore Gardens	Rich, D.
	London W. 14, 34 Dewhurst Rd.	Sutherland-Read, J.
	Mutley, Plymouth, 21 Connaught Ave.	Hughes, R. W.
India	Bihar, Domchanch, via Kodarma, E.I. Ry.	Sastry, S. S.
New Zealand	Auckland, Wilson St., Ellerslie	Parsons, H. E.
Uruguay	Montevideo, Nueva York 1590	Primavesi, J. C.

For Election to the Junior Grade

Illinois	Mooseheart, Lake Camp Hall	Masteryanni, T.
Texas	Corsicana, 2000 W. 3rd Ave.	Tripp, B. E.
England	Bradford, Yorks., 440 Harewood St.	Harrison, T.
	Shipley, Yorks., 10 Wellcroft	Hardy, R. A.
	Shipley, Yorks., 3 Nabwood Bank	Ledger, H.
	Shipley, Yorks., 52 Saltaire Rd.	Wade, F.

For Election to the Student Grade

California	Alameda, 2148 Lincoln Ave.	Duguid, H. Q.
	Stanford University, Box 2173	Blanchard, H. P.
Iowa	Iowa City, 1025 E. Washington St.	Huston, O. H.
Michigan	Wyandotte, 602 Chestnut St.	Maddock, B. H.
Washington	Seattle, 1415½ E. North Lake	Trowbridge, H. M.
Canada	Westmount, P.Q., 619 Belmont Ave.	Mellor, A. G.
Puerto Rico	San German, Box 243	Sanchez, R.

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INSTITUTE NEWS AND RADIO NOTES

Election Notice

ARTICLE VII

NOMINATION AND ELECTION OF PRESIDENT, VICE PRESIDENT, AND THREE DIRECTORS AND APPOINTMENT OF SECRETARY, TREASURER, AND FIVE DIRECTORS

Sec. 1—On or before July 1st of each year the Board of Directors shall call for nominations by petition and shall at the same time submit to qualified voters a list of the Board's nominations containing at least two names for each elective office, together with a copy of this article.

Nominations by petition shall be made by letter to the Board of Directors setting forth the name of the proposed candidate and the office for which it is desired he be nominated. For acceptance a letter of petition must reach the executive office before August 15th of any year, and shall be signed by at least thirty-five Fellows, Members, or Associates.

Each proposed nominee shall be consulted and if he so requests his name shall be withdrawn. The names of proposed nominees who are not eligible under the Constitution, as to grade of membership or otherwise, shall be withdrawn by the Board.

On or before September 15th, the Board of Directors shall submit to the Fellows, Members, and Associates in good standing as of September 1st, a list of nominees for the offices of President, Vice President, and three Directors. This list shall comprise at least two names for each office, the names being arranged in alphabetical order and shall be without indication as to whether the nominees were proposed by the Board or by petition. The ballot shall carry a statement to the effect that the order of the names is alphabetical for convenience only and indicates no preference.

Fellows, Members, and Associates shall vote for the officers whose names appear on the list of nominees, by written ballots in plain sealed envelopes, enclosed within mailing envelopes marked "Ballot" and bearing the member's written signature. No ballots within unsigned outer envelopes shall be counted. No votes by proxy shall be counted. Only ballots arriving at the executive office prior to October 25th shall be counted. Ballots shall be checked, opened, and counted under the supervision of a Committee of Tellers, between October 25th and the first Wednesday of November. The result of the count shall be reported to the Board of Directors at its first meeting in November and the nominees for President and Vice President and the three nominees for Directors receiving the greatest number of votes shall be declared elected. In the event of a tie vote the Board shall choose by lot between the nominees involved.

Sec. 2—The Treasurer, Secretary, and five appointive Directors shall be appointed by the Board of Directors at its annual meeting for a term of one year or until their successors be appointed.

In accordance with the above, notice is hereby given that the following names have been placed in nomination to be balloted on in 1934.

FOR PRESIDENT

Stuart Ballantine — William Wilson

FOR VICE PRESIDENT

Heinrich Barkhausen — P. O. Pedersen

FOR DIRECTORS

L. C. F. Horle	E. L. Nelson
Haraden Pratt	B. E. Shackelford
H. A. Wheeler	L. E. Whittemore

Joint Meeting with International Scientific Radio Union

A joint meeting of the Institute was held with the American Section of the International Scientific Radio Union in the National Academy of Sciences Building, Washington, D.C., on April 27, 1934. The morning session was presided over by Dr. Dellinger and the afternoon session was under the direction of President Jansky. Twenty-one papers were presented and abstracts of them follow.

THE DEVELOPMENT AND CHARACTERISTICS OF NINE-CM RADIATION

G. R. KILGORE
(Westinghouse Electric & Manufacturing Company)

ABSTRACT

The application of "Micro-Rays" or radio-optical waves less than one meter in length to transmission of intelligence is continuously gaining in importance because of the convenience with which such waves can be focused in narrow beams by means of properly shaped reflectors. The shorter the waves used, the easier the design of an efficient optical system; but at the same time the more difficult is the generation of oscillations of proper frequency. Magnetostatic tubes with suitable circuits seem at present to be the most efficient generators of radio-optical waves from 50 down to one-centimeter. A description of a practical transmitting and receiving apparatus for nine-centimeter waves is given. A method of modulation of radio-optical waves by voice and music is outlined. A tube detector for reception of nine-centimeter waves is described.

NOTES ON PROPAGATION AT A WAVELENGTH OF 73 CENTIMETERS

B. TREVOR AND R. W. GEORGE
(RCA Communications, Inc.)

ABSTRACT

Quantitative field strength observations have been made on a wavelength of 73 centimeters with improved apparatus. The methods are described. Propa-

gation data have been obtained with the receiver installed in an automobile and an airplane. Further observations were made on the ground at a distance of 111 miles, 8000 feet below the line of sight from the transmitter. The results show the nature of the propagation of 73-centimeter waves over distances up to 175 miles. Below the transmitter's horizon, rapid attenuation occurs with distance from the transmitter, the plane of polarization of the transmitted signal remains unchanged, and various types of fading are observed.

VACUUM TUBES FOR GENERATING FREQUENCIES ABOVE 100 MEGACYCLES

C. E. Fay, and A. L. Samuel
(Bell Telephone Laboratories, Inc.)

ABSTRACT

The failure of conventional vacuum tubes to oscillate above some critical frequencies is analyzed and illustrated by data on a tube capable of oscillation at frequencies up to 300 megacycles. A Barkhausen tube giving output of the order of five watts in the range from 450 to 600 megacycles is described. For higher frequencies (up to 2500 megacycles) spiral grid Barkhausen tubes have been used. By departing from conventional construction principles it is possible to extend the operation of negative grid oscillators above 300 megacycles, one tube described in detail giving six watts at 500 megacycles. By further refinement appreciable power has been obtained at 1000 megacycles and the possibilities have by no means been exhausted.

THE 1932 ECLIPSE OBSERVED BY RADIO FACSIMILE

E. F. W. Alexanderson
(General Electric Company)

ABSTRACT

In connection with the eclipse in August, 1932, a good deal of interest was shown in the theory of the electronic shadow which was expected during the two hours before the optical eclipse. A radio receiving station equipped to take continuous facsimile records of a signal from Schenectady was therefore set up at Conway, New Hampshire. The records show a marked change in conditions during the period of the expected electronic shadow. The paper contains samples of the facsimile records and an interpretation of the phenomena observed.

SOME RECENT WORK ON THE IONOSPHERE IN CANADA

J. T. Henderson
(Canadian National Research Council)

ABSTRACT

A brief résumé of the results of ionospheric measurements at McGill University, by Dr. W. B. Ross for the period January to August, 1933, is given. Generally speaking, these are in good agreement with those of other workers but an anomalous case where it has been possible apparently to get echoes from re-

gion F at times when region E was much more intensely ionized than F, was noted in a few cases. The method of automatic recording recently set up at the National Research Council in Ottawa is described. The output of the receiver is impressed on a cathode-ray oscillograph and a special form of time base devised which permits reference lines, denoting definite layer heights, to be recorded continuously on the photograph.

STUDIES OF THE IONOSPHERE BY MULTIFREQUENCY AUTOMATIC RECORDING

T. R. Gilliland
(Bureau of Standards)

ABSTRACT

This paper shows the diurnal and seasonal changes of ionosphere layer heights and critical frequencies, observed with an automatic recorder covering a frequency band from 2500 to 4400 kilocycles. Records were obtained each hour during most of the past year. During the daytime the E, F_1, and F_2 layers return reflections at these frequencies. Other strata between E and F_1 layers are often indicated. The F_1 layer disappears in the late afternoon and is much less pronounced in winter than in summer. At night the ordinary and extraordinary rays from the F layer become critical. The intensity of ionization in this layer often increases during the night or may remain practically constant for several hours.

IONOSPHERE MEASUREMENTS AT LOW LATITUDES

L. V. Berkner and H. W. Wells
(Carnegie Institution of Washington)

ABSTRACT

Ionosphere measurements at the Huancayo, Peru, Observatory (latitude 12°S.) of the Department of Terrestrial Magnetism, Carnegie Institution of Washington, identify the E, F_1, and F_2 layers in the daytime, as found at Washington, D. C. Two reflection components are found for the F_1 and F_2 layers, reaching critical values at different frequencies, the difference corresponding closely to that calculated for the effect of magneto-ionic double refraction due to the earth's magnetic field. From these data further information on the actual ionization conditions is obtained, indicating that the ionization is due to electrons rather than heavier ions. The F_2 critical frequency appears to be several thousand kilocycles higher at Huancayo than at Washington, and its diurnal variation is much more erratic than that for the lower layers.

A HIGH-FREQUENCY ELECTRODYNAMIC AMMETER

H. M. Turner and P. C. Michel
(Yale University)

ABSTRACT

An instrument designed as a laboratory standard for the measurement of current at frequencies above twenty megacycles consists of a closed ring freely suspended in the field of an exciter coil. The exciter current is calculated

from the physical dimensions of the instrument and the observed period of the torsional vibration of the ring. The smallness of the physical dimensions minimizes the capacity effects. The impedance of the instrument is independent of the current amplitude. The power consumed is relatively small.

SOME THERMAL METHODS OF MEASURING POWER LOSS IN VACUUM TUBES

F. P. Cowan
(Harvard University)

ABSTRACT

The reading of an indicator of the thermal radiation from a tube is noted when the tube is oscillating. The same reading is produced with the tube not oscillating by proper adjustment of power input. Power loss in the oscillating tube then equals power input when not oscillating. Various indicators have been used including an air chamber with temperature controlled outer jacket and a thermopile. The latter, when used with a filter seems best. This filter removes the heat coming from the glass and leaves that from the plate thereby reducing the time of equilibrium from 15 minutes to 2. Errors due to shift in heat distribution between oscillation and nonoscillation and possible changes in filament power have been examined. Values of power loss correct to within 1 or 2 per cent even at 20 or 30 megacycles can be obtained.

THE PRIMARY FREQUENCY STANDARD AND MONITORING STATION OF THE CANADIAN RADIO BROADCASTING COMMISSION

W. A. Steel
(Canadian Radio Broadcasting Commission)

ABSTRACT

The Canadian Radio Broadcasting Commission has recently installed a frequency standard and main monitoring station, in order to control the frequency of all broadcasting stations in Canada. A series of special receivers has been constructed to cover the frequency band from 500 to 25,000 kilocycles, so that not only the broadcast band but all harmonics and high-frequency broadcast stations may be measured. The frequency standard has been specially designed for this purpose. Three quartz oscillators, each with a separate heat and power supply, constitute the standard. One crystal vibrating at 50 kilocycles drives a 1000-cycle clock through a chain of multivibrators. The average frequency of this crystal is determined from a comparison of the clock with Arlington time signals. This comparison is recorded automatically four times a day. The instantaneous frequency of the clock crystal is obtained from its comparison with the frequencies of the other two crystals which are adjusted about 0.15 cycles below 50 kilocycles. The beats between the clock crystal and each of these crystals are automatically counted over a nine-minute interval and recorded alternately. Although the equipment is used principally as the standard of frequency for Canadian broadcast stations, it also furnishes frequencies of one cycle and its decimal multiples to the laboratories of the National Research Council.

A METHOD OF MEASURING NOISE LEVELS ON SHORT-WAVE RADIOTELEGRAPH CIRCUITS

H. O. Peterson
(RCA Communications, Inc.)

ABSTRACT

This method is essentially a means for continuously measuring the percentage of elapsed time that the noise level exceeds a certain predetermined voltage level. The predetermined voltage level may be made any level at which data are desired. The device consists of a biased tube circuit such that when input signal exceeds a certain level, the output tube passes a normal value of plate current through a ballistic meter of slow period. If the input continuously exceeds the threshold level, the ballistic meter reads "100 per cent." The data obtained are particularly useful in connection with the engineering of short-wave telegraph circuits, since these are usually operated through vacuum tube relay devices at the receiving end, set to register normal output whenever the signal "marks" at any level above a controllable threshold value.

FREQUENCY DISTRIBUTION OF THE INTENSITIES OF RADIO ATMOSPHERICS

K. A. Norton
(Bureau of Standards)

ABSTRACT

The variation of intensity of radio atmospherics with frequency is determined by two factors, (a) the production of atmospherics, and (b) the propagation of the atmospherics to the receiver. A general formula for the propagation factor is derived. The Eckersley-van der Pol attenuation formulas for the ground wave are presented in a simplified graphical form. These formulas are then used in the general formula to determine the propagation factor under conditions such that most of the atmospherics arrive as ground waves. In conjunction with the empirical Austin-Cohen formula, the method is used to calculate the frequency distribution of atmospherics in the frequency band from 10 to 60 kilocycles. The agreement between theory and measurement data is satisfactory.

DEVELOPMENTS IN AUTOMATIC SENSITIVITY CONTROL

G. E. Pray
(Signal Corps Laboratories)

ABSTRACT

First is presented a short review of several circuits in common use, in which detection and automatic sensitivity control are combined in one tube. Advantages and disadvantages of each circuit are illustrated by means of curves showing operating characteristics. A discussion follows, showing how all of the desirable features, including delayed detection and delayed automatic sensitivity control, are obtained by use of a standard tetrode or pentode tube, doing away with the necessity for using special types of tubes.

PHASE ANGLE OF VACUUM TUBE TRANSCONDUCTANCE AT VERY HIGH FREQUENCIES

F. B. Llewellyn
(Bell Telephone Laboratories, Inc.)

ABSTRACT

Theoretical considerations indicate that the transconductance of a vacuum tube exhibits a phase angle when the transit time of electrons from cathode to anode becomes an appreciable fraction of the high-frequency period. Measurements show that such a phase angle actually occurs and that its behavior is in general agreement with the theoretical predictions.

A NEW METHOD OF DETERMINING THE OPERATING CHARACTERISTICS OF POWER OSCILLATORS

C. N. Kimball and E. L. Chaffee
(Harvard University)

ABSTRACT

The performance of vacuum tube power oscillators is determined by tests made at 60 cycles. The low-frequency potentials are applied to the plate and grid electrodes from the 60-cycle power mains through suitable transformers. Wattmeters placed in the circuit indicate the magnitudes of the plate and grid power losses. The wave forms of and the phase angle between the alternating plate and grid voltages are determined from 60-cycle measurements. Radio-frequency measurements with the same tubes show close agreement with the results of the 60-cycle tests. A new method of presenting the results is also given in which the operating characteristics of the vacuum tube are shown as contours on the $e_p - e_g$ plane.

GRID CIRCUIT LOSSES IN VACUUM TUBES AT VERY HIGH FREQUENCIES

B. J. Thompson and W. R. Ferris
(RCA Radiotron Company, Inc.)

ABSTRACT

This paper presents the results of a theoretical and experimental study of the input resistance of vacuum tubes at high frequencies. H. O. Peterson of RCA Communications, Inc., first observed that at wavelengths as long as 22 meters the input resistance of a radio-frequency amplifier tube is comparatively low and is a function of the operating conditions of the tube and of the frequency. In this paper it is shown that the input loading effect is due to the appreciable time of transit of the electrons across the space between cathode and anode, that the theory presented agrees qualitatively with the results, and that this effect is one of the major limitations to the use of vacuum tubes at high frequencies.

A "SHORT-CUT" METHOD FOR CALCULATION OF HARMONIC DISTORTION OF MODULATED RADIO WAVES

I. E. Mouromtseff and H. N. Kozanowski
(Westinghouse Electric and Manufacturing Company)

ABSTRACT

Precalculation of harmonic distortion produced by vacuum tubes used in all stages of a class-B audio amplifier, for a great variety of operating conditions, is an important item in the work of the radio design engineer. A graphical method for this and similar calculations, which gives considerable saving of time has been developed. It is applicable to all symmetrical periodic curves containing harmonic components plotted to the sine of the fundamental frequency, sin ωt. The procedure consists in connecting the ends of such a curve by a straight line and measuring the ordinate differences between the curve and the chord for five definite values of abscissas. Simple expressions allow, then, for a rapid calculation of harmonic components up to 11th order. The method has proved to be very useful in class-B modulator design.

SPACE-CHARGE EFFECTS IN PIEZO-ELECTRIC RESONATORS

W. G. Cady
(Wesleyan University)

ABSTRACT

When an X-cut quartz plate vibrates extensionally at its fundamental frequency in the direction of its thickness, the mechanical stress, and hence the piezo-electric polarization, may be assumed, as a first approximation, to be distributed sinusoidally in the x-direction, and to be uniform in the y and z directions. Equations are derived for the resulting space and surface charges. The components of electric intensity due to these charges are determined, as well as their effect upon the equivalent stiffness and frequency, for resonators having air-gaps of any magnitude.

LOW-FREQUENCY TRANSMISSION OVER TRANSATLANTIC PATHS

H. H. Beverage and G. W. Kenrick
(RCA Communications, Inc.) (Tufts College)

ABSTRACT

Continuous records of field intensity taken at several receiving points and on various types of antenna systems are compared. Evidence of incoherent low-frequency fading is found. Several measures of variability are discussed and changes observed are compared with variations in earth potentials. Evidence for analogous transmission phenomena to those commonly associated with high-frequency transmission is discussed.

INPUT IMPEDANCE OF VACUUM TUBE DETECTORS AT ULTRA SHORT WAVES

A. B. Crawford
(Bell Telephone Laboratories, Inc.)

ABSTRACT

An impedance substitution method, employing a Lecher frame, has been used to measure the variation with frequency of the input impedance of several

types of receiving tubes operating as balanced detectors in the wavelength range 1 to 5 meters. The measurements give the values of the parallel combination of resistance and reactance that is equivalent to the tube impedance at a particular frequency. Practically all the tubes measured exhibit the properties of a multi-resonant network. Only the smaller and special tubes are still capacitive at 1 meter. The necessity of reducing the inductance of the leads within the tube envelope to a minimum is shown by the curves obtained.

IONOSPHERE STUDIES AT FAIRBANKS, ALASKA

H. B. Maris
(Naval Research Laboratory)

ABSTRACT

Apparent ionosphere heights range from 130 to 700 kilometers, the average increasing from 260 kilometers in December, 1932, to 300 kilometers in March 1933. Observations made during a zenith auroral display indicated that the maximum ion density of the ionosphere may be 100 kilometers or more above the position of maximum auroral brilliancy. Signals from distant stations heard during an active auroral display indicated that auroral activity may give a swinging Doeppler shift to the frequency of a signal. Signal absorption was found to be much more important than maximum ion density in the reception of high-frequency signals. This is in agreement with the results of all attempts at commercial use of frequencies above 3000 kilocycles in Alaska.

A series of active magnetic disturbances of eight days each during January, February, March, and April, obliterated nearly all echo signals. Another series of mild one day disturbances through April, May, June, and July resulted in a great increase in efficiency of signal reception especially during daylight hours.

May Meeting of the Board of Directors

The Board of Directors met on the afternoon of May 2 in the Institute office and those present were: C. M. Jansky, Jr., president; Melville Eastham, treasurer; Arthur Batcheller, O. H. Caldwell, Alfred N. Goldsmith, R. A. Heising, J. V. L. Hogan, L. C. F. Horle, C. W. Horn, E. L. Nelson, E. R. Shute, H. M. Turner, A. F. Van Dyck, H. A. Wheeler, William Wilson, and H. P. Westman, secretary.

Thirty-one applications for Associate membership, one for Junior, and two for Student membership were approved.

The secretary was named as alternate representative on the Electrical Standards Committee.

A request from L. E. Whittemore asking that he be released from membership on the Tellers Committee because his name was placed in nomination for director was complied with.

The Medal of Honor for 1934 is voted to S. C. Hooper, Captain, U.S.N., for his orderly planning and systematic organization of radio communication in the government service with which he is associated, and the concomitant and resulting advances in the development of radio equipment and procedure. The Medal of Honor, until further action by the Board, will be made of bronze in view of the present conditions affecting the use of gold.

The Morris Liebmann Memorial Prize was voted to V. K. Zworykin of the RCA Victor Company for his contributions to the development of television.

The president was empowered to appoint a committee to prepare a review of radio developments during 1934 for possible presentation at the New York meeting to be held in December.

The possibilities of continuing the preparation of material for the "Radio Abstracts and References" section of the PROCEEDINGS were discussed at length and the president was empowered to appoint a small committee to investigate the possibilities of obtaining funds for this purpose and outline the type of bibliographical work which can be accomplished.

The Emergency Employment Service reported the placing of thirteen registrants during April and a total registration of 668. Two hundred and fifty men have been placed in positions during the past year.

Committee Work

Awards Committee

A meeting of the Awards Committee was held on the afternoon of May 2 in the Institute office and H. M. Turner, chairman; A. F. Van Dyck, William Wilson, and H. P. Westman, secretary, were in attendance.

The recommendations of the committee were prepared and submitted to the Board of Directors at its meeting later that afternoon.

Sectional Committee on Radio

Technical Committee on Vacuum Tubes

A meeting of the Technical Committee on Vacuum Tubes of the Sectional Committee on Radio which was attended by M. J. Kelly, D. F. Schmidt, and H. P. Westman, secretary, was held on the morning of April 17 to consider material which has been submitted for standardization by the International Electrotechnical Commission at a meeting to be held in The Hague late in May.

Radio Transmissions of Standard Frequencies

The Bureau of Standards transmits standard frequencies from its station WWV, Beltsville, Md., every Tuesday except legal holidays.

The transmissions are on 5000 kilocycles per second. The transmissions are given continuously from 12 noon to 2 P.M., and from 10:00 P.M. to midnight, Eastern Standard Time. The service may be used by transmitting stations in adjusting their transmitters to exact frequency, and by the public in calibrating frequency standards, and transmitting and receiving apparatus. The transmissions can be heard and utilized by stations equipped for continuous-wave reception through the United States, although not with certainty in some places. The accuracy of the frequency is at all times better than one cycle per second (one in five million).

From the 5000 kilocycles any frequency may be checked by the method of harmonics. Information on how to receive and utilize the signals is given in a pamphlet obtainable on request addressed to the Bureau of Standards, Washington, D. C.

The transmissions consist mainly of continuous, unkeyed carrier frequency, giving a continuous whistle in the phones when received with an oscillating receiving set. For the first five minutes the general call (CQ de WWV) and announcement of the frequency are transmitted. The frequency and the call letters of the station (WWV) are given every ten minutes thereafter.

Supplementary experimental transmissions are made at other times. Some of these are made at higher frequencies and some with modulated waves, probably modulated at 10 kilocycles. Information regarding proposed supplementary transmissions is given by radio during the regular transmissions.

The Bureau desires to receive reports on the transmissions, especially because radio transmission phenomena change with the season of the year. The data desired are approximate field intensity, fading characteristics, and the suitability of the transmissions for frequency measurements. It is suggested that in reporting on intensities, the following designations be used where field intensity measurement apparatus is not used: (1) hardly perceptible, unreadable; (2)weak, readable now and then; (3) fairly good, readable with difficulty; (4) good, readable; (5) very good, perfectly readable. A statement as to whether fading is present or not is desired, and if so, its characteristics, such as time between peaks of signal intensity. Statements as to type of receiving set and type of antenna used are also desired. The Bureau would also appreciate reports on the use of the transmissions for purposes of frequency measurement or control.

All reports and letters regarding the transmissions should be addressed to the Bureau of Standards, Washington, D. C.

Institute Meetings

Atlanta Section

A meeting of the Atlanta Section was held at the Georgia School of Technology on February 15 and was presided over by H. L. Reid, chairman. Twenty-nine members and guests were in attendance.

A paper on "Harmonic Analyzers" was presented by I. H. Gerks, Professor of Electrical Engineering at the Georgia School of Technology. In addition to presenting a theoretical discussion of the design and operation of harmonic analyzers, Professor Gerks demonstrated its operation with equipment set up for that purpose. The paper was discussed by Messrs. Daugherty, Reid and Wills.

The March meeting of the Section was held on the 15th at the Georgia School of Technology. Twenty-five were in attendance and the presiding officer was H. L. Reid, chairman.

Professor Gerks presented a paper on "Cathode Ray Oscillographs." He pointed out that brilliance was directly proportional to the applied voltage. The threshold effect was discussed and its importance in measurement work outlined. The use of the tubes for measuring purposes was described and a demonstration given by means of such equipment. Messrs. Bangs, Daugherty, Reid, and Wills participated in the discussion.

Boston Section

A meeting of the Boston Section was held on March 23 at Massachusetts Institute of Technology. G. W. Kenrick, secretary, presided and the attendance was 100.

A paper on "Municipal Police Communication Systems" was presented by C. E. Tucker, Professor of Electrical Engineering at Massachusetts Institute of Technology. The subject was introduced with a description of the requirements and equipment used in municipal police communication systems. The relation of wire and telephone circuits to the radio facilities of such a system adapted for large urban areas was discussed. Illustrations were chosen from the results of recent surveys in the vicinity of Boston. Advantages and disadvantages of radio as compared to wire facilities and blinker systems were considered and the logical fields of each pointed out. Several of those present participated in the discussion. The dinner which preceded the meeting was attended by twenty.

The April meeting of the Boston Section was held on the 20th at Harvard University and was presided over by E. L. Chaffee, chairman. Stuart Ballantine of the Boonton Research Corporation presented the paper on "High Quality Broadcast Transmission Reception" which was published in the May PROCEEDINGS. It was discussed by a number of the fifty members and guests in attendance. Twenty-five were present at the informal dinner which preceded the meeting.

Buffalo-Niagara Section

On April 11 a meeting of the Buffalo-Niagara Section was held at the University of Buffalo. L. Grant Hector, chairman presided and the attendance was forty-one.

A paper on "Airway Radio Beacons and Transmitters and Their Operation" was presented by J. G. Petrison, chief operator in charge of WWAB, the Department of Commerce station at the Buffalo Airport. A general discussion followed its presentation.

Cincinnati Section

A meeting of the Cincinnati Section was held on March 13 at the University of Cincinnati with R. E. Kolo, chairman, presiding. Thirty members and guests were present.

A paper on "General Tube Design" was presented by W. R. Jones a commercial engineer for Hygrade Sylvania Corporation. In it, he pointed out many of the troubles experienced in the manufacture of tubes and explained the various changes made to correct these faults. He also explained in detail some of the defects peculiar to particular types of tubes and described their limits and precautions necessary for their use.

The meeting was concluded with the showing of a motion picture entitled the "Bradleyometer" which was commented on by D. S. W. Kelly, chief engineer of the Allen-Bradley Company.

The April meeting of the section was held on the 17th at the Engineers Club in Dayton, Ohio. Two hundred and fifty were present and Chairman Kolo presided.

The meeting was devoted to the subject of "Blind Landing Systems for Aircraft" which was preceded by an inspection trip to Wright Field which was taken during the afternoon. The first paper presenting the pilot's viewpoint was given by Captain A. F. Hegenberger of the Wright Field Aircraft Laboratory. He introduced the subject with a discussion of the need of the army and commercial aviators for control of their planes when weather conditions do not permit satisfactory visibility. Blind flying using station beacons and artificial horizon in conjunction with the other instruments found on aircraft does not completely solve the landing problem. The Wright Field Laboratory developed a method of landing employing two beacons of slightly different frequency separated about a mile and a half and in a direct line toward the runway on which the plane is to land. The pilot locates his ship over the line of the two transmitters in landing.

The second paper was on "The Radio Compass" and was presented by G. G. Kruesi of the Wright Field Aircraft Laboratory. His discussion covered the equipment carried by the plane to enable the pilot to make use of the signals from the beacon transmitters. The history of early developments showed the first types gave direction over one quadrant but that all four quadrants were identical. The later developments reduced this difficulty to give absolute location by slight frequency changes in the marker beacons. The demonstration of radio compass equipment as applied to broadcast station transmission was shown.

The third portion of the meeting was devoted to a "Technical Description of Ground Equipment" by C. D. Barbulesco of the Wright Field Aircraft Laboratory. He discussed ground equipment necessary for the functioning of the flight instruments. The apparatus consisted of small portable transmitters which could be stationed at proper locations as direction beacons. In addition, a boundary marker is placed on the edge of the field. This operates at ultra-high frequencies and is picked up by a tuned antenna and receiver which operates a neon light on the control panel at the time when the plane passes directly over the boundary of the field. It requires no adjustment by the pilot and a demonstration was given of its operation.

Cleveland Section

F. T. Bowditch, chairman, presided at the April 26 meeting of the Cleveland Section.

A paper on "Radio Receiver Circuits" was presented by J. A. Burtch, President of the Shellburn Radio Manufacturing Company in whose offices the meeting was held. The speaker outlined improvements in the direct coupled audio-frequency amplifier which permits stable operation with variation in resistor values of as much as twenty per cent. Loud speaker developments to permit a single cone to reproduce all frequencies from 16 to 12,000 cycles were outlined. Equipment embodying these advances was demonstrated by means of an audio-frequency oscillator as well as the use of records and direct pick-up by piezo-electric microphones.

A new type of automobile radio receiver was demonstrated as was short-wave reception by means of converters and regular receivers. The paper was closed with a short discussion of some of the factors which advertising and sales aspects dictate in design of radio equipment. The attendance totaled fifty-eight.

Los Angeles Section

The regular meeting of the Los Angeles Section was held at the Exposition Park Studios of the Department of Parks on February 20 and was attended by sixty-one. H. C. Silent, chairman, presided.

E. H. Price, manager of the Mackay Radio and Telegraph Company presented a paper describing the "Mackay Communication System." He was followed by L. Winser, Manhattan Beach Marine Superintendent of the same organization, who spoke on the technical and operational features of the system. A general discussion followed. Fifteen were present at the informal dinner which preceded the meeting.

The March meeting of the Los Angeles Section was held on the 20th at the Exposition Park Studios. H. C. Silent, chairman, presided and forty-nine were present. Fourteen attended the informal dinner which preceded the meeting.

A discussion on "Tube Testing from the Service Viewpoint" was presented by J. L. Mahon of the Hygrade Sylvania Company. The design features of various tube checker circuits employed for servicing of broadcast receiving tubes were considered and their limitations and possible sources of error were pointed out. R. G. Leitner, engineer for the Radio Kit Company, then spoke on vacuum tube measurements from the radio manufacturer's viewpoint. He discussed methods for determining the suitability of a particular tube type for special purposes other than those for which the tube was originally designed. It was pointed out that for this purpose, measurements in a complete circuit simulating operating conditions are generally to be preferred to predictions of performance based on conventional output characteristic measurements.

In the general discussion which followed these two papers other members contributed the results of their own work on various phases of vacuum tube measurement.

The April meeting of the section was held jointly with the American Institute of Electrical Engineers Los Angeles Section and student branches of the University of Southern California and the California Institute of Technology. At the University of Southern California, A. P. Hill, chairman of the Los Angeles Section of the American Institute of Electrical Engineers, presided. The following papers were presented:

"Dielectric Strength of Oil Immersed Solids," by R. E. Kidd, California Institute of Technology.

"A New Illuminometer," by W. L. Patton, California Institute of Technology.

"Aeronautical Radio Applications," by G. W. Weaver, California Institute of Technology.

"Characteristics of Coil Form," by T. Onaka, University of Southern California.

"Oscillographic Studies of Adjustable Speed Commutator Type Alternating-Current Motors," by Frank L. Long, University of Southern California.

"Operating Characteristics of Adjustable Speed Commutator Type Alternating-Current Motors," by George S. Rives, University of Southern California.

About 200 members of the various sections and branches were present at the meeting and 160 were present at the informal dinner which preceded it.

New Orleans Section

The December 29 meeting of the New Orleans Section was held in the Monteleone Hotel with J. A. Courtenay, chairman, presiding. It was devoted entirely to a business meeting and twenty-four were present.

A meeting was held on February 21 at the St. Charles Hotel. J. A. Courtenay, chairman, presided and sixty members and guests were in attendance.

A paper on "Radio as Used in Military Communications Systems" was presented by J. E. Raymond, Lieutenant, U.S.A. Infantry School at Fort Benning, Ga. In it he outlined the advantages and disadvantages of the use of radio in military field operations. The chief disadvantages are that communication may be broken up by interfering signals, code material received by the enemy for deciphering, and the enemy may locate the position of the transmitter by taking cross bearings, and might judge the size of the force from the number of messages handled. A complete set of field equipment was available for inspection and a number of those present participated in the discussion of the paper.

Shushan Airport was the place of the March 16 meeting which was attended by thirty. Inspection of the airport and its equipment was made under the direction of D. A. Madere, chief dispatcher. The radio and public address equipment was described and demonstrated. A transmitter which is for local traffic control is of fifteen watts output and may be either voice or telegraph modulated. It is crystal controlled and operates on a frequency of 208 kilocycles. It is remotely operated from the control tower as is all the electrical equipment. The public address system consists of two forty-watt channels with centralized radio. A velocity microphone is employed for pick-up and the equipment may be used from several points about the main building. Two automatic phonographs are used with it.

The April meeting was held on the 2nd at the Monteleone Hotel. Eighteen members and guests were in attendance. J. A. Courtenay, chairman, presided.

A paper by A. A. Schiele of the Electron Engineering Company on "Practical Applications of Photo-Electricity" was presented. It covered in detail the commercial application of photo-electric devices for counters, fire alarm apparatus, door closing and opening and other uses. The paper was illustrated by a demonstration of equipment which was made available for the purpose. Several of those present participated in the discussion.

New York Meeting

A joint meeting of the Institute and the Amateur Astronomers Association was held on May 2 in the American Museum of Natural History where a paper on "Astronomical Factors and Radio Transmission" was presented by J. L. Richey, chief technical operator of the Overseas Radiotelephone Service of the American Telephone and Telegraph Company.

The speaker pointed out that studies of radio transmission have shown a regular daily and seasonal change, which, of course, is intimately related to the earth's position with respect to the sun. In addition to this other changes occur, which appear to be associated with solar phenomena and other cosmic influences. Experience has indicated that the effects of the solar phenomena are felt with different intensities on radio paths across different zones. The changed radio conditions appear to follow periodicities, which are related to the rotation period of the sun, and the sun-spot cycle. A relationship between these phenomena and other earthly happenings, such as frequency of auroras, fluctuations in terrestrial magnetism, and the presence of enormous electric current sheets, which oscillate in the earth's outer crust, have been observed. Supplementing this discussion, lantern slides of related astronomical phenomena were shown as were moving pictures illustrating the fluctuations taking place in the Kennelly-Heaviside layers.

Philadelphia Section

A meeting of the Philadelphia Section was held at the Engineers Club on April 5. W. F. Diehl, chairman presided and 300 members and guests were present at the meeting.

A paper on "Television Image Pick-Up and Reproduction by Means of Electron Streams" was presented by V. K. Zworykin, director of electrical research of the RCA Victor Company.

The speaker presented a detailed description of the iconoscope used for television pick-up and the kinescope employed for receiving television pictures. These are cathode ray tubes. The iconoscope has a receiving screen composed of an immense number of small condensers which are charged by light and discharged by a scanning cathode ray beam. The amount of current discharged from each condenser element is proportional to the amount of light falling upon it. Curves of different factors limiting the physical possibilities of television pick-up were shown and the method of fitting the iconoscope to a camera for pick-up purposes was described. The focusing of cathode ray tubes to a point by means of an electrical lens was also covered. New tubes sensitive to ultra-violet light which may be used for taking microscopic pictures of objects too small to be seen by the visible light spectrum were described. Messrs. Budche, Cook, McIlwain and others participated in the discussion. Fourteen were present at the informal dinner which preceded the meeting.

The annual meeting of the Philadelphia Section was held on May 3 at the Engineers Club and was presided over by W. F. Diehl, chairman. Two hundred members and guests were present and fourteen attended the informal dinner which preceded the meeting.

A paper on "Radio City Studios and Equipment" was presented by O. B. Hanson, chief engineer of the National Broadcasting Company. The paper was profusely illustrated with views of the studios and equipment used for broadcasting at the RCA building in New York City. The intricacies of this system was indicated by the description of a control board containing over 2000 signal lights, each one designating a particular action in a program. Descriptions were given of numerous large and small studios enclosed in floated soundproof rooms with air conditioning, echo and delay sounds, stages with solid glass curtains permitting the audience to see but not to be heard, sound absorbing panels and drapes for modifying the acoustics of studios and many other interesting features of building and construction equipment.The paper was discussed by Messrs. Cook, McIlwain and others.

During the short business meeting which was held, the report of the secretary-treasurer was read and the annual election of officers held. The newly elected chairman is E. D. Cook, RCA Victor Company; vice chairman, Knox McIlwain of the Moore School of Engineering, University of Pennsylvania; and secretary-treasurer, R. L. Snyder, Bell Telphone Company of Pennsylvania.

Pittsburgh Section

A meeting of the Pittsburgh Section was held on January 23 at the Fort Pitt Hotel. It was presided over by Lee Sutherlin, chairman.

A paper on "Control Room Operation in a Modern Broadcast Station" was presented by D. W. Myer, plant manager of KDKA and W8XK. In it, the speaker outlined the early history of broadcasting and traced the development in control room equipment and operation. The changes necessitated by the use of several studios to permit continuous programs were stressed. Operating procedure was discussed and technical descriptions given of the equipment and circuits used. Methods of routine testing, loading, and trouble shooting were covered. A general discussion followed and was participated in by many of the thirty-one members and guests in attendance.

The March meeting of the section was held on the 20th at Carnegie Institute of Technology. The attendance was fifty-two and Chairman Sutherlin presided.

The meeting was devoted to the subject of low voltage cathode ray oscillographs and two papers were presented. The first by C. Williamson, Professor of Physics at Carnegie Institute of Technology, covered the historical and elementary side of the subject. He exhibited several of the old type Braun tubes which obtained electron emission from a cold cathode through the use of a high voltage spark coil producing potentials of the order of 100 kilovolts or more. He demonstrated how a magnetic field would deflect cathode rays and showed various figures produced by superimposing different frequencies on a carrier frequency.

L. E. Swedlund of the Westinghouse Research Laboratories then discussed the newer applications of cathode rays. Three nine-inch tubes were operated in parallel and demonstrated many figures due to the application of varying excitation voltages. How the grid of a vacuum tube acts as a barrier to the electrons in their movement from the filament to the plate of a vacuum tube was also demonstrated.

San Franscico Section

The San Francisco Section held a meeting at the Bellevue Hotel on March 24 which was presided over by G. T. Royden, chairman, and attended by forty-two members and guests. Twelve were present at the dinner which preceded the meeting.

A. R. Rice of the Navy Department presented a paper on "Radio Time Signals." A general discussion followed.

Seattle Section

C. E. Williams, vice chairman, presided at the March 26 meeting of the Seattle Section which was held at the University of Washington.

A paper on "An Experimental High Gain Amplifier" was presented by W. R. Hill of the University of Washington. The amplifier which was described and demonstrated was built for amplification of frequencies from zero to serval thousand cycles. It comprised a 57 tube resistance coupled to a 2A3. The plate load resistor for the 57 was a 199 type tube with "flashed" filament constituting a very high resistance load which made possible a power gain of 10^9 which is quite unusual in a two stage amplifier. A discussion was participated in by Messrs. Fisher, Hackett, and Libby.

The second paper presented was on "Frequency Allocation in Our International Services" by J. R. Redman, Lieutenant Commander, U.S.N. He presented a review of the early history of radio in the United States Navy and the bearing of this history on the frequency allocation which we have today. The chief results of the international radio conferences to which he was a delegate were the decisions to stabilize transmission frequencies and to assign frequencies to services rather than to nations. He brought out the complexity of the allocation problem due to its international character and to the conflicting and sometimes selfish demands of the interested nations. The paper was discussed by Messrs. Bach, Kraft, Libby, and Smith.

Forty-two members and guests attended the meeting.

The April meeting of the Seattle Section was held on the 6th at the University of Washington and Howard Mason, chairman, presided. Fifty-three members and guests were in attendance.

"Engineering Problems in Sound Reproduction" was the subject of a paper by H. C. Silent of Electrical Research Products, Inc. Mr. Silent is also chairman of the Los Angeles Section of the Institute.

In his paper, he tabulated the outstanding methods of sound reproduction which are sound on disc, sound on film, and the telegraphone. He then proceeded to outline the system used by the Western Electric Company in recording and reproducing. The vast number of changes of form through which the sound is transferred from its arrival at the recording microphone to its reproduction in the theater as sound again was stressed. Some twenty-eight such transformations occur in a typical case. Variations in recording technique in studio, location, and news work were outlined. The paper was closed with a discussion of the Selsyn system which permits the camera and recording apparatus to be maintained in synchronism though they may be located at substantial distances from each other.

The second paper of the evening was on "Noise Measuring Devices" by D. P. Loye of Electrical Research Products. He described and demonstrated apparatus developed in the Bell System for recording average noise levels as well as apparatus for analyzing noise spectra.

Toronto Section

The Toronto Section met at the University of Toronto on March 13 with W. F. Choat, chairman, presiding.

A paper on "Some Problems in the Design of Radio Receivers for Automobiles" was presented by V. M. Graham of the Stromberg-Carlson Telephone Manufacturing Company. The paper was opened with an outline of the principal design characteristics essential in automobile radio receivers. The need for standardization in various design requirements was discussed and the minimum space requirements considered. Ignition interference at broadcast and high frequencies and the need for more effective suppression methods and the necessity of installing suppression of all automobiles at the time of manufacture were discussed.

Messrs. Bayly, Fose, Oxley, Patience, Van Sickle, and Wright participated in the discussion of the paper.

W. F. Choat, chairman, presided at the April 18 meeting which was held at the University of Toronto and attended by fifty-two.

Arnold Pitt of the Department of Physics of the University of Toronto presented "Some Experiments in Physics with Electron Tubes." The first demonstration was that of a Coherer type receiver indicative of early radio activity. The principles of the direct-current amplifier which is widely used in physics, the Crooks tube showing that cathode rays travel in straight lines, and the cathode ray oscillograph showing the effect of a variable audio frequency plotted against a fixed frequency were next demonstrated. A condenser microphone and audio-frequency amplifier demonstrated heartbeat sound reproduction. Some records showing the effect of suppressing various portions of the audio-frequency spectrum in music and speech were played.

The presence of harmonics in sound was then demonstrated, the harmonics being picked out of a sound wave and amplified. The measurement of audio frequencies in amplitudes by matching against a standard frequency oscillator was demonstrated. The Doeppler effect of standing audio-frequency waves was illustrated. Alpha rays were detected through the use of an FB54 tube followed by a high gain amplifier. Cosmic and gamma rays were made audible by use of suitable detection and amplifying equipment.

Washington Section

A meeting of the Washington Section was held on April 9 in the auditorium of the Potomac Electric Power Company. T. McL. Davis presided and the attendance was seventy-eight. Twenty-six were present at the informal dinner which preceded the meeting.

A paper on "Negative Resistance Devices" was presented by E. W. Herold, research engineer for the RCA Radiotron Company. The paper started with a general discussion of negative resistance presenting a picture of why and how negative resistance is possible. The treatment led up to the establishment of two main classes of negative resistance, the current controlled and the voltage controlled types. The major differences between these two were tabulated and illustrated with the aid of slides. The desirable characteristics of a negative resistance for use in communications were then discussed in some detail; the consideration of the unavoidable self-reactance of a negative resistance enabled a figure of merit to be established for each class. Means of producing negative resistance were than arranged in three groups according to the principle of operation, the simple, the direct coupled, and the reverse-phase coupled groups being differentiated. Each device was then discussed from the point of view of the desirable characteristics already mentioned. One of the more promising arrangements of the direct coupled group was illustrated by presenting a circuit and curves for a 57 tube used as a retarding field negative transconductance tube to produce a negative resistance whose properties appear to be superior to the hitherto much used dynatron. Applications of negative resistance in the production of oscillations, as circuit elements in a stable state and in measurement work were then mentioned. In conclusion, applications and general properties of the negative transconductance tube used not as a negative resistance but as an amplifier were briefly discussed.

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TECHNICAL PAPERS

SUPPRESSION OF INTERLOCKING IN FIRST DETECTOR CIRCUITS*

By
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(Palo Alto, California)

Summary—*Interlocking in the oscillator circuit of superheterodynes has been a subject of study for several years. The electron coupled oscillator did a great deal to solve the problem. The advent of the pentagrid converter brings up a new set of causes by which interlocking can occur. These causes, and the manner in which they manifest themselves, are discussed in this paper. A new circuit is proposed which has been found to reduce greatly the amount of interlocking which is found to occur in the conventional circuits. A new tube structure is also proposed which should accomplish the same results.*

Introduction

The pentagrid converter, introduced about a year ago, seemed at the time of its introduction the answer to a great many problems incident to first detectors. Its general acceptance at the present time is indicative of how well the problems have been met.

There are applications, however, in which the converter tube and circuit may prove to be too subject to interlocking. For example, for short waves, or beat frequency oscillators, wherein the different frequency is a very small percentage of the signal frequency, the interlocking may be so excessive as to render the system usable only with considerable difficulty.

It is the purpose of this paper to offer means of reduction of the interlocking which exists in conventional circuits by two means: (1) By utilizing existing tubes and introducing a new circuit; (2) by leaving the external circuits unaltered and suggesting alterations of the tubes.

Causes of Interlocking

Fig. 1 is a diagram of the conventional pentagrid circuit, with circuit constants as recommended by the manufacturer.† While experimenting at low frequencies with a circuit of this type, the writer found that there was an excessive tendency for the oscillator to synchronize

* Decimal classification: R162. Original manuscript received by the Institute, January 26, 1934.
† See "RCA-Radiotron-Cunningham Radio Tube Manual," series RC-11, application of 1A6, 2A7, 6A7.

with the frequency of whatever voltage was being injected into the signal control grid, or grid no. 4. This may be illustrated by Figs. 2 and 3 which show the change of frequency of the oscillator as the frequency of the injected voltage is varied through the normal frequency of the oscillator. These are based on actual data, but used here for qualitative explanation only and hence not drawn to any particular

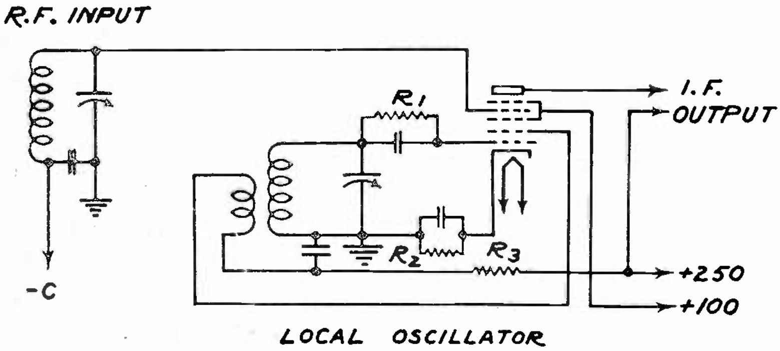


Fig. 1—Conventional pentagrid converter circuit.

scale. The figures show how the "local oscillator" will change in frequency as the frequency of a voltage injected into the signal grid of a converter circuit is varied through the normal frequency of the local oscillator.

The cause of this synchronization was found to be due to the division of current between the anodes as the voltage on the signal con

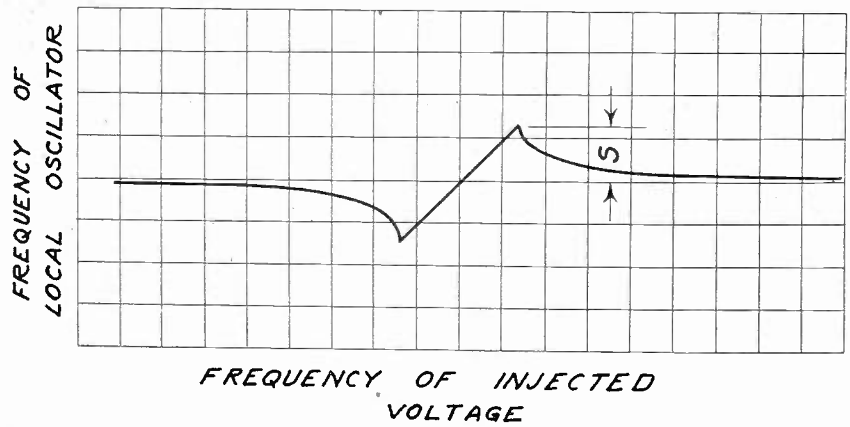


Fig. 2—Typical interlocking characteristic.

trol grid was varied. The static curves of the tube showing this effect are shown in Fig. 4. It is apparent that an oscillatory voltage on the control grid will induce a current of the same frequency in the oscillator anode and its associated circuit so that, in effect, the voltage on the control grid is amplified and fed into the oscillatory circuit, where it tends to pull the oscillator into step with it.

The explanation of this change in current is as follows: The effect of the voltage on the control grid is to change the division of current between the plate and the other anodes. As the plate current decreases the current to the other anodes, i.e., grids 2 and 3, must increase since

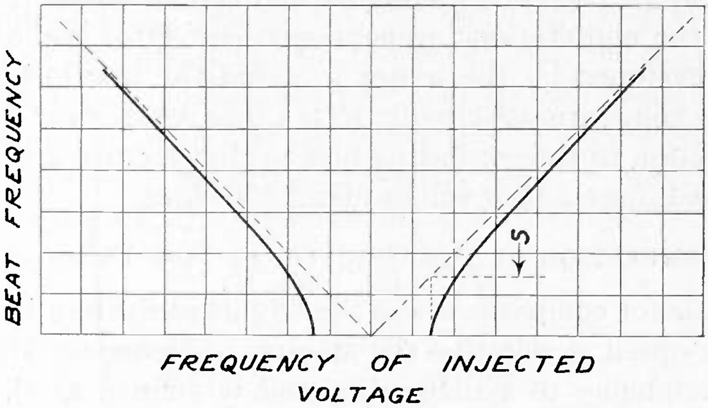


Fig. 3—Typical interlocking characteristic.

the shielding prevents the control grid from having much effect on the total emission. The fact that this current change is shared by the screen and the oscillator anode grid is due to the space charge being

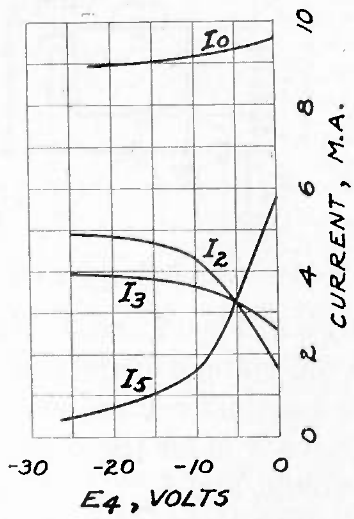


Fig. 4—Static curves for 2A7. I_0, total space current; I_2, current to grid no. 2 or oscillator anode; I_3, current to screen grids nos. 3 and 5; I_5, current to plate. E_4 is the potential on the signal control grid no. 4.

caused to vary in the region of the screen so that the final destination of an electron is a result of the relative voltages on the two anode grids, and the velocity of the electron.

The New Circuit

Apparently, if this change in oscillator anode current can be prevented from flowing in the oscillator tank circuit, this cause of interlocking would be removed. This was accomplished by by-passing the current change to ground. As the oscillator anode is thus at ground potential as far as the oscillator or injected frequencies are concerned, it is necessary to use the floating cathode circuit used extensively with electron coupled oscillators using 24 and similar tubes. The complete circuit for the noninterlocking pentagrid converter is shown in Fig. 5, and was developed by the writer to avoid the interlocking found to exist in the conventional circuit. R_2 is a bias resistor to furnish a small bias in addition to the oscillating bias so that control grid current can be minimized; for a 2A7 it will be about 300 ohms.

Comparison of the Circuits at Low Frequency

The basis for comparison was the "Synchronization Factor" which is a unit adopted to describe the amount of frequency shift resulting from the tendency to synchronize, and is defined as the maximum amount by which the frequency changes, expressed in per cent of the normal frequency of the oscillator. The term is abbreviated to the

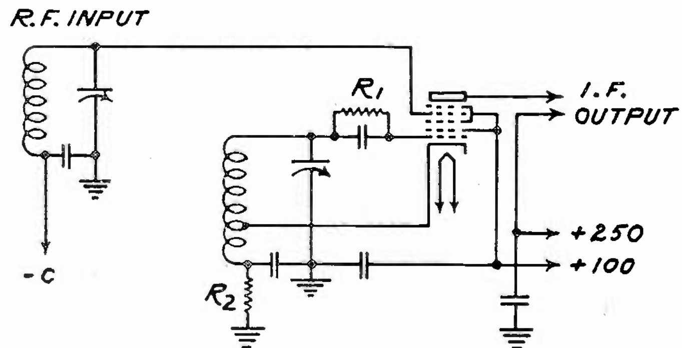


Fig. 5—Noninterlocking circuit.

letter "S" and is so shown in Figs. 2 and 3. The value of S is ordinarily nearly proportional to the voltage of the injected frequency.

Curves of "S" plotted against voltage level of the injected frequency are shown in Fig. 6. Curve A is for the conventional circuit of Fig. 1; curve B is for the new circuit, Fig. 5.

The tests were run at 315 cycles where the tube capacities were not effective in producing any coupling; the only coupling was that due to the static characteristic of the tube. The Q, L/C ratio, and losses of the two circuits were maintained at as nearly the same value as possible.

It might be expected that the proposed changes would render the tube totally free from any tendency to synchronize. But in Fig. 4 it will be noted that there is a change in the total emission current, I_0, which is coupled into the ground end of the tank inductance through the cathode lead. This gives rise to a small residual coupling, which accounts for the fact that even the circuit of Fig. 5 has some tendency to synchronize. However, it must be recognized that the circuit of Fig. 5 is superior by a ratio of ten or twenty to one over the conventional circuit.

Comparison at High Frequencies

The circuit of Fig. 7 was set up with constants such that it would operate over the broadcast band. The change from the conventional to the new circuit was accomplished by merely changing plug-in coils. The coil design in the case of the conventional circuit was as recommended by the tube manufacturers; for the new circuit a simple Hartley coil was used, the design of which will be discussed later.

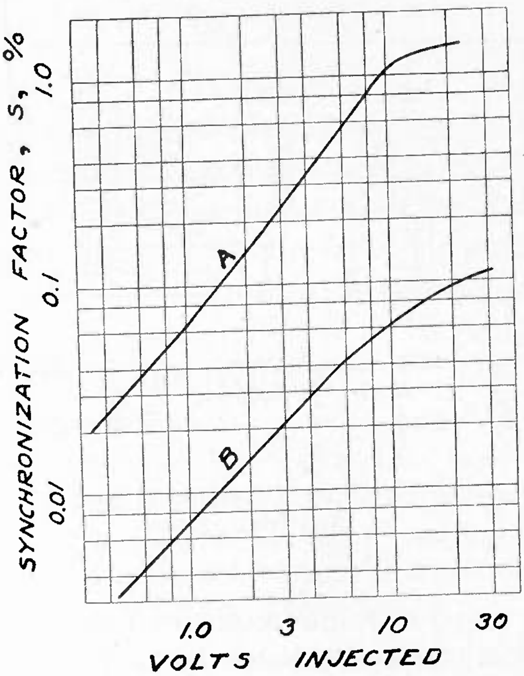
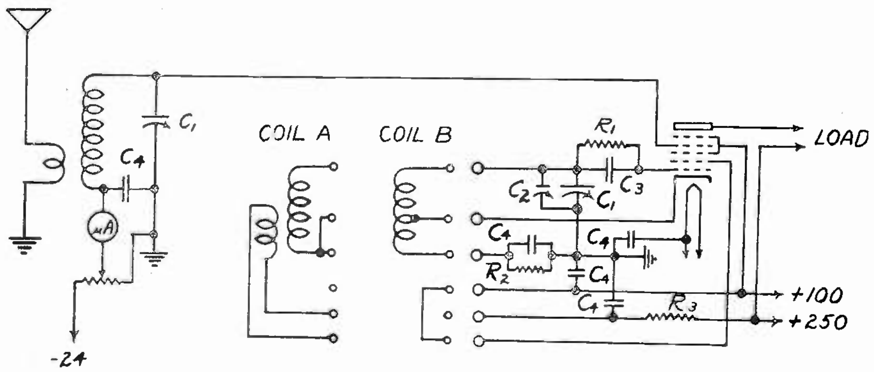


Fig. 6—Comparison of interlocking performance of circuits of Figs. 1 and 5. Synchronization factors; curve A, for circuit of Fig. 1; curve B, for circuit of Fig. 5.

A measure of the interlocking was obtained by observing the "silent interval," or region on the oscillator tuning dial over which zero beat frequency was obtained. For this interval to be appreciable, a strong signal must be used. This silent interval for the conventional circuit was approximately two dial divisions. For the new circuit, the interval was only 0.3 division. This dial was on the small condenser C_2 where one division represented a little over 100 cycles. The signal used was the carrier of KPO, 680 kilocycles, with a carrier strength at the control grid of two volts with the oscillator turned off. With the local oscillator running, the signal control grid was drawing current, so that the input signal control grid might have been excited by almost any voltage other than that measured. The cause of the presence of control-grid current will be explained later; it is due to the first cause mentioned in the explanation.

Due to the change of effective Q with grid current flowing, and thus not knowing what voltage was actually being injected, values of the synchronization factor could not be based on this measurement. As a basis for comparison of the two circuits, the measurement is quite valid in showing that the new circuit tends to synchronize over only about one sixth of the tuning range as compared to the conventional circuit.



C_1 350 μμF Max.
C_2 5 to 10 " Approx.
C_3 200 "
C_4 0.1 μF

R_1 = 50,000
R_2 = 300
R_3 = 20,000

Fig. 7—Change-over circuit. The circuit used for comparison at high frequencies.

Interlocking Without an Injected Voltage

Another effect noted at radio frequency was that of the impedance of the input or signal tank circuit on the frequency of the local oscillator. It was found that the oscillator frequency would increase as the signal circuit was tuned through the oscillator frequency. In the case of the conventional circuit, the change from 680.3 kilocycles was 1050 cycles; in the case of the new circuit, only 30 cycles. This change is due to the capacity coupling between grids 1 and 4 (manufacturer's data gives 0.15 micromicrofarad) which permits the local oscillator to induce a voltage on grid 4. The higher the impedance in the circuit of grid 4 the higher this induced voltage becomes. This voltage acts just as an injected voltage would, and while it is the same frequency as the oscillator, it is leading in phase so that in "pulling the oscillator into step" it tries to cause each oscillation to occur a little sooner than it should so that the effect is to increase the oscillator frequency. The effect in the new circuit is only 3 per cent of the effect in the conventional circuit.

Control-Grid Current

The control grid may draw current in either circuit. As noted in the preceding paragraph, the voltage on the oscillator grid has a capacitive coupling path to the signal control grid. If the return circuit impedance to the signal control grid is high enough, the voltage generated on it may reach several volts (5 volts root-mean-square were measured in one case) so that if this induced voltage exceeds the bias, current will flow. The condition for high impedance would occur when the input grid circuit is tuned to the oscillator frequency.

Another cause for grid current, in the case of the new circuit only, arises from the oscillating "bias" resulting from the floating cathode. This will be discussed under "Circuit Design."

Sensitivity

Inasmuch as the new circuit is intended to replace the conventional circuit in superheterodyne receivers, some knowledge of the relative sensitivity should be had. Measuring the voltage input and the voltage of the beat frequency output led to the following data:

	Conversion transconductance Micromho
Conventional circuit	250
New circuit	170

which show for the particular tube and circuit constants used the new circuit has about 3 decibels less gain than the conventional circuit. This is due to the self-bias generated by the oscillator.

Circuit Design

About the only element requiring extra care in the use of the new circuit is the design of the oscillator tank coil. The placing of the cathode tap is rather critical. If too many turns are included in the "plate" or ground end of the coil, the cathode has such a high positive potential during the part of the oscillation when the space current is flowing, the signal control grid being thus biased to such a high negative potential with respect to the cathode, that the plate current and conversion conductance are greatly reduced. If the tap is too low, too near the ground end, the circuit may not oscillate at all; or, if it does, the space current may flow during more than 180 degrees of the oscillation cycle so that the cathode may be negative with respect to the signal control grid while space current is flowing, with the result that the signal control grid draws current. The best compromise seems to be that of adjusting the cathode tap until about 1.5 to 2.5 milliamperes plate current flows when the oscillator anode, screen, and plate potentials are as indicated in Fig. 5. At broadcast frequencies, it was found that this condition would be realized with the cathode tap at about 11 per cent of the turns from the ground end of the coil.

The circuit was used to feed into an intermediate-frequency amplifier. Operation was normal with no peculiarities of performance which cannot be predicted from a knowledge of the conventional circuit and the information given here.

Circuit for Filament Type Tubes

The circuit of Fig. 5 is applicable only to heater type cathodes. The circuit of Fig. 8 is offered as one possible combination for tubes with

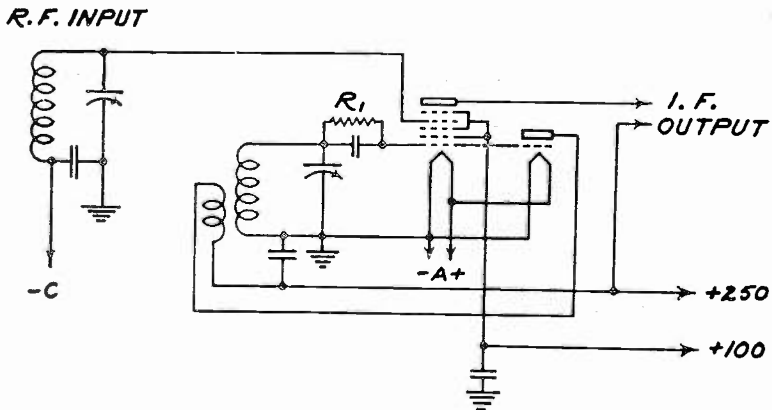


Fig. 8—Noninterlocking circuit for tubes with filament cathodes.

filament cathodes. The performance of this circuit should exhibit even less interlocking than the circuit of Fig. 5, using available tubes, since the small change in emission current (I_0 of Fig. 4) is not coupled into the oscillator tank but is grounded, as well as the oscillator anode current (I_2) being independent of the voltage applied to the signal control grid.

Changes in Tube Structure

The circuit of Fig. 8 suggests the possibility of getting the same results with a single tube which will eliminate the causes of interlocking as in circuit 5; the floating bias of Fig. 5 is also eliminated so that the conversion conductance will remain high. This circuit is probably the best for either unipotential or filament cathode tubes for short-wave work, since oscillations can be maintained at a higher level without critical adjustment of the oscillator coil. Fig. 9b shows a possibility in the design of a tube to replace the pentagrid converter without any circuit changes outside the tube itself (i.e., no alterations in the conventional circuit connections up to and including the socket).

This proposed tube form rather suggests the "triode-pentode" or 6F7 but is actually radically different in principle. It consists of the two tubes of Fig. 8 in a single envelope and with a common cathode and first or oscillator control grid.

If the signal control grid is reasonably shielded from the other elements, and the screen and oscillator plate sufficiently separated, or a baffle of some sort placed between them so that electrons which are turned back by the signal grid cannot readily reach the oscillator plate but must arrive at the screen, the use of such a tube, with the associated circuit, should result in a system which interlocks to a considerably less degree even than the circuit of Fig. 5, at the same time exhibiting the high conversion conductance of the present pentagrid converter.

Such a tube could obviously be constructed either in the unipotential or filament cathode type. Numbers in circles in figure indicate base pin numbers for heater tube; heater pins not shown. Such a tube could be plugged into the socket of a conventional circuit, and the resulting combination would really be that of circuit 8.

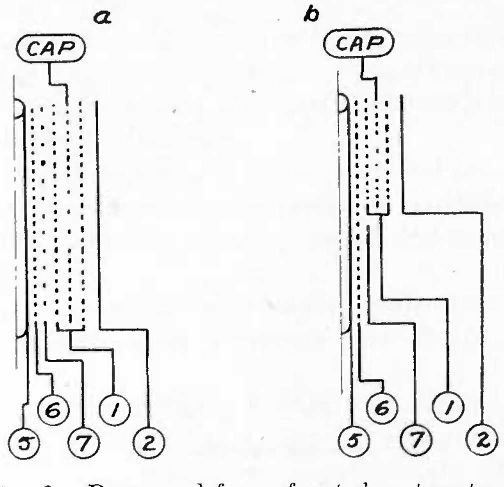


Fig 9—Proposed form for tube structure.
a. Existing form of 2 A 7
b. Proposed change

It is rather interesting to speculate on other tube arrangements which might eliminate the causes of interlocking. For example, if the shielding between grid 4 and the rest of the elements in the present converter were improved, the oscillator could not induce as much voltage on the signal control grid, which in turn could not reflect back on the oscillator frequency. The addition of a suppressor grid between the oscillator anode and the screen, connected to the cathode inside the tube, would probably limit the anode current change induced by a voltage on the signal control grid to the plate and screen only, leaving the oscillator anode unaffected, and hence offering no means of coupling a signal voltage into the oscillator circuit.

ACKNOWLEDGMENT

The writer wishes to express his appreciation to Dr. F. E. Terman in whose laboratory at Stanford University this work was done, for his guidance and suggestions during the course of the research and for his constructive criticism during the preparation of this paper. Mr. J. M. Sharp, also, deserves a large measure of thanks for suggestions resulting in greater facility and reliability in some of the measurements.

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RESISTANCE TUNING*

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Summary—*Resistance tuning is analyzed postulating the absence of inter-stage reaction and grid conductance. Reactance tuning is also analyzed under these same conditions.*

Formulas are developed by which it is possible to compute both the amplification and selectivity it is possible to attain in both cases.

The fact is brought out that if we desire to attain a maximum of selectivity (and therefore a minimum band width causing improvement in the signal-to-static ratio) without undue sacrifice in the audio quality of reception and to maintain this same "optimum" compromise over the entire tuning range, it is of importance to keep the value of inductance divided by effective resistance in the circuits equal and unchanged over the entire tuning range. With reactance tuning this is not feasible. With resistance tuning it might be feasible.

Using a minimum of two resistance tuned stages, consisting of tuned antenna and coupled by a tube to a second tuned circuit, and assuming both side bands are to be received, a possible amplification as great as 430,000 between antenna and detector grid with adequate selectivity is indicated.

With reactance tuning we would probably be forced to use additional stages to attain this result. For this reason the substitution of resistance tuning for reactance tuning might result in considerable economy of both apparatus cost and space occupied.

Resistance tuning has an advantage over reactance tuning in that without making circuit changes, we may tune in any frequency lower than the maximum frequency which the set can receive.

A single stage of resistance tuning used as an oscillator has also this same advantage and its use is indicated for supplying "heterodyne" effects in long-wave reception and carrier effects in telephonic reception where the carrier and one side band have been eliminated in the transmitter.

It might also be used as an adjustable audio-frequency oscillator with economy as it eliminates the necessity of heterodying two high frequencies in order to produce the desired range of audio frequencies.

A BROAD definition of a tuned circuit might be worded somewhat as follows: An impedance consisting of two components in one dimension and a third component in a second dimension at quadrature to the first dimension.

The above-mentioned two components are of opposite sign, one of them being directly proportional to frequency and the other inversely proportional to frequency, and thus, at a particular desired

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frequency, they become equal and being of opposite sign, they cancel. At this point their magnitude is relatively great compared to that of the above-mentioned third component. This ratio measures the voltage "resonant rise." If we desire to control this ratio, we may do so by adjusting the value of the third component.

This control may be secured by some form of adjustable negative resistance effect and if we adjust the value of the third component to zero by this means, the circuit becomes self-oscillatory at the desired frequency.

Reactance Tuning

In engineering practice the two above-mentioned components are plus and minus reactances and are located in the nondissipative or "J" dimension. The third component above mentioned is located in the dissipative or "resistance" dimension at quadrature to the "J" dimension.

Tuning is done by variation of reactance which governs the effects in the nondissipative or reactance dimension. This might be defined as "reactance tuning." It has been with us since the time of Maxwell and Heaviside.

Now the point of the above wordy definition of tuning is to try to find a general definition of tuning which might also cover in the same words another means of tuning in which we merely exchange the planes of the two dimensions above mentioned.

Resistance Tuning

With this exchange of dimensions, tuning is done by variation of a relatively large adjustable resistance. The two aforesaid components are positive and negative resistances and the third aforesaid component is a small negative reactance which may be made as small as desired or zero by adjustment of the negative resistance effect. This might be defined as Resistance Tuning.[1]

Comparing the two foregoing paragraphs, it will be evident that the resistance tuned circuit is in effect a "J" transformation of the reactance tuned circuit.

B. van de Pol has pointed out the possibilities of "J" transformation which involve the use of a negative resistance.[2] See also L. C. Verman.[3]

This particular transformation is brought about by establishing particular relations in an alternating-current bridge network.

[1] Cabot, U. S. Patent 1,853,604, April 12, 1932.
[2] Van der Pol, Proc. I.R.E., vol. 18, p. 221; February, (1930).
[3] Verman, Proc. I.R.E., vol. 19, p. 676; April, (1931).

Practical Applications of Tuning

Tuned radio-frequency amplification is often used to receive and amplify selectively in radiotelephony. To obtain a satisfactory performance, we must arrange to cover the most useful audio frequencies in the "side bands" with the least possible change in voltage amplification, and then, as the frequency "off resonance" increases to a point where the audio effects are of less importance, we must provide for the greatest possible decrease in voltage amplification in order to get adequate selectivity. We must, in any case, make some form of compromise between adequate audio performance and adequate selectivity.

For the purpose of calculating the performance of the set in this respect, engineers often plot a "selectivity" curve using cycles "off resonance" as abscissa and the ratio of voltage "at resonance" to voltage at a specified number of cycles off resonance over the audio range to be received as ordinates.

The more rectangular in shape we can make this curve, the greater will be the selectivity and the less will be the band width of frequencies to which the set responds which in turn causes a greater ratio of desired signal to undesired static.

We shall find on analysis (see Appendix I) that at least two tuned stages are necessary to accomplish adequate rectangularity, and that for optimum results in this respect, it is of importance that the value of the "time constant" L/R in each stage be the same, and furthermore that it be maintained constant over the entire tuning range. With a minimum of two tuned stages and assuming we desire to include both side bands, the optimum indicated value for the "time constant" L/R is about 6.5×10^{-5}, for frequencies within the tuning range.

With the present reactance tuned sets which contain no means for keeping the above-mentioned third component which is the effective resistance in the oscillating circuit at a constant value so that $L/R=6.5\times10^{-5}$ in each of the two stages independent of the carrier frequency to which we tune, it is not possible to maintain the above-mentioned best compromise between adequate selectivity and adequate audio performance.[4]

The reason for this is that the effective resistance has a way of varying approximately as the square of the proportional change in frequency so that over the tuning range, which is limited to about 3 to 1 in frequency, we shall find on analysis (see Appendixes I, II, and IV) that we would have to go 9 to 1 in cycles off resonance to get the same voltage cut-off at the high-frequency limit of the tuning range as we

[4] Polydoroff, Proc. I.R.E., vol. 21, p. 690; May, (1933).

get at the low-frequency limit. This is a very real defect and requires the use of at least three stages in reactance tuning where two stages could be made to give as good results if the defect were overcome. Furthermore it is not possible to make coils in the band of frequencies between 500 and 1500 kilocycles which have an equivalent value of L/R as great as 6.5×10^{-5}.

The formulas for evaluating the form of the selectivity curve over the audio side band and the amplification per stage are relatively simple if we make the assumption that effects caused by plate-to-grid capacity are negligible.

In the practice of reactance tuning the value of L/R varies from about 1.3×10^{-5} at 500 kilocycles to about one ninth of this value or 1.45×10^{-6} at 1500 kilocycles which is the high-frequency limit of the American broadcast band.

For the short-wave bands above 1500 kilocycles the lack of selectivity due to this inherent decrease in the value of the time constant L/R becomes so important that most short-wave sets which do not use the superheterodyne principle are forced to use a regenerative circuit to reduce the value of R by interstage reaction. This method of reduction of R is unfortunately limited to one stage due to inherent instability introduced when we try to use this method for more than one stage in a train of stages. We know from experience that attempts to get all of our selectivity from a single stage tends to audio distortion but this is the only method remaining to get adequate selectivity in the high-frequency channels.

An indicated method to overcome this difficulty is to confine the control of R to the circuit involving the R which we desire to control. This is possible with resistance tuning. J. R. Nelson[5] has worked out these formulas for reactance tuning (Appendixes II and IV). It will be found when we apply the numerical values used in practice with shield grid tubes that the effects of plate-to-grid capacity are unimportant.

Later in this article similar formulas are worked out for resistance tuning using the same methods employed by Nelson.

Resistance tuning allows us to maintain a constant selectivity curve over the tuning range and also gives us very much greater amplification per stage. This increase in amplification might cause plate-to-grid capacity effects to become important even with shield grid tubes so that it might be necessary to use one of the numerous well-known bridge means for eliminating plate-to-grid capacity effects. For sake of simplicity in developing the formulas the assumption is made, however,

[5] Nelson, Proc. I.R.E., vol. 20, p. 1203; July, (1932).

that the shielding is adequate to prevent appreciable interstage reaction.

With "resistance tuning" adequate performance might be attained with two stages only, involving a saving in apparatus and space.

With resistance tuning we shall also find upon further analysis that the tuning range, instead of being limited to about 3 to 1 in frequency without changing apparatus (as is the case in present reactance tuned sets), may be extended to cover all frequencies below a maximum frequency of about 1500 kilocycles. This figure is dependent on how low a stable source of negative resistance the tube makers can furnish us.

Resistance Tuning

Single Stage Theory

The "J" transformation of "reactance" tuning which gives us resistance tuning, reduced to its simplest terms, is shown in Fig. 1. The transformation is accomplished by the conventional "bridge" circuit, with particular relations existing between the impedances forming the bridge network.

It will be noted that the "galvanometer" branch is short-circuited and that the battery branch contains a variable resistance, R_3, which adjusts the frequency to which the circuit exhibits maximum response or is self-oscillatory.

Two of the bridge arms between one "battery" branch terminal and the short-circuited galvanometer branch, consist of the negative resistance $-R_p$ of a dynatron forming one arm and the inherent tube capacity C, which shunts it, forming the other arm.

The two remaining bridge arms between the short-circuited galvanometer branch and the other battery branch terminal, consist of a fixed inductance shunted by a fixed resistance.

To bring about the "J" transformation and obtain a "resistance" tuned self-oscillatory circuit, we put in the following particular relations in the bridge arms:

Make the value of the fixed positive resistance, R_1, which shunts the inductance, equal to the negative resistance of the dynatron $-R_p$, and make the fixed inductance L in henrys equal to the square of this resistance multiplied by the value of the above mentioned capacity C in farads.

If we now calculate (see Appendix III) the value of the impedance through the bridge, between the terminals of the battery branch, we shall find that all reactance terms cancel, leaving two "effective" resistance terms in series, one of which is positive and proportional to the impressed frequency and the other of which is negative and inversely proportional to the impressed frequency and at a frequency of $1/(2\pi\sqrt{LC})$ these two terms are equal and of opposite signs so that they cancel. If the frequency is less than $1/(2\pi\sqrt{LC})$ the resultant impedance through the bridge is an increasingly effective negative resistance, reaching a maximum of $-R_1$ at zero frequency. If we now complete the circuit through a variable resistance R_3 in the battery branch, the circuit will be self-oscillatory at a frequency which causes the effective negative resistance through the bridge to equal the value to which the positive resistance R_3 in the battery branch is adjusted. Varying this adjustable resistance between the values of zero and R_1 adjusts the frequency of the self-oscillation between the values of $1/(2\pi\sqrt{LC})$ and zero.

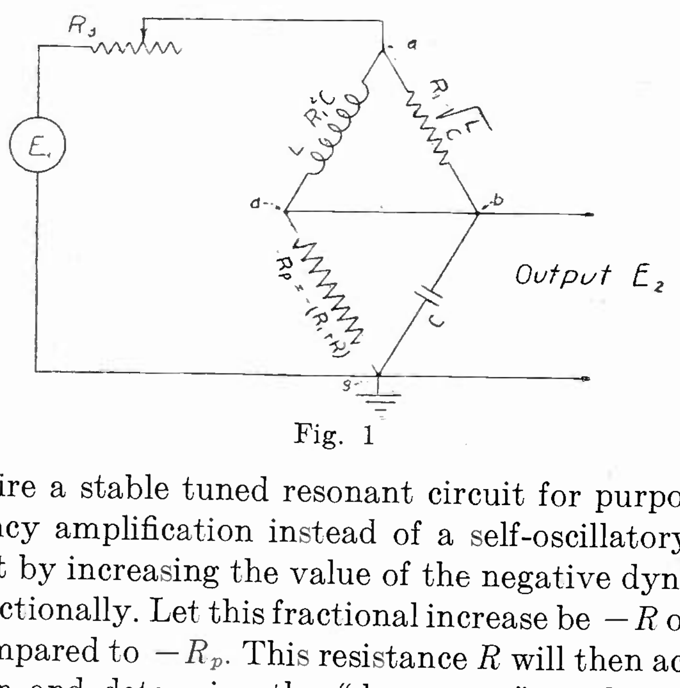

Fig. 1

If we desire a stable tuned resonant circuit for purposes of tuned radio-frequency amplification instead of a self-oscillatory circuit, we may obtain it by increasing the value of the negative dynatron resistance $-R_p$ fractionally. Let this fractional increase be $-R$ ohms a quantity small compared to $-R_p$. This resistance R will then act to prevent self-oscillation and determine the "decrement" in the same manner as in the reactance tuned circuit.

In evaluating the formula for calculating the practical engineering performance of this "resistance tuned" resonant circuit, we may greatly simplify them by using the same approximations used by Nelson previously referred to, which make the assumption that the points of engineering interest lie close to the resonant frequency and that cycles "off resonance" are small compared to the resonant frequency.

With these assumptions, the formula for the impedance Z around the resonant circuit takes the form

$$Z = \frac{2m_0 R}{(1+m_0{}^2)^2}[y - j]. \tag{1}$$

The factor y has the same value $4\pi fL/R$ as previously deduced in the analysis of reactance tuning, f being cycles "off resonance" (plus or minus).

The factor m_0 assumes a knowledge of the desired resonant frequency and is the ratio of this frequency to the maximum frequency $1/(2\pi\sqrt{LC})$ to which the circuit can be tuned by short-circuiting the resistance R_3 in the battery branch.

If a knowledge is desired of the resistance R_3 to put in the battery branch to produce a desired frequency ratio m_0 we may calculate it from the approximate relation $R_3 = R_1(1 - m_0^2/1 + m_0^2)$.

Points of interest to note are as follows:

1. m_0 must be less than 1.
2. R_3 must be less than R_1.
3. $2m_0/(1+m_0^2)^2 =$ approx. 0.5, between m_0 values ranging between 1 and 0.25.
4. At resonance ($f=0$) the impedance Z equals approximately $-j.5R$ within the above-mentioned range of values of m_0.

The engineering performance, or ratio of output to input voltage of the circuit, may be best, for purposes of calculation and comparison with the reactance tuned circuit previously analyzed to be put in the form of $A_0(1/y - j)$.

The factor $1/y - j$ which determines the selectivity has the same effective value as $1/1+jy$ previously analyzed in connection with reactance tuning, and thus needs no further comment.

The factor A_0 which gives the ratio of output to input voltage at resonance depends upon where the electromotive force is impressed in the circuit and upon where the output terminals are connected to the circuit. If we assume a voltage E_1 is impressed directly in series with the battery branch, it will set up a current around the resonant circuit equal to this voltage divided by the impedance of the resonant circuit at resonance ($y=0$). This impedance $Z_0 = -j2m_0R/(1+m_0^2)^2$.

The output voltage E_2 is the product of this current E_1/Z_0 and the impedance in the resonant circuit at resonance between the two points to which the output terminals are connected. Let this impedance be Z_{20}.

The ratio of output to input voltage at resonance $E_2/E_1 = A_0$ which gives the amplification per stage at resonance will thus be equal to Z_{20}/Z_0. This is the principal factor of interest in engineering calculations since the performance is specified by $A_0 = 1/\sqrt{1+y^2}$.

For proper operation of the circuit it is essential that the impedance in the battery branch remain a pure resistance, independent of frequency. Now means for impressing an electromotive force in series with a pure resistance are not obvious. We may, however, arrange to impress an electromotive force in series with one of the branches of any "mesh" in the resonant circuit, consisting of a pure resistance usually shunted to a fixed capacity. Let the impedance of this mesh at resonance consisting of two branches in parallel be represented by Z_{im0} and the impedance of the branch of this mesh in series with which the electromotive force is impressed be represented by Z_{ib0}.

The "reciprocity" theorem furnishes a short cut in figuring the ratio of output to input voltages or A_0 in which we are interested.

In applying it we assume an impressed voltage E_1 in series with R_3 and calculate the value of the current I_1 around the resonant circuit $I_1 = E_1/Z_0$. We next calculate the value of the voltage which the current develops across the terminals of the input mesh $E_1 Z_{im0}/Z_0$. We then divide this voltage by the impedance of the input branch to obtain the current in the input branch.

By the reciprocity theorem we now reverse the positions of E_1 and I_2 bringing the voltage E_1 into the branch in which it is actually impressed and calculate the product of I_2 and the impedance through which it flows in the resonant circuit Z_{20} to get the value of the output voltage, $E_2 = I_2 Z_{20} = E_1 Z_{im0} Z_{20}/Z_0 Z_{ib0}$.

This gives us the value of the amplification per stage

$$\frac{E_2}{E_1} = A_0 = \left(\frac{Z_{im0}}{Z_{ib0}}\right)\left(\frac{Z_{20}}{Z_0}\right). \tag{2}$$

This determines the performance of the circuit when multiplied by the selectivity factor $1/y - j$ before mentioned.

Resistance Tuned "Hook-Up"

As an illustration of a possible application of resistance tuning in its simplest form Fig. 2 shows a two-stage system of "Resistance Tuned" radio-frequency amplification.

Apparatus

The first stage consists of an antenna C_a, associated with a resistance tuned resonant circuit $g-db-a-a'-d'b'-g$. Voltage picked up from the antenna develops a current around this resonant circuit and causes a greatly magnified voltage to exist across its output terminals $b'-g$. Resistance effects and decrement in this resonant circuit may be governed by adjustment of the negative resistance of the "dynatron" tube No. 1 and is accomplished by varying the electron emission by variation of the adjustable resistance R_2 which forms an adjustable "C bias." The apparatus also comprises a fixed inductance L_a and two equal fixed resistances R_a which serve to keep the value of the impedance between the points $a-g$ a pure resistance of value R_a ohms at all frequencies.

A fixed resistance R_1, a fixed inductance L, a variable resistance R_3 by which the circuit is tuned, together with a radio-frequency choke and by-pass condenser complete the list of apparatus appertaining to the first stage.

Apparatus appertaining to the second stage is a duplicate of apparatus in the first stage with the exception that the antenna system between the points $a-g$ is replaced by a resistance of the same value R_a which is fractionally variable by a small amount $(1 \pm \Delta)$ for pur-

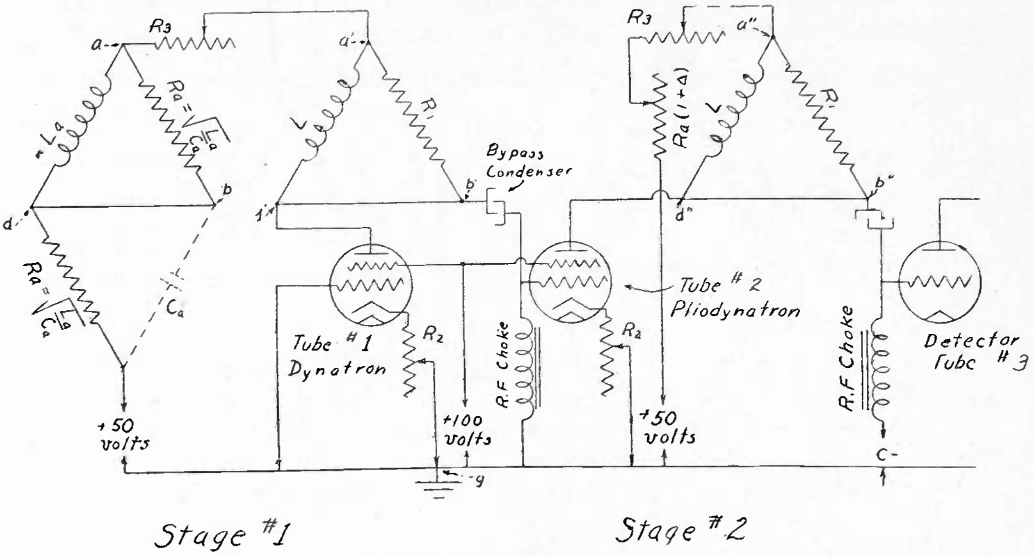

Fig. 2

poses of "trimming" or "staggering" the tuning of the two stages. It is assumed that in tube No. 2, by proper electrode design, an adequate value of mutual conductance may be attained (say of the order of 0.001) and that the polarizing grid provides adequate shielding so the capacity effects between the plate and control grid are sufficiently small to prevent any appreciable amount of the interstage reaction.*

If such a "pliodynatron" is not at present available, it is always possible to use in addition a high-mu shield-grid tube to impress the voltage derived from the first stage upon the second stage and use for tube No. 2 a dynatron similar to tube No. 1.

As an alternative, we might also use the "magnetron" principle of causing the magnetic field of the inductance L in stage 1 to act to control the output tube No. 2.

* *Note*: The development of all the formulas in this paper postulates the absence of interstage reaction.

Apparatus common to the two stages consists of the usual power supply and means for simultaneously varying by a single motion the value of the two equal tuning resistances R_3.

Operation of the Set

We may tune in any desired signal having a lesser frequency than $1/(2\pi\sqrt{LC}) = 1/(2\pi R_1 C)$ by adjusting the gang control which varies the adjustable tuning resistances R_3 in the two stages. We may tune the two stages "in line" or we may "stagger" the tuning to obtain a more rectangular response curve over the side bands by adjustment of the control varying the value of Δ in the variable resistance $R_a(1+\Delta)$ in the tuning circuit of the second stage.

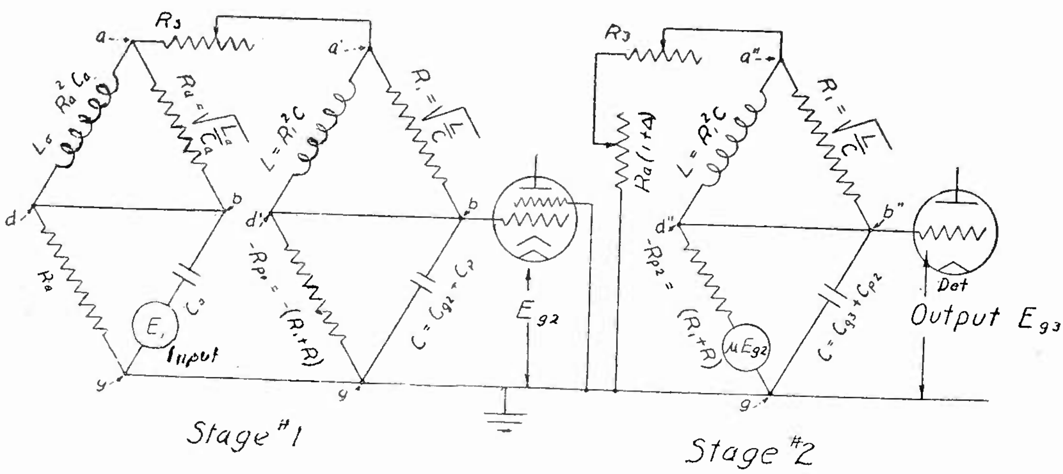

Fig. 3

Control of R in each stage is secured by adjustment of the resistances R_2. This brings all factors necessary to secure optimum performance under control.

Theoretically as the circuit is shown there should be no necessity for varying the adjustment of the resistances R_2 once best adjustment is ascertained as we tune to different frequencies within the tuning range.

In practice, however, due to the fact that the inductances L are not pure inductances as shown but in reality contain a small resistance term dependent on frequency, it would be necessary, at the higher frequencies, to make small fractional changes in the resistances R_2 to keep the value of L/R in two circuits equal and of definite value.

Several different means of compensation of this necessity for change in adjustment will suggest themselves to the trained engineer, but the scope of this article does not permit entering into this question.

Calculation of Voltage Amplification

The ratio of output to input voltage A_0 attained in a resistance tuned resonant stage is given by evaluating (2) for the stage.

Fig. 3 shows the equivalent simplified circuits of Fig. 2 for purposes of making this calculation. The subscripts g and p shown indicate the circuit is taken between grid and ground or between plate and ground respectively. The small number subscripts following g and p indicate the number of the vacuum tube shown in Fig. 3. C_a is the capacity to ground of the antenna.

The calculations will be somewhat simplified if we establish the relation $R_a/R_1 = C/C_a$ since it cancels the effect of antenna capacity out of the formula. (This is not necessary however.)

At the frequency of resonance $m_0/(2\pi\sqrt{LC})$ the impedances entering into (2) will have the following values with the particular relation noted in Fig. 3.

In stage No. 1,

$$\frac{Z_{im0}}{Z_{ib0}} = \frac{jm_0}{1 + jm_0}$$

$$Z_{20} = \frac{-\sqrt{\dfrac{L}{C}}}{1 - jm_0}$$

$$Z_0 = \frac{-jm_0 2R}{(1 + m_0^2)^2}.$$

Thus the voltage amplification Eg_2/E_1 for stage one is given by inserting these values into (2).

$$A_{01} = \frac{Eg_2}{E_1} = \frac{(1 + m_0^2)\sqrt{\dfrac{L}{C}}}{2R} = \left(\frac{1 + m_0^2}{2}\right)\left(\frac{L}{R(LC)^{1/2}}\right). \quad (3)$$

In stage No. 2,

$$\frac{Z_{im0}}{Z_{ib0}} = \frac{1}{1 - jm_0}$$

$$Z_{20} = \frac{-\sqrt{\dfrac{L}{C}}}{1 - jm_0}$$

$$Z_0 = \frac{-jm_0 2R}{(1 + m_0^2)^2}.$$

Thus the voltage amplification Eg_3/Eg_2 for stage two is given by inserting these values into (2).

$$A_{02} = \frac{Eg_3}{Eg_2} = \frac{[2m_0 - j(1 - m_0^2)]\mu\sqrt{\frac{L}{C}}}{m_0 2R}$$

$$= \left(\frac{1 + m_0^2}{2m_0}\right)\frac{\mu L}{R_1 RC} = \frac{(1 + m_0^2)}{2m_0} g_m \frac{L}{RC}. \tag{4}$$

The complete expression giving the voltage amplification for the two stages at resonance is as follows:

$$\frac{Eg_3}{E_1} = \frac{A_{01}A_{02}}{S_{s0}} = \frac{(1 + m_0^2)^2}{4m_0}\mu\left(\frac{L}{RR_1C}\right)^2\left(\frac{1}{S_{s0}}\right). \tag{5}$$

If the two stages are tuned to the same frequency $1/S_{s0}$ will be equal to 1 and $f_s = 0$.

For derivation of the numerical values of L/R and $1/S_{s0}$ if we use stagger tuning refer to Appendix I. The value of $1/S_{s0} = 0.2$ and the value of $L/R = 1/2\pi f_s$ in case we choose the suggested curve to give the optimum compromise between selectivity and audio performance.

Analysis of Formula (5)

Note: (1) That the factor $(1+m_0^2)^2/4m_0$ remains fairly close to unity for values of m_0 between 1 and 0.3 and for values of m_0 smaller than 0.3 the over-all amplification tends to become inversely proportional to the frequency to which the two stages are tuned.

(2) That the factor $1/(R_1C)^2 = 1/LC$, which is the square of maximum periodicity $1/(LC)^{1/2}$ to which the set can be tuned. Assuming that the minimum value of C is fixed by inherent tube capacities (about 20 $\mu\mu$f), the maximum frequency to which the set may be tuned becomes $1/2\pi R_1C$, and the amplification proportional to $\mu/(R_1C)^2 = g_m/R_1C^2$ where g_m is the mutual conductance of the shielded pliodynatron.

(3) Most of the present production tubes may be used as dynatrons to produce a negative resistance, down to a minimum of roughly 15,000 ohms without overloading. Assuming a minimum value for the capacity C of 20 micromicrofarads this would make the maximum frequency to which we can tune the set 530 kilocycles or about the low-frequency limit of the American broadcast band.

Tubes for use as dynatrons have been recently improved by special methods of pumping and use of anode metal having higher secondary emission so that the figure of 15,000 ohms might be reduced to about

5000 ohms which would make the maximum frequency about 1500 kilocycles which would thus include the broadcast band.

The writer has no data with regard to the value of g_m which might be obtained on a shielded pliodynatron as shown for tube No. 2.

If one cannot at present be obtained having a value of g_m as large as 0.001, this difficulty may be overcome by adding to the circuit a shield-grid tube, (many of which have a value of g_m as large as 0.001) with its plate circuit connected in parallel with the plate circuit of a dynatron used for tube No. 2. We would then connect the output of stage 1 to the grid of this new tube. We shall find on analysis in (5) (using above-indicated methods) that there is no change introduced by this.

An alternative method would be to connect the plate circuit of the extra tube across the terminals of the resistance R_1 in stage 2. This has the advantage that if we desire to use the set to tune in the lower frequencies between the values of $m_0=0.3$ and $m_0=0$, that it eliminates the factor m_0 from the denominator of (5) leaving the over-all amplification practically unchanged within this range of m_0.

In this connection it is of interest to note that if we took stage 2 alone, thus changed, and connected a small controlled amount of its output to its input terminals, we would have an oscillator whose frequency could be adjusted from say 50,000 cycles to 50 cycles by adjustment of the resistance R_3. This might have its advantages (due to greatly increased tuning range) as an audio oscillator, for supplying carrier frequencies in "wired wireless," and for general purposes for supplying heterodyne oscillation in long-wave reception.

Design Procedure

The design of resistance tuned radio-frequency amplifiers is much more simple than reactance tuned amplifiers due to the elimination of the coupling transformer when we use resistance tuning. (With reactance tuning, to obtain optimum results, the transformer would have to have a variable ratio of transformation over the tuning range and is not practicable.)

The design might perhaps best be shown by working out the numerical values of resistance, inductance, and capacity to fulfill a specific problem in resistance tuning, as follows:

Problem 1.

Preliminary assumptions: Assume the tube makers can furnish a shielded pliodynatron or magnetron giving a stable negative resistance as low as 5000 ohms with a mutual conductance of 0.001 ($\mu=5$) and also a dynatron giving this same value of negative resistance.

Assume that the value that the plate-to-ground capacity of these dynatrons is 10 micromicrofarads and the grid-to-ground capacity of the pliodynatron and the detector has this same value.

Assume the set consists of a tuned antenna and a second tuned circuit as shown in Fig. 2.

With these assumptions we have established the value of C in both stages of Fig. 3 at $C=2\times10^{-11}$ and the value of R_1 in both stages at $R_1=5\times10^3$. The value of L in both stages must therefore be $L=R_1^2C=5\times10^{-4}$.

The maximum frequency to which the set may be tuned will be approximately (neglecting the resistance R_a), $f_m=1/2\pi R_1C=1600$ kilocycles. Assume we stagger "tune" the two stages by adjustment of the resistance $R_a\Delta$ in stage 2 so that the stages are tuned 5 kilocycles apart regardless of the frequency to which we tune, and assume that we adjust the resistance R in both stages so that $L/R=6.5\times10^{-5}$ from which $R=7.7$ ohms.

This may be done by adjusting the resistances R_2 in each stage until the negative resistances of the dynatrons are -5007.7 ohms each.

With this adjustment, we shall establish the uniform selectivity curve which shows approximate uniformity of amplification over the side bands out to about 3 kilocycles and with a cut-off at 10 kilocycles off resonance of about 10 to 1. (See Appendix I.)

The value of the factor S_{s0} will then be 5. We have now established all the numerical values entering into (3), (4), and (5) and may now use it to determine the over-all amplification of the set.

$$\frac{Eg_2}{E_1}=\frac{1+m_0^2}{2}(6.5\times10^{-5})(1\times10^7)=\frac{1+m_0^2}{2}(650) \tag{3}$$

$$\frac{Eg_3}{Eg_2}=\frac{1+m_0^2}{2m_0}(6.5\times10^{-5})(5\times10^7)=\frac{1+m_0^2}{2m_0}(3300) \tag{4}$$

$$\frac{Eg_3}{E_1}=\frac{(1+m_0^2)^2}{4m_0}(650)(3300)\left(\frac{1}{5}\right)=\frac{(1+m_0^2)}{4m_0}(430{,}000). \tag{5}$$

The factor $(1+m_0^2)^2/4m_0$ does not vary greatly from unity from 1600 kilocycles down to 400 kilocycles for the value of the carrier frequency to which the set is tuned. At carrier frequencies below 400 kilocycles it tends to become inversely proportional to the change in carrier frequency, reaching a value of 16 approximately at carrier frequency of 25 kilocycles.

A voltage amplification of 4.3×10^5 corresponds to a gain of 113 decibels approximately with a set comprising a tuned antenna coupled by a tube to second tuned stage. This gain would probably be considered sufficient for most conditions to be met in radio-frequency amplification. It represents the maximum possible gain we might hope to attain with two tuned circuits subject to the above-mentioned limitations which assume no appreciable interstage reaction and grid-to-ground conductance in tubes No. 2 and 3.

If we desire to compare the results of resistance tuning with those which might be attained with reactance tuning under the same conditions, the necessary data are given in Appendix IV.

If we use a fixed value of L without negative resistance control R will tend to vary more or less proportionately with the square of the proportional change in frequency. If we assume a 3-to-1 change in frequency, the value of y will be one ninth as large at the high-frequency limit of the range as it is at the low-frequency limit. This would make it necessary to go about 90 kilocycles off resonance to attain the same voltage attenuation at the high-frequency end of the range as we would get with 10 kilocycles off resonance at the low-frequency limit of the range. This is the largest inherent trouble with reactance tuning and attention has been called to it by Nelson in the first paragraph of his article.[5]

Appendix I

Assuming that all stages are tuned to the same frequency and also that the value of the time constant L/R in each stage is the same, this curve is obtained by plotting the function

$$s = (1 + y^2)^{n/2}$$

where,

n = number of stages
f = cycles off resonance
L = inductance in the oscillating circuit
R = resistance in the oscillating circuit
$y = 4\pi fL/R$ (see Appendixes II and III).

If we use two stages tuned to slightly different frequencies, we shall get a more rectangular curve than with three stages tuned to the same frequency. This is sometimes called "stagger" tuning.

Stagger Tuning

Stagger tuning[5] in its simplest form, consists in tuning two consecutive stages, one above and the other below the center of the audio band of frequencies it is desired to receive. Let this number of cycles $= f_s$. The combined selectivity factor of the two stages will then be given by the expression

[5] *Loc. cit.*, Fig. 8, curve *C*.

$$S_s = (y_1^2 + 1)^{1/2}(y_2^2 + 1)^{1/2}$$

where,

$$y_1 = 4\pi \frac{L}{R} f_s\left(\frac{f}{f_s} - 1\right)$$

and,

$$y_2 = 4\pi \frac{L}{R} f_s\left(\frac{f}{f_s} + 1\right).$$

In general we would make f_s approximately equal to the audio frequency away from the center of the audio band at which we desire the attenuation of voltage amplification to start. The smaller we make f_s, the greater will be the selectivity, and due to decreased band width, the greater will be the ratio of signal to static. Also the greater will be voltage amplification for the two stages.

Its lower limit is governed by too great voltage attenuation in the audio frequencies furthest away from the center of the audio band which brings about undue audio distortion of the reception. A fair compromise value for f_s might be about 2500 if we desire to receive both side bands or about 1250 if we desire to receive only one side band.

In order to give this "staggered tuning" a chance to produce optimum results, it is primarily necessary that the curve of voltage amplification over the audio frequencies be kept symmetrical on each side of the middle. This can only be accomplished if we keep the value of L/R in each stage the same and also render interstage reaction negligible. These are the two essentials of primary importance if we wish to obtain a maximum possible ratio of signal to static combined with a maximum possible selectivity for an adequate audio performance.

In the present sets which tune over a specified range by variation of capacity there is no provision for keeping L/R constant over the tuning range. With tuning accomplished by variation of resistance, the means are provided.

In both cases it is assumed that interstage reactions and grid-to-ground conductance are negligible, and the following analysis postulates that this is so.

To determine what value of L/R gives the most promising curve of voltage attenuation over the audio range, it is necessary to plot the curve of S_s using as abscissas the variable f/f_s between the values from $+10$ to -10. (If we plot three of these curves using the values 1, 2, and 3 for the value of the quantity $4\pi(L/R)f_s$ we should probably select

$4\pi(L/R)f_s=2$ as giving the most promising shape of general curve of voltage attenuation over the audio range, and being superior in selectivity with approximately the same audio distortion to the curve for three stages tuned without "stagger."

If we desire to receive both side bands with $f_s=2500$ this fixes the best value of $L/R=1/2\pi f_s=6.5\times10^{-5}$, a quantity which is independent of the specified carrier frequency.

Stagger tuning has probably been in use consciously or unconsciously since about 1925 to a certain extent since most sets contain an instrumentality which consists in tuning two or more stages with a single tuning handle ("gang tuning") and then providing an independently adjustable tuning means ("trimmer condensers") by which the amount of the "stagger" is controlled.[6]

It will be found on analysis that a more rectangular curve which gives a better compromise between selectivity and audio performance may be obtained with two stages staggered than with three stages tuned to the same frequency. We should thus choose two stages "staggered" in place of three stages tuned to the same frequency provided the over-all amplification attained in two stages is adequate for our purpose.

To obtain the shape of the curve over the audio bands to be received, we may plot S_s in units of f/f_s as abscissas. We then divide the ordinates by the value of S_s at $f=0=S_{s0}=(4\pi(L/R)f_s)^2+1$. This curve will then show "relative voltage input for standard output," and is somewhat more convenient for inspection. It may be directly compared with a similar curve shown by Nelson[5] which gives results of actual measurements taken from a three-stage set.

If we assume that the best shape of curve is obtained by making the above-mentioned value of $4\pi(L/R)f_s=2$, and furthermore that the best value for f_s, if we desire to receive both side bands, is 2500, we may make a direct comparison between Nelson's measurements on a three-stage stagger tuned receiver and a two-stage receiver having a value of $L/R=6.5\times10^{-5}$ in both circuits which are tuned 5000 cycles apart.

This comparison shows approximately as follows:

Relative Input for Standard Output

	$f=0$	$f=\pm2500$	$f=\pm5000$	$f=\pm10{,}000$
Nelson, Fig. 8 (three-stage measurements)	1	1.3	2	6
Calculated three-stage tuned without "stagger" $L/R=1.2\times10^{-6}$	1	1.3	2	6
Calculated two-stage, 5000-cycle stagger ($f_s=2500$) $L/R=6.5\times10^{-5}$	1	0.8	3	10

[6] Cabot, U. S. Patent 1,545,940, July 14, 1925.

Appendix II

The "selectivity factor" $1+jy$ is the number by which we multiply the impedance at resonance $=Z_0$ taken around the tuned resonant circuit to obtain the value of this impedance at a specified number of cycles off resonance $= \pm f$

$$Z = R + j\left(L\omega - \frac{1}{C\omega}\right) \tag{1}$$

let,

$$m = \omega(LC)^{1/2}$$

$$Z = R + j\left(\frac{L}{C}\right)^{1/2}\left(m - \frac{1}{m}\right) \tag{2}$$

let,

$$f = \frac{1}{2\pi(LC)^{1/2}} = \text{resonant frequency}$$

$$m = 1 \pm \frac{f}{f_0} = 1 \pm 2\pi f(LC)^{1/2} = 1 \pm \delta.$$

If δ is small compared to 1,

$$-\frac{1}{1+\delta} = -1 + \delta, \text{ approximately}.$$

Substituting this value in (2)

$$Z = R + j\left(\frac{L}{C}\right)^{1/2}(\pm 2\delta) = R\left[1 \pm j4\pi\frac{L}{R}f\right].$$

Let,

$$y = \pm 4\pi\frac{L}{R}f$$

$$Z = R(1 \pm jy)$$

and,

$$\overline{Z} = R(1 + y^2)^{1/2}$$

$$Z_0 = R.$$

Near resonance the impedance across the terminals of the capacity C is equal to $-j(L/C)^{1/2}$ (approximately), from which it follows if we impress a voltage E_1 in series in the resonant circuit and measure the voltage E_2 produced by resonant rise across the terminals of the capacity C that the ratio of these voltages

$$\frac{E_2}{E_1} = \frac{-j\left(\frac{L}{C}\right)^2}{Z} = -j\frac{L}{R(LC)^{1/2}}\left(\frac{1}{1+jy}\right) = A_0\left[\frac{1}{(1+y^2)^{1/2}}\right].$$

For the derivation of this value of y in a transformer coupled resonant circuit, refer to J. R. Nelson.[5] If we assume that we are interested only in that portion of the selectivity curve which covers the audio frequencies to be received and that the total range of audio frequencies to be received is small compared to the carrier frequency so that the approximation $(1+2\delta)=(1+\delta)^2$, we may write Nelson's formula (6)

$$y = \pm\frac{\omega_0 L 2\delta}{R + \dfrac{\omega_0{}^2 L^2}{T^2 R_p}}.$$

The denominator of this expression will be found upon analysis to be the value in ohms of the effective resistance in the generalized oscillating circuit. Let us call this R instead of calling R the resistance of the transformer secondary. We then have

$$y = \pm\frac{\omega_0 L 2\delta}{R}.$$

But,

$$\pm\,\delta = \frac{\omega - \omega_0}{\omega_0} = \frac{2\pi f}{\omega_0},\quad \text{where } f = \text{cycles} \pm \text{“off resonance.”}$$

This gives

$$y = \pm\, 4\pi\frac{L}{R}f$$

which is identical to the value of y in the simple resonant circuit above analyzed.

Appendix III

The impedance around the battery branch R_3 and the bridge circuit $a-db-g$ (shown in Fig. 1) taken in series with the battery branch may be calculated as follows:

Let,

$$\omega = 2\pi x \text{ impressed frequency}$$

$$m = \omega(LC)^{1/2}$$

$$R_3 = K_3\left(\frac{L}{C}\right)^{1/2}$$

K = an arbitrary multiplier dependent on adjustment

$$R = \Delta\left(\frac{L}{C}\right)^{1/2}.$$

The impedance between the points a and db will then be equal to

$$\left(\frac{L}{C}\right)^{1/2}\left(\frac{jm}{1+jm}\right).$$

The impedance between the points db and g will be equal to

$$\frac{-\left(\frac{L}{C}\right)^{1/2}(1+\Delta)}{1-jm(1+\Delta)}.$$

The impedance of these two meshes in series (or between the points a and g) will then be

$$Z_{(a-g)} = \left(\frac{L}{C}\right)^{1/2}\frac{[m^2(1+\Delta)-(1+\Delta)-j\Delta m]}{m^2(1+\Delta)+1-j\Delta m}.$$

The total impedance $Z_{(a-g)}+R_3$ will then be

$$Z = \left(\frac{L}{C}\right)^{1/2}\frac{[m^2(1+\Delta)(K_3+1)-(1+\Delta)+K_3-j\Delta m(K_3+1)]}{m^2(1+\Delta)+1-j\Delta m}. \qquad (1)$$

The impedance Z will be a minimum when the value of m^2 in the numerator is such as to cause the real term in the numerator to equal zero. Let this value of $m^2 = m_0{}^2$.

$$m_0{}^2 = \frac{1-K_3+\Delta}{(1+K_3)(1+\Delta)}.$$

Let,

$$m^2 = \omega^2 LC = m_0{}^2(1+K\Delta)^2$$

when,

$$K\Delta = \text{ratio of cycles "off resonance" to resonant frequency} = \frac{f}{f_0}$$

(K being an arbitrary multiplier)

$$\Delta = R\left(\frac{C}{L}\right)^{1/2}.$$

If we substitute $m_0^2(1+K\Delta)^2$ for m^2 in (1), we get

$$Z=\Delta\left(\frac{L}{C}\right)^{1/2}$$

$$\frac{\left[(2K+K^2\Delta)(1-K_3+\Delta)-j(1+K\Delta)(1+K_3)\left(\frac{1-K_3+\Delta}{(1+K_3)(1+\Delta}\right)^{1/2}\right]}{\frac{(1-K_3+\Delta)(1+K\Delta)^2}{1+K_3}+1-j\Delta(1+K\Delta)\left[\frac{1-K_3+\Delta}{(1+K_3)(1+\Delta)}\right]^{1/2}}. \quad (2)$$

In evaluating (2) we are principally interested in its value over the range of audio frequencies we desire to cover. Under these conditions the value of $K\Delta = f/f_0$ will be very small compared to 1 and we may drop all terms containing Δ from the denominator and from the numerator (inside the brackets) without causing any important error in engineering calculations.

This gives us

$$Z = R\left\{\frac{(1+K_3)^2}{2}\left(\frac{1-K_3}{1+K_3}\right)^{1/2}\left[2K\left(\frac{1-K_3}{1+K_3}\right)^{1/2} - j\right]\right\} \quad (3)$$

but,

$$\left(\frac{1-K_3}{1+K_3}\right)^{1/2} = m_0 \text{ (approximately)}$$

and,

$$\frac{(1+K_3)^2}{2} = \frac{2}{(1+m_0^2)^2}.$$

Substituting these values in (3)

$$Z = \frac{2m_0R}{(1+m_0^2)^2}(2Km_0 - j). \quad (4)$$

The frequency f_0 of resonance at which Z in a minimum as before mentioned is

$$f_0 = \frac{m_0}{2\pi(LC)^{1/2}}$$

from which,

$$\frac{f}{f_0} = K\Delta = \pm \frac{2\pi f(LC)^{1/2}}{m_0}$$

but,

$$\Delta = R\left(\frac{C}{L}\right)^{1/2}, \text{ from which } 2Km_0 = \pm 4\pi \frac{L}{R} f = y.$$

Note that $4\pi fL/R$ is the same as the value of y previously determined in the reactance tuned circuit (Appendix I) which interpreted means that the selectivity factor remains unchanged and has its value $1+jy$ in the reactance tuned circuit and $-j(1+jy)=y-j$ in the resistance tuned circuit.

We may then write the equation for Z over the audio frequencies near resonance,

$$Z = \frac{Zm_0R}{(1+m_0^2)^2}(y-j). \tag{5}$$

The expression for Z at the frequency of resonance where $f=0$, $K=0$, and $y=0$ is

$$Z_0 = \frac{-jZm_0R}{(1+m_0^2)^2}.$$

Appendix IV

Reactance Tuning

J. R. Nelson[5] has analyzed the factor A_0 for the usual stage of high-mu shield-grid input tube, which is transformer coupled to a resonant output circuit.

We may simplify Nelson's expression considerably by assuming that the transformer has the optimum ratio of transformation.

Let,

R = effective resistance in oscillating circuit
R_2 = effective resistance of transformer secondary
$R_2 = R/2$
μ = amplification factor of tube
R_p = plate resistance of tube
L = inductance of secondary
C_0 = capacity of tuning condenser at resonance
T_0 = optimum ratio of transformation.

The factor

$$A_0 = 0.7\mu\left[\left(\frac{1}{R_p}\right)\left(\frac{L}{R}\right)\left(\frac{1}{C_0}\right)\right]^{1/2} \tag{1}$$

$$= 0.5\mu T_0.$$

The ratio output to input voltages over the side bands is given by $A_0(1/1+jy)$ where $y=4\pi(L/R)f$ and where $1+jy$ is the selectivity factor before referred to. (See Appendix II.)

If we let m_0 equal the ratio of the desired resonant frequency to the maximum frequency to which we can tune and let C equal the minimum value of the tuning capacity, we may write $C_0=C/m_0^2$ and formula (1)

$$A_0 = 0.7\mu m_0\left(\frac{L}{R_pRC}\right)^{1/2}.$$

Let,

$$g_m = \frac{\mu}{R_p} \text{ and } A_0 = 0.7m_0\left(\frac{\mu g_m L}{RC}\right)^{1/2}.$$

Over the tuning range the value of m_0 will vary from 1 at the highest frequency to about 0.3 at the lowest frequency. With L/R constant, this will cause variation of 10 to 1 in amplification for two stages over the tuning range. This variation is too large to be tolerated. It will be found that this difficulty may be overcome by using "resistance tuning" in place of "reactance tuning."

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QUARTZ CRYSTAL CONTROLLED OSCILLATOR CIRCUITS*

By

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Summary—*A brief account is given of the developments conducted previous to 1930 which resulted in improved frequency stability for quartz crystal controlled oscillators.*

An improved oscillator circuit is shown with the author's theory of operation and both laboratory and operating data on its performance.

Commercial equipment using this circuit has demonstrated consistent stability of better than 10 parts in 10^6.

Much of the work done on quartz crystal controlled oscillators has had for its purpose the improvement of the frequency stability of these circuits. However, an examination of the literature will reveal the fact that few changes have been made in the circuits themselves since 1923. The reason for this is that the conditions imposed upon the crystal controlled oscillator when it was put to work competing with high power self-excited oscillators were so obviously unfavorable to a high degree of frequency stability that it has been found more economic to better these conditions than to experiment with new circuits, that is, until 1930.

At that time, the crystal controlled oscillators used in broadcast transmitters built by the General Electric Company were operating under the following conditions:

1. The quartz plate was mounted in a Monel-metal cell in which the top electrode was accurately spaced from the bottom electrode, or anvil, by three spacers cut from the same slab of quartz as the plate itself. Metal pins in the anvil restricted the lateral and longitudinal motion of the quartz plate. The effect the vibrating or tilting of this holder had upon the frequency was usually less than one part in a million. The temperature frequency coefficient of the crystal and holder was about twenty parts in a million per degree centigrade.
2. The holder was mounted in a crystal heater box which held the temperature within plus or minus one-fifth degree centigrade of its correct operating temperature under ordinary operating conditions.

* Decimal classification: R355.65. Original manuscript received by the Institute, September 20, 1933.

3. The oscillator circuit used was the inductance-loaded type using the UX-210 tube which was operated with a plate voltage of about 200 volts.
4. The load on the oscillator was held practically constant by means of a two-stage buffer amplifier using UX-865 tubes operated with a plate voltage of about 500 volts.

These oscillators were incorporated in an oscillator-amplifier unit which was named the 0A-1A unit. This unit was used in all broadcast transmitters from the 100-W, a 100-watt set, to the 50-B, a 50-kilowatt set. These transmitters had a frequency stability of plus or minus 50 cycles at any frequency in the 550- to 1500-kilocycle range.

It was observed that the carrier frequencies of some of these transmitters drifted gradually during a day's operation. The change in temperature indicated by the crystal thermometer was not large

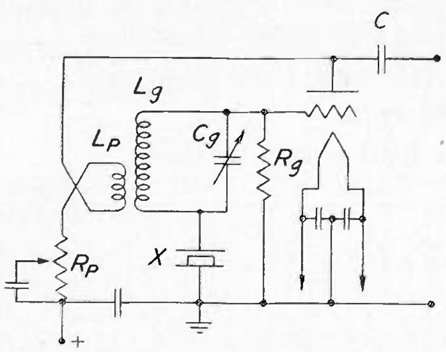

Fig. 1—Oscillator circuit.

enough to account for this drift. The effect was found to be greatest on days of rapidly changing room temperature, consequently, it was judged to be some kind of temperature effect.

It was assumed that the heat dissipated in the quartz plate was the cause of this frequency drift. Consequently, the search for a circuit which would provide sufficient excitation for the two-stage buffer amplifier without causing so much heat to be dissipated in the quartz plate was started.

The result of this search was the circuit shown in Fig. 1, and the result of substituting this circuit for the inductance-loaded circuit in a standard 0A-1A unit used in the WGY transmitter is shown in Fig. 2. This startling reduction in frequency drift was caused by the circuit change alone.

Since the unit in the WGY transmitter was being used on regular programs eighteen hours a day, another 0A-1A unit was modified by installing the circuit of Fig. 1, and considerable data were taken at

1500 kilocycles, the top of the broadcast range. The conditions and results of these tests follow:

Tubes	1—UX-210 as oscillator
	2—UX-865 as buffer-amplifiers
E_b	210 volts (normally)
Crystal	Type 1A-522, Y cut.
L_p	81 microhenries
L_g	386 microhenries
R_p	2000 ohms
R_g	1 megohm
C (coupling)	40 micromicrofarads.
T_x	$56.5°C \pm 0.2°C$

Test	Resulting change cycles per second
Tuning grid circuit 1122–1451 kilocycles	150.0
Tuning 1st buffer plate 548–1580 kilocycles	2.0
Tuning 2nd buffer plate 548–1580 kilocycles	0.5
Changing E_b 190–230 volts	1.0
Changing 2nd buffer load 0–5.5 watts	1.0

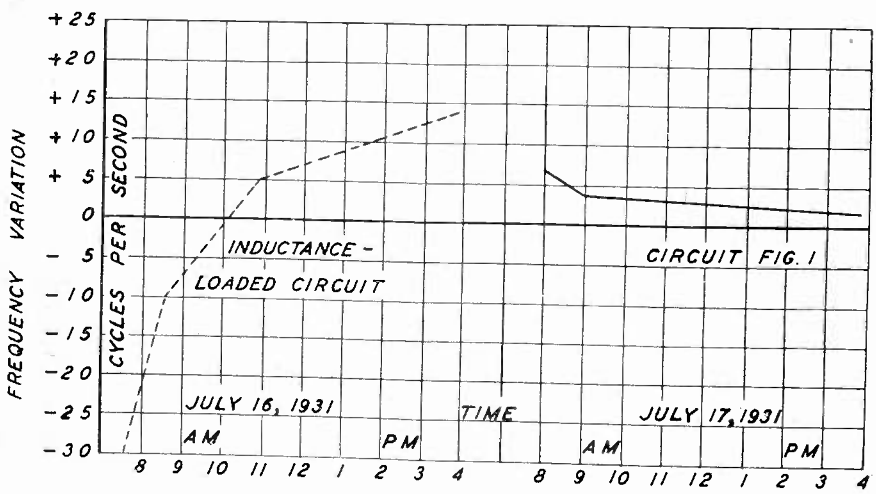

Fig. 2—Characteristic operation—790 kilocycles.

These data show that a change of 2193 cycles per second in the uncoupled frequency of the grid tuning circuit caused a change of only one cycle in the controlled frequency.

Professor Terry[1] found that a change of 961 cycles per second in the uncoupled frequency of the plate circuit in a tuned-plate circuit type of crystal controlled oscillator caused a change of one cycle per second in the controlled frequency.

[1] E. M. Terry, "Frequency of piezo-electric oscillators," Proc. I.R.E., vol. 16, p. 1486; November, (1928).

MacKinnon[2] found that a change of 20 per cent in plate voltage on the tuned-plate type of crystal controlled oscillator caused a change in frequency of from 2 to 5 cycles per second in 10^6, and that the crystal controlled dynatron changed frequency from 0.5 to 2 cycles per second in 10^6 for a 20 per cent change in screen voltage.

These data show that the circuit of Fig. 1 compares favorably with the other circuits on the two points mentioned.

Its performance in the WGY transmitter since July 17, 1931, which shows that on a large percentage of the days, the frequency drift during the day is one cycle per second or less proves how favorably it compares with other circuits for over-all stability under regular broadcast operating conditions.

This circuit has been applied to both the new RCA Victor broadcast frequency control equipment type EX-4170 and the broadcast monitor type EX-4180.

The frequency control equipment is a complete radio-frequency generator with an output of about five watts and a frequency stability of better than ten cycles per second in 10^6. The equipment consists of the following units:

(1). A rectifier using two hot-cathode mercury-vapor tubes in a single-phase full-wave circuit.
(2). A crystal oscillator using an RCA-843 tube and two precision type holders (one as spare).
(3). Two crystal heater boxes capable of holding the crystal temperature within ± 0.1 degrees centigrade of its correct temperature under all reasonable operating conditions. (These mount in (2).)
(4). Two buffer amplifiers each using an RCA-844 tube in a tuned plate buffer-amplifier circuit.
(5). A rack on which the separate units are mounted.
(6). Three blank panels.

When the first of the broadcast monitors was tested, the room temperature was varied between the limits of 14 and 45 degrees centigrade, the line voltage was varied from 107 to 120 volts, and the oscillator tube was changed in the middle of the test. The largest change in frequency was that caused by changing the tube; it was twelve cycles per second, and after one hour of operation the frequency had returned to within two cycles per second of where it was before the tube was changed. Except for this, the maximum deviation from 1499.5 kilo-

[2] K. A. MacKinnon, "Crystal control," Proc. I.R.E., vol. 29, p. 1689; November, (1932).

cycles was nine cycles per second, and this occurred after the crystal had been cooled down to 22 degrees centigrade from its operating temperature of 57 degrees centigrade in the middle of the run and reheated. The average daily change for the seven-day run was five cycles per second.

Theory

It is the opinion of the author that the circuit of Fig. 1 operates in the following manner:

(a). Energy is fed back through the plate-to-grid capacity of the tube in the phase which will sustain either oscillations of the type found in the inductance-loaded circuit (considering the crystal and the plate coil only) or those of the tuned-plate tuned-grid type (considering the tuned-grid circuit and the plate coil only).

(b). Energy is fed back through the magnetic coupling of the plate and grid coils in the phase which will neither sustain oscillations of the types mentioned in (a) nor those of the type found in the Armstrong regenerative circuit.

(c). The resistance in series with the plate coil suppresses the tendency toward self-excitation of the tuned-plate tuned-grid type. (This type of oscillation cannot be easily stopped by degenerative magnetic feed-back at the higher frequencies unless electrostatic shielding is provided between the two coils.)

(d). The fact that the tuned grid circuit is in series with the crystal makes it possible to have the same grid-to-filament voltage swing on the oscillator tube that is used in other circuits and have only part of it across the crystal. This naturally results in less energy dissipation in the crystal than is found in other circuits.

Conclusions

Although this crystal controlled oscillator circuit is not perfect, it has made the increased frequency stability which was the object of this development an actuality. It is believed to be the best circuit available at the present time for use in the broadcast frequency spectrum in applications requiring good frequency stability and economic design. It is believed to be suitable for use in laboratory frequency standardizing and measuring equipment, also.

It seems clear that the developing of the circuit of Fig. 1 has helped advance the art of controlling the frequency of electromagnetic waves.

Acknowledgment

The author wishes to make known his indebtedness to Messrs. C. A. Priest, A. B. Tripp, W. A. Ford, F. J. Moles, and R. L. Downey for their hearty coöperation in the search for better means for controlling the frequency of electromagnetic waves.

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ELECTRODYNAMIC SPEAKER DESIGN CONSIDERATIONS*

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Summary—*The operation of the diaphragm is shown to be so involved as to make its development difficult with any treatment other than an empirical one. Measurements, however, can be made upon a diaphragm to determine its efficiency and to determine the optimum mass of the voice coil to be associated with it. With the voice coil mass so determined, a method is shown for selecting the voice coil diameter and the magnet dimensions to give a desired performance at a minimum cost. The design procedure is applicable not only to electrodynamic speakers but is of use also in the design of any device utilizing an electrically-driven vibramotive force.*

IT IS traditional that the early forms of devices making use of new principles or phenomena are, for the most part, based upon practical expediency or empirical findings. The development of the "cone" type electrodynamic loud speaker makes no exception to this general rule. It can be safely stated that, even at the present time workers in this field depend almost entirely upon "cut-and-try" methods to achieve a desired result. The reason for the continued use of empirical methods lies in the difficulty encountered in reaching an exact mathematical analysis. Such an analysis would be extremely involved, if not impossible, because of the continuous change in mode of vibration of the radiating surface with respect to frequency. Experimental data concerning a particular diaphragm may be easily acquired, however, and coupled with a knowledge of the manner in which the device functions, this information can be readily utilized in establishing the performance of a given design.

The Diaphragm

Of fundamental importance to the design procedure is a knowledge of the radiation resistance and effective mass reactance of the diaphragm throughout the frequency range to be reproduced. The diaphragm, when in operation, breaks up into various radial and circular nodes or combinations of the two, depending upon the size, shape, material, constructional features, frequency, and amplitude. If the diaphragm moved strictly as a piston, its motional mass would be equal to its static mass; but since the surface movement is complex,

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varying with frequency, the effective mass of the diaphragm likewise varies with frequency. Because of this variation in area of effective radiating surface and the effect of frequency upon the radiating efficiency of a given area, the radiation resistance likewise undergoes a great variation throughout the frequency range to be reproduced. The complex variation of these two quantities is impossible of realization except by a direct measurement upon the diaphragm under consideration. There are two effects of a beneficial nature in the high-frequency range which must be given consideration because it is through them that a single cone-shaped structure responds to high-frequency signals as well as to those of low frequency. The unit mass, elasticity, and radial length of the cone are such as to cause it to be in transverse resonance[1] at about 2500 cycles. The velocity of wave transmission radially along the diaphragm is such that at this frequency, a wave is propagated from the apex to the outer rim and reflected back to the apex in the time of one-half cycle. This results in a great reduction in effective mass reactance in this frequency range, and hence serves to increase the high-frequency response,—in some cases to such an extent as to be undesirable. The amount of this increase, however, may be controlled by a choice or treatment of the material of the diaphragm and by its assembled angle. In general, the intensity of this high-frequency peak is reduced through the use of so-called "soft" materials in which the internal loss is great, or by circular corrugations, or by an increase in the cone interior angle to effect a reduction in the stiffness.

The other effect of benefit to a sustaining of the high-frequency response is the directional characteristic. At low frequencies, the locus of pressure forms substantially a semicircle with respect to the apex. At about 1000 cycles, however, the radiation begins to assume a beam shape which becomes quite evident at 2000 cycles. To listeners in line with the speaker, this has the effect of maintaining the high-frequency response even though the sound power is decreasing. These latter two effects are difficult to carry into the speaker design in a quantitative way. They are usually of a beneficial nature or can be so controlled as to give no difficulty.

Fig. 1 illustrates both of these effects. The increased response in the 2500-cycle range as well as the change in pressure at 30 degrees and 60 degrees from the cone center line are shown. Note that the directional effect begins to occur at about 1000 cycles.

At the outset, we are confronted with the choosing of the size of the diaphragm. Lord Rayleigh and Crandall have shown that the di-

[1] By transverse resonance is here meant the natural resonance of the conical surface to sound waves propagated radially through the cone material.

ameter of a piston vibrating in an infinite baffle should approximate a quarter wavelength of sound at the lowest frequency to be reproduced. Practice has shown, however, that large paper or fabric cones when made sufficiently light so as to function as a speaker diaphragm, cannot be made to move as a piston when driven from the apex, even at the lowest frequencies, such as 30 cycles. Because of this circumstance, any increase in diameter beyond a certain value results in no substantial extension of the low-frequency cut-off. For a household device, in which a uniform response down to 80 or 90 cycles is quite satisfactory, little or no benefit is experienced in utilizing a diaphragm whose diameter is greater than eight inches unless it be the psychological effect of size upon the prospective purchaser. Fig. 2 illustrates this point. This group of sound pressure curves was obtained from a speaker

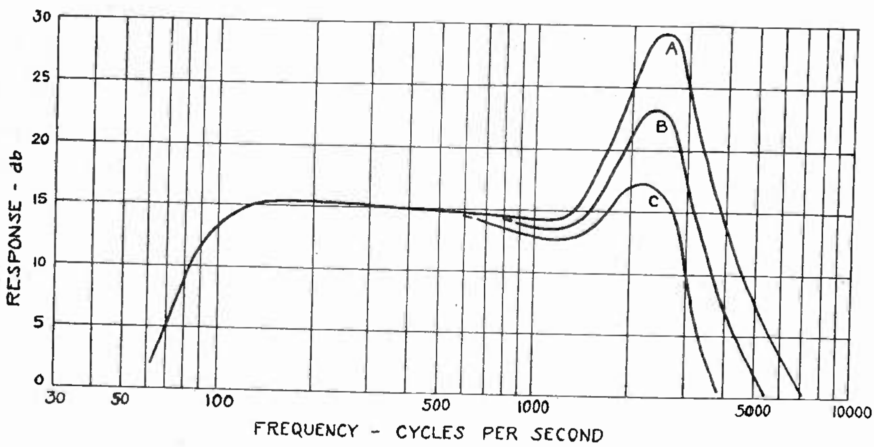
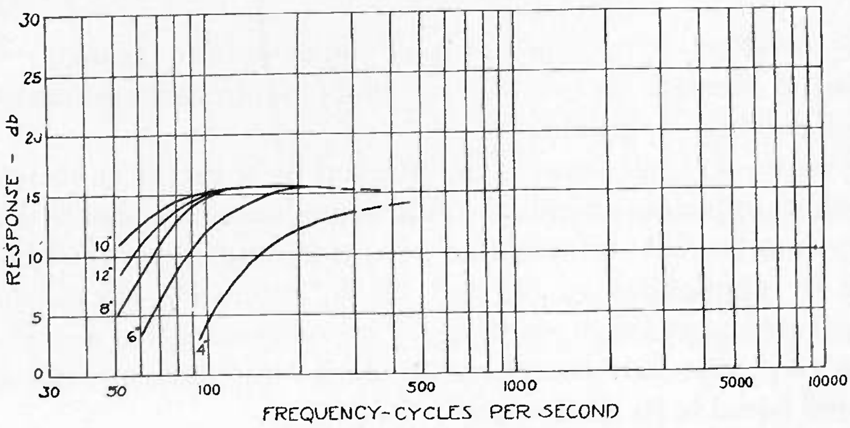


Fig. 1—Typical average response of an electrodynamic cone speaker with 48- by 48-inch flat baffle. Curve A is the response directly in line with speaker; curve B is 30° from center line; curve C is 60° from center line. Microphone 12 inches from cone.

having a twelve-inch diameter conical diaphragm. The restoring stiffness of the centering device was small so that the low-frequency cut-off may be safely considered as due to a decrease in the radiating efficiency of the diaphragm. The pressure curves at decreased cone diameters were obtained successively by recutting the original cone down to the next smaller size. No substantial variation in response between the various sizes occurs above 500 cycles and it can be seen that little gain in useful response can be realized through an increase in cone diameter beyond eight inches. It is evident that from a cost standpoint and adaptability the cone diameter should be held down to a minimum commensurate with adequate performance. It will be observed that the low-frequency cut-off of the twelve-inch diameter cone is higher than that of the ten-inch diameter cone. This is caused by the mass reactance of the twelve-inch size increasing at a rate greater than did the radiation resistance.

Determination of Diaphragm Mechanical Impedance

The radiation resistance and effective diaphragm mass reactance e determined through measurements of the voice coil impedance. he vector difference of the blocked and free impedance represents the ange in impedance due to motion and has hence been called the mo- nal impedance, being the vector sum of the motional resistance and e motional reactance. The mechanical resistance includes the acous- load and the hysteresis loss within the cone and its suspension. The echanical reactance is influenced at low frequencies by the compli- nce of the centering device and outer support, both external to the



ig. 2—Low-frequency response of an electrodynamic cone speaker with 48- by 48-inch baffle showing effect of diaphragms of from 4- to 12-inch diameters upon the low-frequency cut-off.

ne, and tending to introduce a resonance peak usually from 70 to 20 cycles. Consider a speaker voice coil having the following values:

E_c = counter e.m.f., r-m-s volts
B = flux density, gausses
l = length of conductor, cm
v = effective velocity, cm/sec
I = current, r-m-s amperes
Z = free electrical impedance, ohms
Z_0 = blocked electrical impedance, ohms
Z_Δ = motional impedance, ohms
Z_m = mechanical impedance, dyne sec/cm.

$$E_c = Blv \cdot 10^{-8} \text{ volts}$$

$$v = \frac{BlI}{10Z_m} \text{ cm/sec}$$

$$Z_\Delta = \frac{E_c}{I} = \frac{B^2l^2}{Z_m} 10^{-9} \text{ ohms}$$

$$= R_\Delta + jX_\Delta$$

$$Z_m = \frac{B^2l^2}{Z_\Delta{}^2}(R_\Delta - jX_\Delta)10^{-9} \text{ mechanical ohms}$$

$$= (r - jx)$$

$$r = \frac{B^2l^2}{Z_\Delta{}^2} R_\Delta \cdot 10^{-9} \text{ dyne sec/cm, mechanical resistance} \quad (1)$$

$$x = \frac{B^2l^2}{Z_\Delta{}^2} X_\Delta \cdot 10^{-9} \text{ dyne sec/cm, mechanical reactance} \quad (2)$$

$$= \omega(m_1 + m_2). \quad (3)$$

m_1 is the effective mass of the voice coil conductor, grams.

m_2 is the effective mass of the complete diaphragm exclusive of the voice coil conductor, grams.

The motional impedance is determined by a vector subtraction of the blocked impedance from the free impedance. By separating this quantity into its real and imaginary parts R_Δ and X_Δ, it then becomes possible by application of (1) and (2) to determine the mechanical resistance and mechanical reactance. The voice coil being a rigid body, its mass (m_1), does not change with change in frequency and can be considered equal to its static mass.

Efficiency

After the constants r and m_2 have been determined for a particular diaphragm, consideration can be given to the proper choice of voice coil mass and diameter, and the gap density which will result in a desired mechanical efficiency at minimum cost. By mechanical efficiency is here meant the ratio of mechanical power transmitted to the diaphragm to the electrical power supplied to the diaphragm and voice coil. The acoustic efficiency is the ratio of the acoustic power radiated to the mechanical power supplied to the diaphragm. In order to have this acoustic efficiency as high as possible, it is evident that the radiation resistance should be high and the effective mass of the moving parts kept at a minimum. The actual acoustic efficiency is not easily determined because r, in addition to the acoustic load, contains certain losses not readily separated. The mechanical efficiency, however, can be found.

The mechanical efficiency,

$$\eta = \frac{R_\Delta}{R_\Delta + R_0} \quad (4)$$

$$= \frac{1}{1 + \dfrac{R_0 Z_m^2 \cdot 10^9}{B^2 l^2 r}} \tag{5}$$

$$R_0 = \frac{\rho \delta l^2}{m_1}$$

where,

ρ = resistivity of conductor, ohms/cm^3

δ = density of conductor, grams/cm.^{3}

Substituting for R_0 in (5),

$$\eta = \frac{1}{1 + \dfrac{\rho\delta}{B^2 m_1 r}(r^2 + \omega^2(m_1 + m_2)^2)10^9} \cdot \tag{6}$$

This same result can be arrived at by considering the rate at which work is done by the diaphragm.

$F = \dfrac{BlI}{10}$ grams effective vibramotive force exerted by the coil

$v = \dfrac{F}{Z_m}$ effective velocity of voice coil, cm/sec

$P_\Delta = v^2 r \cdot 10^{-7}$ watts output

$$= \frac{B^2 l^2 P r}{(R_0 + R_\Delta) Z_m^2} 10^{-9}$$

$$\eta = \frac{P_\Delta}{P}$$

$$= \frac{1}{1 + \dfrac{\rho\delta}{B^2 m_1 r}(r^2 + \omega^2(m_1 + m_2)^2)10^9} \tag{6}$$

which is identical with the result obtained above.

It is evident that in the interest of a high efficiency, the mass reactance (ωm) should be kept as low as practicable and the radiation resistance be maintained as high as possible. The diaphragm area of cone speakers must be relatively large in order to have sufficient low-frequency radiation resistance, and hence, the diaphragm mass is relatively large. The efficiency for this reason never exceeds approximately five per cent. Horn type speakers, on the other hand utilize diaphragms of very small proportions, and hence, low mass, the radiation resistance being enhanced greatly by the loading effect of the horn. Horn speakers may be made to have an efficiency of forty per cent, or more. To increase the efficiency of cone speakers, some method must be found to increase the ratio of the radiation resistance to mass reactance without the use of horns or other cumbersome appurtenances or else through the use of a diaphragm material of lower density than is now being used, yet having the requisite rigidity.

Equation (6) also discloses that a conductor should be used in which the product of the resistivity and density is a minimum. This prefers the use of aluminum rather than copper for the voice coil conductor. This is true if the flux density is held constant, but it can be shown that for a given cost of field structure, the use of a copper voice coil will give greater efficiency than one of aluminum because the latter requires a greater air gap volume and as a result B^2 is lowered to a greater extent than is the coil resistance.

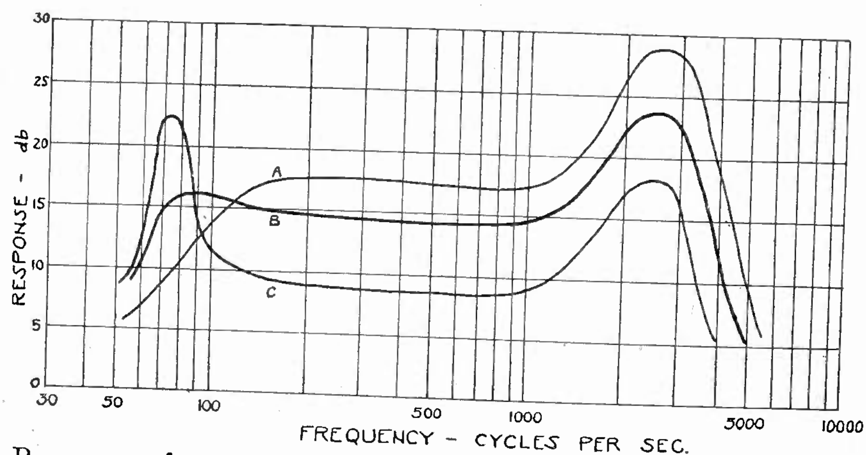


Fig. 3—Response of an electrodynamic cone speaker at various gap densities. Curve A is with gap density of 9600 gausses, curve B is with 7000 gausses, curve C is with 3500 gausses.

Equation (6) is of the form

$$y = \frac{1}{1+x}$$

in which (x) is large compared to unity so that substantially, the form is

$$y = \frac{1}{x}.$$

The efficiency is thus proportional to B^2, and the sound pressure is a linear function of the gap density. This effect is illustrated in Fig. 3, showing the sound pressure for a particular cone speaker for various gap densities. At the extreme low frequencies, this relation is disturbed by the natural period of the vibratory system and by the change in net driving force with change in flux density. In a loaded shunt motor, some particular field strength gives maximum speed. Likewise, in an electrodynamic speaker, a particular field strength will give maximum vibratory velocity or sound pressure at some particular frequency. Ordinarily, this maximum is reached in the low resonant range where the cone mass reactance is neutralized by the compliance of the support. This resonant effect can be used to good advantage to enhance the low-frequency response provided an optimum flux density is used. In the case under consideration in Fig. 3, a flux density of about 7000 gausses seems best. Higher densities than this increases the general signal strength but at a sacrifice in the response in the low-frequency range.

Optimum Voice Coil Mass

With the gap density held constant, any increase in the voice coil conductor mass will increase the driving force, but this mass will add to the combined diaphragm mass, increasing the mechanical load impedance. By differentiating η of (6) with respect to m_1, we find η to be a maximum when

$$\omega m_1 = \sqrt{r^2 + \omega^2 m_2^2} \tag{7}$$

or when the mechanical impedance of the driving coil is equal to the mechanical impedance of the load exclusive of the driving coil. This is an interesting parallel to the proposition that maximum power is drawn from a generator when the load impedance is equal to the impedance of the generator. Over the whole range of frequency, the mechanical impedance of the cone is essentially its mass reactance, the radiation resistance being small in comparison. The voice coil mass, then, can be matched directly to the mass of the cone. The effective mass of the cone, however, decreases greatly with increase in frequency. Fig. 4 illustrates the general change in mass of a typical diaphragm. The actual change is more or less erratic—the curve covers the general tendency. This decrease in mass with frequency is a fortunate circumstance making possible the production of a wide frequency spectrum. The voice coil mass is usually matched to the cone mass at the mid-range so as to give the proper balance to the response. A decrease in voice coil mass will increase the high-frequency output at a sacrifice of lows, and likewise, an increase in voice coil mass will increase the low-frequency response at a sacrifice of highs. Because of cost considerations of the magnet structure, the mass of the voice coil is kept at a minimum commensurate with the desired balance in order that the gap size, and hence, the magnet size be maintained as small as possible.

With the mass of the voice coil selected, it is then possible to determine the optimum gap density and input volt-amperes required to give a desired performance. The exact determination of these latter two

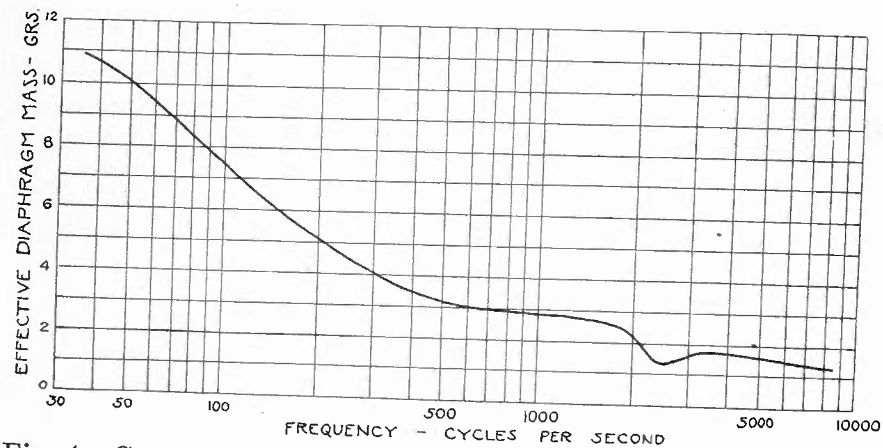


Fig. 4—General change in effective mass of the radiating surface of an electrodynamic speaker.

quantities is of more use in connection with the design of a constant frequency vibratory drive where the load and deflection are known.

The Field Magnet

For a particular mass of voice coil conductor, gap density, and field watts, there is a particular diameter of voice coil which will give a minimum cost of magnet structure. The following considerations will lead to this conclusion. Practice has shown that the use of a core of constant cross section results in a more economical magnet than is obtainable with any other form of core. The leakage flux in this case is as low as can be achieved in other forms and the mean turn of the field coil is kept to a minimum.

The depth of gap is established through the choice of gap diameter and flux density.

D = mean diameter of voice coil, cm
L = depth of gap, cm
B = gap density, gausses
B_c = maximum allowable core density, gausses.

The gap flux is equal to about one half the total core flux, so that

$$2\pi DLB = \frac{\pi}{4} D^2 B_c$$

or,

$$L = 0.125 \frac{DB_c}{B} \text{ cm.} \tag{8}$$

B_c varies from 10,000 to 12,000 depending upon the grade of iron used.

A gap density of about 6500 to 7000 gausses has proved to be a good choice both from an economical and a performance standpoint. This gives a coil depth of about one fourth the diameter. This ratio of diameter to depth has proved quite satisfactory, mechanically speaking, giving a structure quite rugged, insuring a freedom from becoming elliptical which may cause trouble by rubbing the sides of the pole pieces. The width of the gap is fixed after finding the cross-sectional dimensions of the voice coil.

m_1 = mass of voice coil conductor, grams
A = total cross-sectional area of voice coil, cm
S = space factor = 0.6
δ = density of conductor.

Then, if a copper conductor is used,

$$A = 0.06 \frac{m_1}{D} \text{ cm}^2 \tag{9}$$

$$t = \frac{A}{L} \text{ thickness of voice coil, cm} \tag{10}$$

$$g = t + 0.07 \text{ cm, gap width} \tag{11}$$

allowing 10 mils inside clearance
12 mils outside clearance
5 mils coil form thickness.

The design of the field coil and the magnet circuit can now be considered.

E_F = field voltage
P_F = field watts
NI = field ampere turns
R_F = field resistance, ohms
r_c = ohms per cm of conductor
D_m = mean diameter of coil, cm
M = mean turn, cm
a_c = area of conductor, cm^2
A_c = total conductor area, cm^2.

Assuming the reluctance of the iron path to be about 26 per cent of the gap reluctance,

$$Bg = NI \text{ ampere turns} \tag{12}$$

$$NI = \frac{E_F}{Mr_c}$$

$$r_c = \frac{E_F}{MNI} \text{ ohms per cm.} \tag{13}$$

Equation (13) is useful in determining the size of conductor.

$$a_c = \frac{MNI\rho}{E_F} \text{ cm}^2 \tag{14}$$

$$A_c = Na_c = \frac{2M(NI)^2}{P_F} 10^{-6} \text{ cm}^2. \tag{15}$$

The weight of the field coil conductor is

$$W_c = \delta MA_c = \frac{(17.4)M^2(NI)^2}{P_F} 10^{-6} \text{ grams.} \tag{16}$$

The weight of the iron in the magnetic circuit is empirically estimated as follows:

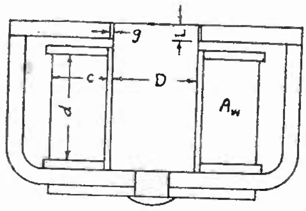


Fig. 5—Cross section of electrodynamic speaker field magnet.

The cross-sectional area of the field coil winding is

$$A_w = cd = \frac{(3.3)M(NI)^2}{P_F} 10^{-6} \text{ cm}^2 \quad 17)$$

where,

c is the radial depth of the field winding
d is the axial length of the field winding.

For a good balance between mean turn, length of iron path, and leakage flux between the sides of the magnetic circuit,

$$d = 2c$$

or,

$$A_w = 2c^2. \tag{18}$$

A value of c must be chosen which will satisfy both (24) and (25), M being equal to $D+0.3+c$. A_w is found through trial calculation by varying c until (17) and (18) are equivalent. With the size of the field coil determined and the diameter of the core known, the weight of the iron constituting the magnetic circuit can be estimated.

Thus,

$$L = 8c \text{ cm, length of iron circuit}$$

$$A = \frac{\pi}{4} D^2 \text{ cm}^2\text{, area of iron section}$$

$$W = 50\ D^2 c \text{ grams, weight of iron.} \tag{19}$$

With a given voice coil mass and gap density, a change in the voice coil diameter will change the amount and the proportion of copper and iron constituting the field. Knowing the ratio of cost per gram of the

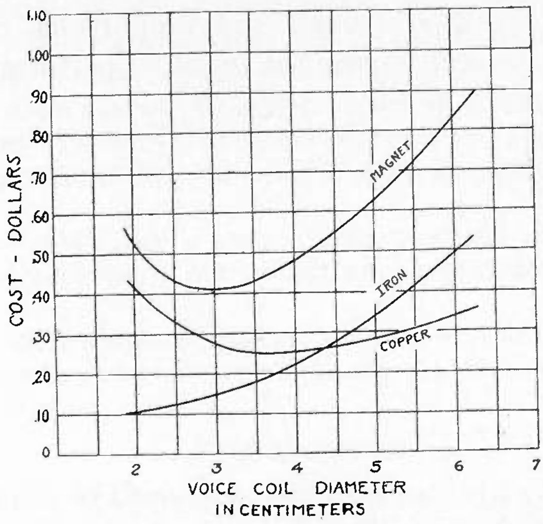


Fig. 6—Showing the effect of voice coil diameter upon the cost of the field magnet of an electrodynamic speaker.

wound copper and finished iron, the voice coil diameter giving the minimum cost of magnet can be ascertained. For example, a magnet for a 2.5-gram voice coil of which 1.5 grams is within the gap (the remainder overhanging the gap to take advantage of the leakage flux), and having a gap density of 7000 gauss at 5 watts, should have a core diameter of about 3 centimeters. This is shown graphically in Fig. 6. At diameters less than this optimum, the copper increases rapidly because of the rapid increase in gap width. At diameters greater than the optimum, the mean length of turn in the field is increasing at a faster rate than the gap is decreasing—the gap approaching as a limit, the fixed clearance of 0.07 centimeter. The weight of the iron in the magnetic circuit continually increases with increase in voice coil diameter.

This method of determining the optimum core diameter neglects many minor considerations and consequently only serves to establish the approximate design dimensions. The final design can only be reached through a continued and more exacting investigation and the building and testing of models.

The primary object was to show the general method to be followed in the design of the magnet. The proportions and gap clearances chosen will depend upon the application.

Conclusion

The nature of the operation of the speaker diaphragm when considered over the whole sound spectrum is so complicated that the major proportion of the design work must be done experimentally. A complete understanding of the device will, however, enable the worker to proceed intelligently. Certain aspects of the design such as the field magnet, and the voice coil mass and proportions can be analyzed exactly to give a speaker having the required performance and with a minimum factory cost.

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THE INNER-GRID DYNATRON AND THE DUODYNATRON*

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Summary—*The falling characteristics of a tetrode due to the secondary emission from the inner-grid electrode are fully examined. A new type of dynatron oscillation is found to occur in a parallel resonance circuit connected to the inner-grid return. Influence of the electrode voltage upon the operating characteristics of such an oscillator is experimentally investigated.*

An abnormal form of oscillation may be maintained when the static negative resistance of the circuit is exceedingly small in comparison with the impedance of the oscillator tank circuit. The possibility of existence of this novel form of oscillation is explained by introducing an idea of the average negative resistance.

When the electrodes are kept at suitable voltage, both the anode and the inner-grid circuits may be made to have negative electric resistances. Thus a new dynatron oscillator is formed which has two oscillation circuits oscillating at different frequencies. This may be used as a simple beat-frequency oscillator comprising only a single tetrode. The author proposes to call this complex dynatron oscillator a "duodynatron" for brevity.

Detailed experimental results are given of this oscillator, with special regard to the synchronizing phenomenon of the weaker inner-grid oscillation by the stronger anode oscillation.

In the mathematical treatment, starting from the differential equations for the primary and the secondary electron currents, oscillation frequencies of the duodynatron oscillator are derived.

I. Introduction

The prototype of dynatron was first introduced by A. W. Hull[1] in 1918, in which he devised the utilization of secondary emission from the anode. We shall call this an anode dynatron.

In 1930, Y. Ito[2] proposed to use as the dynode the grid in a triode or the outer grid in a tetrode. We shall correspondingly call this a grid dynatron or an outer-grid dynatron.

The object of the present paper is to introduce a new falling characteristic device depending upon the secondary emission from the inner-grid electrode of a tetrode. A novel type of dynatron oscillator is subsequently introduced, in which two oscillations of different frequencies are maintained by a single valve simultaneously in the anode circuit and the inner-grid circuit.

II. Three Kinds of Falling Characteristics in a Tetrode

The falling characteristics manifested by a tetrode may be classified into three kinds.

* Decimal classification: R133. Original manuscript received by the Institute, September 25, 1933.

[1] Figures refer to bibliography.

(a). First—anode dynatron. (By A. W. Hull)
(b). Second—outer-grid dynatron. (By Y. Ito)
(c). Third—inner-grid dynatron. (By the writer)

In the dynatron of the first kind, the primary electrons that are accelerated by the positively charged inner and outer grids impinge upon the surface of anode and the secondary electrons emitted from the anode are absorbed by the outer grid whose potential is kept higher than the anode.

If, V_a, V_g, V_r, represent the potential of the anode, the outer grid, and the inner grid, respectively,

$$V_a < V_g, \quad \text{and} \quad V_r < V_g.$$

The dynatron of the second kind depends upon the emission of secondary electrons from the outer grid due to the bombardment by primary electrons accelerated by the inner grid. The anode has to absorb the secondary electrons from the outer grid. Thus there must be the following relations:

$$V_g < V_a, \quad \text{and} \quad V_r < V_a.$$

The dynatron of the third kind depends upon the secondary electron emission from the inner grid. The outer grid collects the secondary electrons, and the anode potential controls the slope of the falling characteristics. In this kind, the electrode voltage relations should be

$$V_r < V_g, \quad \text{and} \quad V_a < V_g.$$

The experiments carried out by the writer with regard to the first and second kinds gave the results quite consistent with those reported by the previous investigators.

In the following, detailed accounts of the experimental and theoretical investigation of the dynatron of the third kind, i.e., the inner-grid dynatron, will be given.

III. Inner-Grid Dynatron

(1). Inner-Grid Dynatron Characteristics

Dynatron of the Third Kind

Throughout the experimental investigations, a concentric tetrode with double grids was used. This is manufactured by Siemens-Halske and named RE-87, or more popularly the "OR" valve. Diameters of the inner grid, the outer grid, and the anode are 6.5, 15, and 21 millimeters, respectively.

The principal constructional difference from the ordinary tetrode lies in the fact that the grid is not spiral but is formed of longitudinal thin narrow ribbons arranged parallel to the axis of the filament. The surface of the grid metal is inclined so that the electrons emitted from the filament will strike it at an angle of about 45 degrees.

According to Okabe's experimental results,[4] the secondary emission is more conspicuous when the primary electrons strike the metal surface at a slanting angle. This is a fact to be anticipated from the con-

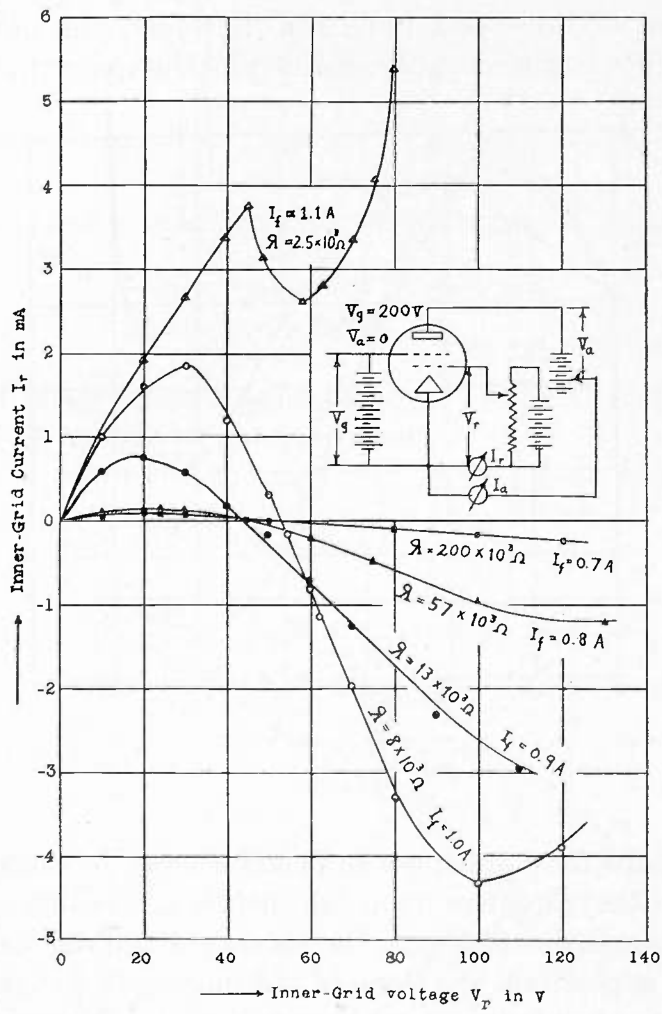

Fig. 1

sideration of the atomic structure of metals. The "OR" valve is thus a suitable one for the utilization of secondary electron emission from the grid. Fig. 1 shows the inner-grid dynatron characteristics at various filament currents. In cases of the anode dynatron and the outer-grid dynatron, the augmentation of filament current above normal gives only a little effect upon the characteristics, but in the case of the inner-grid dynatron, the amount of secondary electron emission becomes very small when the filament current is increased above normal.

In order to elucidate this phenomenon, the presence of space charge must be taken into consideration. The increase of filament current will naturally cause an increase of electron emission, and consequently the increase of space charge in the neighborhood of the filament. In the case of anode or outer-grid dynatron, there exists as least one electrode acting as a space-charge grid and absorbing the unnecessary space charge.

In the case of inner-grid dynatron, however, the motion of the primary electrons is considerably retarded by the space charge, and the

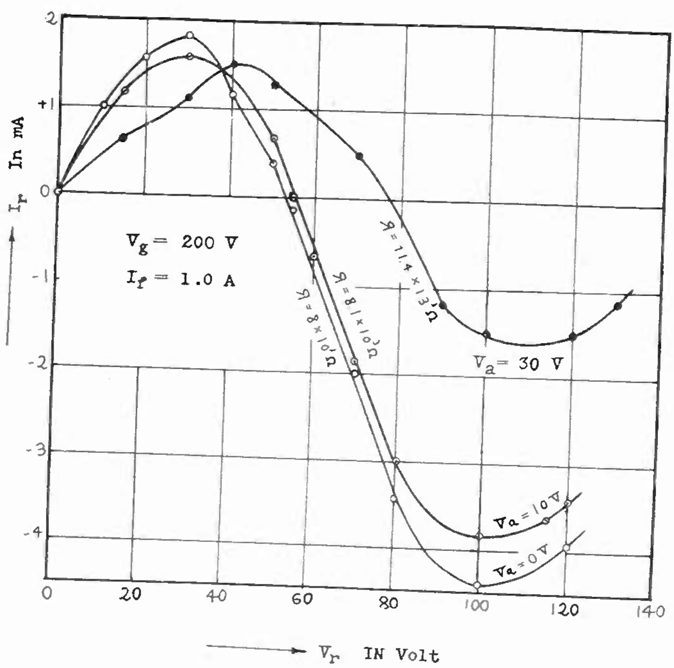

Fig. 2

primary electrons may not have sufficient energy to cause secondary emission when they impinge upon the surface of the inner grid.

In the measurement of Fig. 1, the anode voltage was kept constant at zero. If V_a is positive, the slope of the inner-grid dynatron characteristic tends to be unfavorable as shown in Fig. 2. The negative value of V_a has, as may be anticipated, no influence upon the characteristics.

(2). Inner-Grid Dynatron Oscillator

It is obvious that the falling characteristic, or in other words the negative resistance characteristic, of the inner-grid circuit may cause a dynatron type oscillation in a parallel resonant circuit connected to the inner-grid return. Let us now denote the self-inductance of the coil by L_r, the electrostatic capacity of the condenser by C_r, the effective resistance in series with the inductance by R_r, and the negative electric

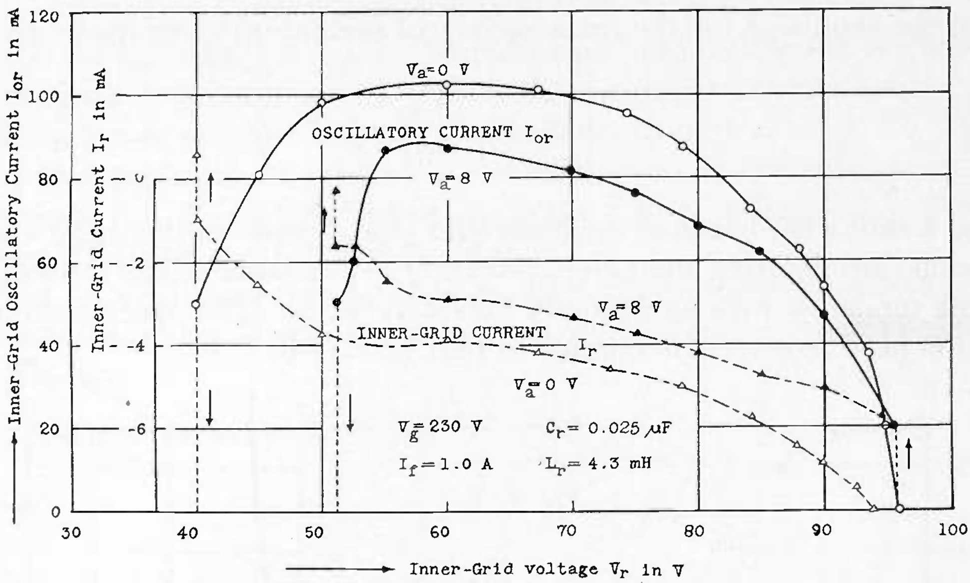

Fig. 3

resistance of the inner-grid dynatron by $|-\text{Я}_r|$. Then the oscillation frequency may readily be obtained from,[1,3]

$$f_r = \frac{1}{2\pi\sqrt{L_r C_r}}\sqrt{1-\frac{R_r}{\text{Я}}} \tag{1}$$

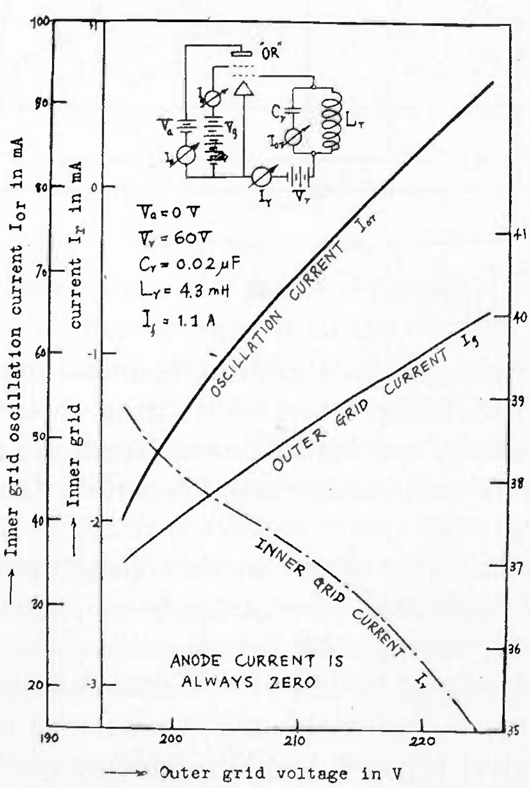

Fig. 4

and the condition for the maintenance of oscillation is expressed by

$$|\mathfrak{R}_r| \leqq \frac{L_r}{C_r R_r} \cdot \tag{2}$$

Fig. 3 shows the effect of the inner-grid (the dynode in this case) voltage on the operating characteristic. The $I_{or}-V_r$ characteristic shows the same tendency with an ordinary anode dynatron. The operating condition at $V_a=8$ volts is inferior to that at $V_a=0$.

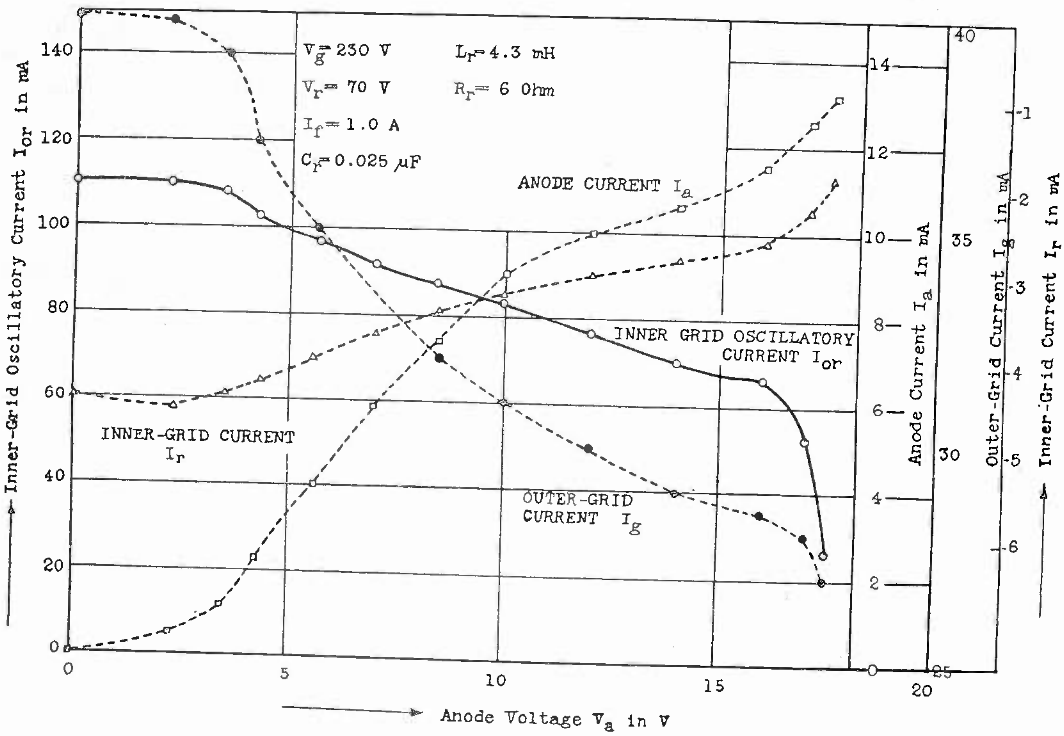

Fig. 5

The outer-grid voltage has a great influence upon the operating characteristic as shown in Fig. 4, and even a drop of 20 volts will reduce the oscillatory current by one half. The characteristic curves of Fig. 4 indicate that the outer-grid voltage is an important factor for modulation.

Fig. 5 shows the influence of the anode voltage upon the operating characteristics. The oscillation current tends to decrease rapidly at a certain anode voltage, say 17 volts in this case.

The inner-grid dynatron oscillator has hitherto found no practical applications; the object of this chapter, however, is nothing but the preparation for a novel type of double frequency oscillator, the duodynatron, described in the subsequent chapter.

IV. The Concept of Average Negative Resistance

In the ordinary retroactive oscillator, self-oscillation will start at a critical anode voltage, say V_a'. While the oscillation is maintained, the diminution of anode voltage down to a certain critical value V_a'' will stop the oscillation. In general, V_a'' does not coincide with but is smaller than V_a'. From the viewpoint of the "Mittere-Steilheit," Barkhausen[5] has explained this phenomenon.

In an analogous manner, the oscillation of anode-dynatron oscillator starts abruptly at a certain positive anode voltage as shown by *s* in Fig. 6.

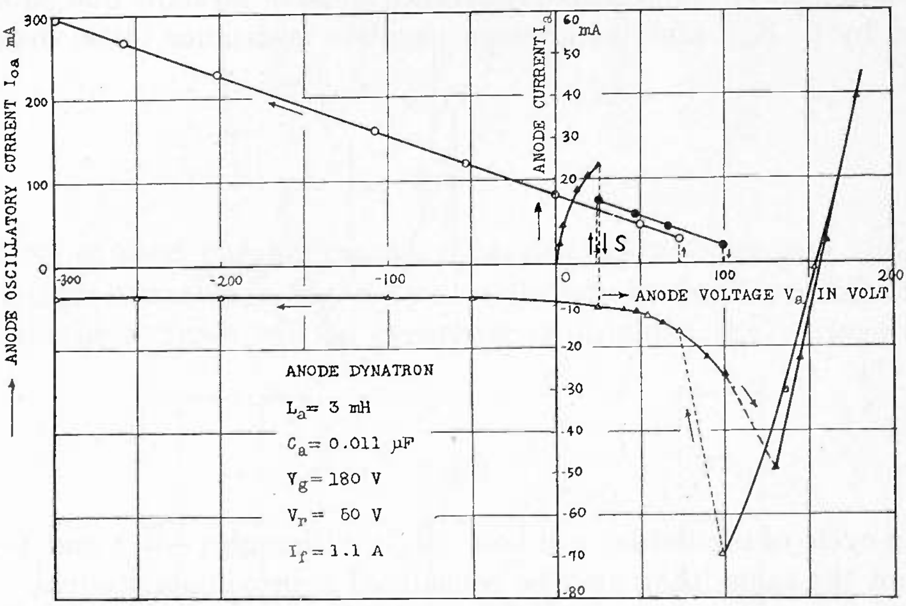

Fig. 6

In certain cases, when the anode voltage is decreased, the self-oscillation is still maintained down to the voltage considerably lower than that required for starting. In Fig. 6, for example, the oscillation is maintained even at a negative anode voltage, and the oscillation current is then rather increasing. The mode of oscillation at this state should of course be different from the simple normal dynatron oscillation. The negative electric resistance is no longer constant in this case, therefore, it is necessary now to introduce the concept of "average negative resistance" analogous to Barkhausen's "Mittere-Steilheit."

In Fig. 7, a self-oscillation will be maintained at the falling region $\overline{bd}$. It is convenient to introduce a new symbol as follows.

$$\alpha = \frac{(L/CR)}{Я} = \frac{L}{CRЯ}$$

where,

L = self-inductance of coil.
C = electrostatic capacity of condenser.
R = ohmic resistance in series with L.
$(-Я)$ = negative resistance of the circuit.

When $\alpha > 1$, the range of oscillation will not be confined to $\overline{bd}$, but will extend as far as α becomes equal to 1. For example, b will tend to become b', and d to d'. When the condition $\alpha > 1$ is still maintained at $V_a = (-a)$, the oscillation cycle may cover a wider range and extend nearly over the range $\overline{aobcd}$.

As in Fig. 7, the negative resistance slope of straight line $\overline{ad}$ is denoted by $(-Я_a)$, and the average negative resistance taken over the

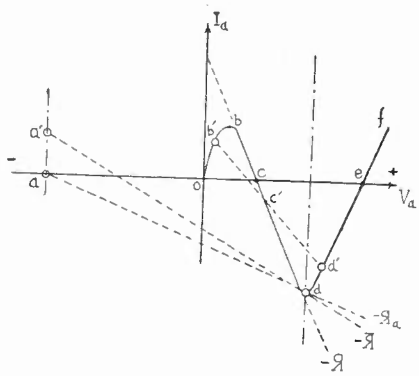

Fig. 7

whole cycle of oscillation will be $(-\bar{Я}_a)$. Although $(-\bar{Я}_a)$ and $(-Я_a)$ are not the same, they may be considered approximately equal.

The necessary condition $\alpha > 1$ for the maintenance of oscillation is a theoretical one with the assumption that the wave form of oscillatory current is sinusoidal. The oscillation now under consideration is not strictly sinusoidal, and hence the following correction should be made:

$$\bar{Я}(I_0) \leqq \frac{L}{CR} \cdot \eta$$

$$\eta = \psi(a_m, b_m)$$

$$I_0 = \sum_{m=1}^{\infty} a_m \cos m\,\omega t + \sum_{m=1}^{\infty} b_m \sin m\,\omega t$$

where,

I_0 = oscillation current.
ω = angular frequency $2\pi f$.

From these relations, it is to be anticipated that the oscillation in Fig. 6 will stop suddenly at a certain negative value of V_a. In the experiment, however, the limit could not be obtained unless the potential difference between the anode and the outer grid was kept higher than 500 volts.

V. Duodynatron Oscillator

By the duodynatron is understood a system in which the dynatron of the first kind and the dynatron of the third kind are incorporated with a single tetrode.

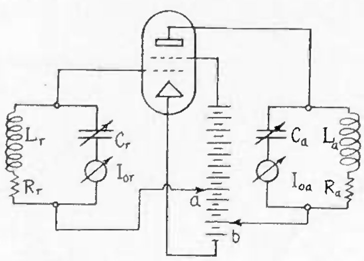

Fig. 8

If we connect parallel resonant circuits in the inner grid and the anode sides as shown in Fig. 8, two oscillations having their respective resonant frequencies will be maintained, provided the various elec-

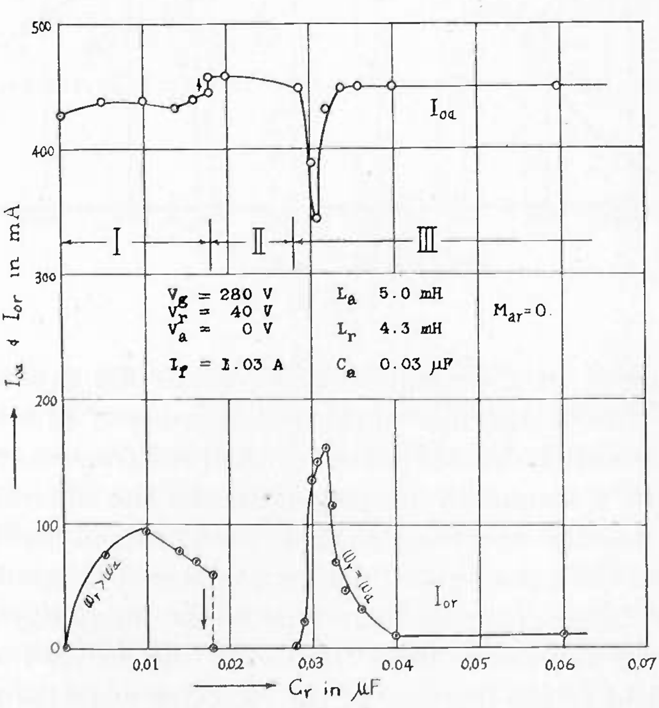

Fig. 9

trodes are kept at suitable voltage. It is conceivable, from the concept of the average negative resistance, that the anode-dynatron oscillation may persist at zero-anode voltage after the oscillation has once set in.

On the other hand, due to the falling characteristic as shown in Fig. 2, another oscillation may take place at the inner-grid side.

The inner-grid oscillatory current I_{or} and the anode oscillatory current I_{oa} are plotted as the functions of C_r. (Fig. 9.)

In range "I" ($C_r = 0 \sim 0.018$ microfarad), the oscillation frequency of the inner-grid circuit differs from that of the anode side, and thus the simultaneous generation of two different oscillations is experimentally established.

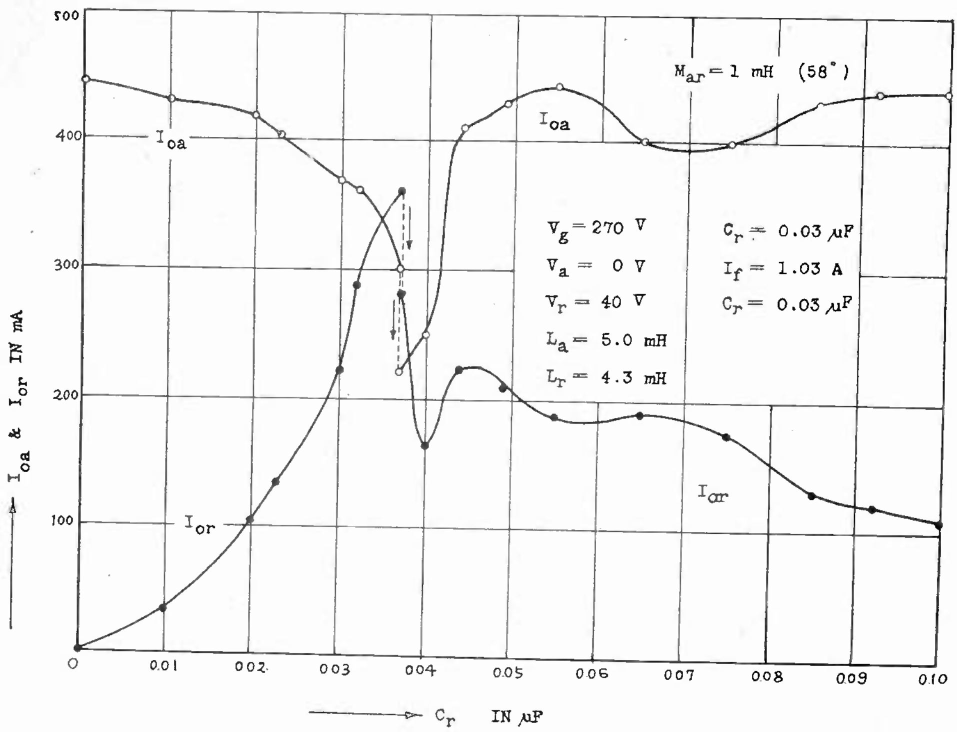

Fig. 10

The frequency of the anode oscillation in this case was 13 kilocycles, while that of the inner-grid oscillation was 15.9 kilocycles at $C_r = 0.016$ microfarad. An oscillation of the beat frequency, namely an oscillation with a frequency corresponding to the difference of these frequencies,* was also observed, so that three kinds of oscillating sound were audible at the same time. Further increase of C_r results in the extinction of the inner-grid oscillation due to the proximity of the natural frequency of the anode tank circuit. In the writer's opinion, this must be ascribed to the increase of the effective negative resistance of the inner-grid side.

When the natural frequency of the inner-grid tank circuit is equal to that of the anode side, a resonance occurs. In this case the oscillatory

* In this case, the beat frequency was about 3000 cycles.

current in the grid circuit may be considered an effect of induction from the anode side. Thus in range "III," the oscillation frequency is fixed by the anode side frequency.

A self-oscillation of the inner-grid side may be produced once again, when the inner-grid natural frequency is made considerably smaller than that of the anode circuit. But, the value of L_r/C_rR_r in this case is comparatively small, and the condition for oscillation does not hold unless the effective negative resistance is exceedingly small. When a mutual inductance M_{ar} exists between L_r and L_a, the oscillation mech-

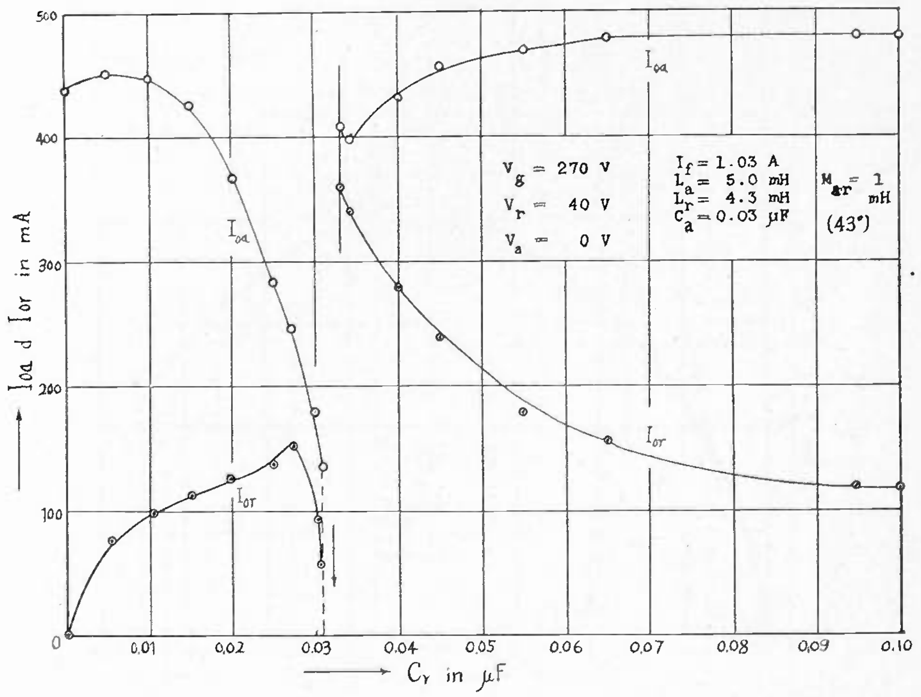

Fig. 11

anism becomes much involved. For example, as in Fig. 10, the state of oscillation changes abruptly at $C_r = 0.0365$ microfarad. This is considered to be the spontaneous extinction of oscillation in the inner-grid side. The oscillatory current of the inner grid I_{or} sometimes increases as high as 300 milliamperes. The reason for this is not yet clear.

When the sign of the mutual inductance M_{ar} is reversed, oscillations on both sides die out at $C_r = 0.031$ microfarad as shown in Fig. 11. Further increase of C_r reveals no inclination to oscillation. But, when the new oscillation is once started at the anode side by a proper setting of the anode voltage (in this case at $C_r = 0.033$ microfarad), I_{or} obtains an intensity of 300 millamperes or more, as indicated by the curve at the right-hand side of Fig. 11. In order to study the retroaction due to M_{ar}, anode oscillation was temporarily stopped, and a series of observations was undertaken in the form of a simple inner-grid dynatron oscillator, setting at $M_{ar}=0$, or $M_{ar}=1$ millihenry. Referring to Fig. 12, it may be stated that the region of the inner-grid oscillation is the widest when $M_{ar}=0$.

Hence, the existence of M_{ar} causes an apparent augmentation of the effective negative resistance. Consequently, the feed-back action in this case is not so simple as in the case of an ordinary retroactive triode oscillator. In practice, the case $M_{ar}=0$ is particularly important, since there are no special merits of the back coupling. It is highly interesting,

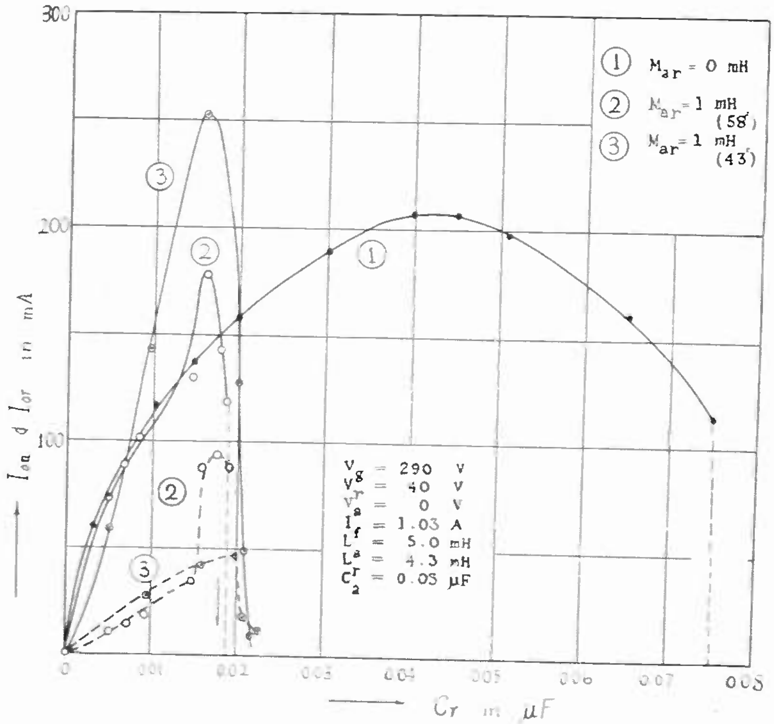

Fig. 12

however, that the beat frequency can be very finely controlled by varying the value of M_{ar}, so the use of back coupling is advantageous in the duodynatron beat-frequency oscillator.

An interesting result was obtained in the duodynatron by a change of inner-grid voltage V_r. (Fig. 13.) When $C_r=0.032$ microfarad, i.e., $\omega_r=\omega_a$, the inner-grid and the anode circuits are at resonance. Under this condition, the inner-grid oscillatory current I_{or} is approximately proportional to the anode oscillatory current I_{oa}, and the oscillations in both circuits vanish simultaneously at $V_r=14$ volts.

It is evident, therefore, that when $C_r=0.032$ microfarad, the oscillation in the inner-grid circuit I_{or} is not the oscillation of the inner grid itself, but, as has previously been mentioned, it is nothing but the induced oscillation from the anode side.

On the other hand, when $C_r = 0.01$ microfarad, i.e., $\omega_r > \omega_a$, the inner-grid oscillatory current I_{or} is inversely proportional to V_r. From the previous description (Cf. Fig. 6), it is understood that the dynode oscillatory current is almost inversely proportional to the dynode potential. In addition, the extinction voltage of the oscillation of the inner-grid side is greater than that of the anode. Therefore, it is reasonable to think that the inner-grid oscillation in this case is of its own.

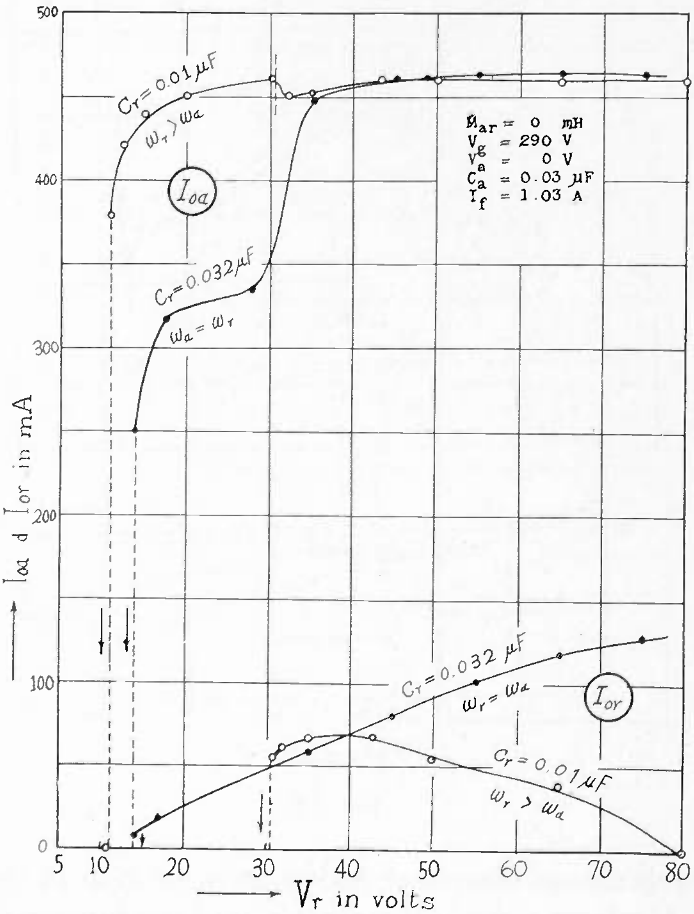

Fig. 13

Fig. 14 shows the effect of anode voltage on the oscillatory system. The real oscillation range is much wider than was anticipated from the static characteristics. The increase of anode voltage is followed by the augmentation of negative resistance in both anode and inner-grid circuits. Hence, the amplitude of oscillations diminishes with the increase of anode voltage.

As mentioned above, when a mutual inductance M_{ar} exists between two oscillation coils, the mechanism of oscillation is exceedingly complicated. When the system is oscillating at a single frequency, the problem is somewhat simpler.

When there are two oscillations of different frequencies, it is necessary to investigate into the synchronizing phenomena and the "Ziehen" phenomena.

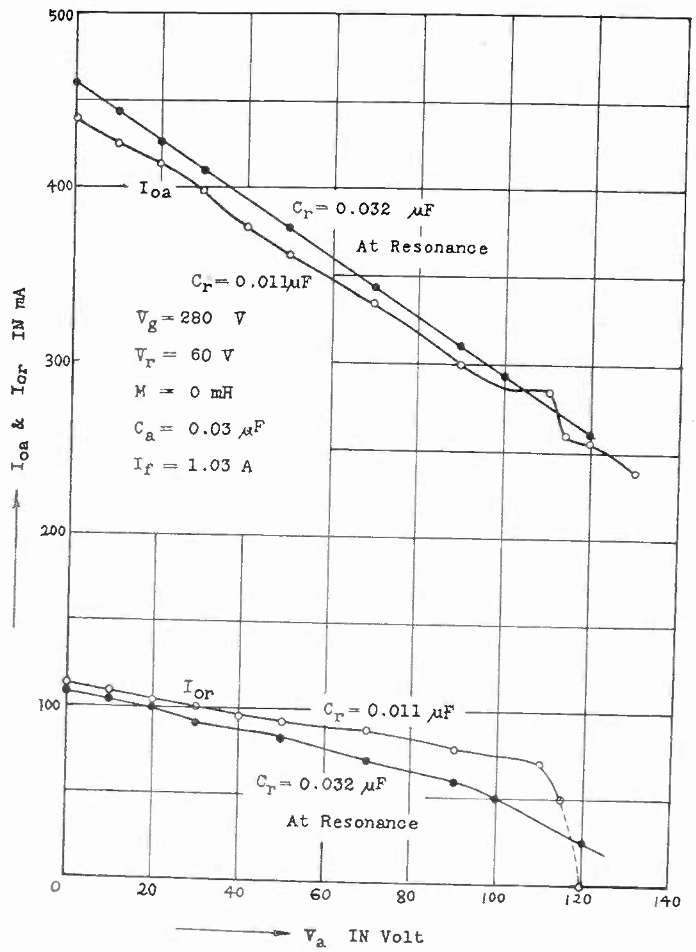

Fig. 14

Fig. 15 represents the effect of variation of M_{ar} on the oscillation amplitude. At about $\omega_r = \omega_a$, the variation of the inner-grid oscillatory current I_{or} takes the shape of an inverted resonance curve, while the anode oscillatory current I_{oa} varies as a normal resonance curve. This fact is also a plain proof that the oscillation in the grid side is merely the induction from the anode circuit.

If the frequency relation is $\omega_r > \omega_a$ ($C_r = 0.011$ microfarad), however, the oscillation mechanism is not so easy to explain. As shown in Fig. 15, jumping phenomenon occurs. Different oscillations are obtainable only before the jumping. After the jumping, the inner-grid oscillation seems to be pulled into another frequency (probably equal to the frequency of anode oscillation) due to the coupling between L_r and L_a.

It is interesting to note that there is a possibility of maintaining four oscillations at different frequencies, by properly combining the writer's duodynatron with another method of double frequency oscillation of McLachlan's.[6]

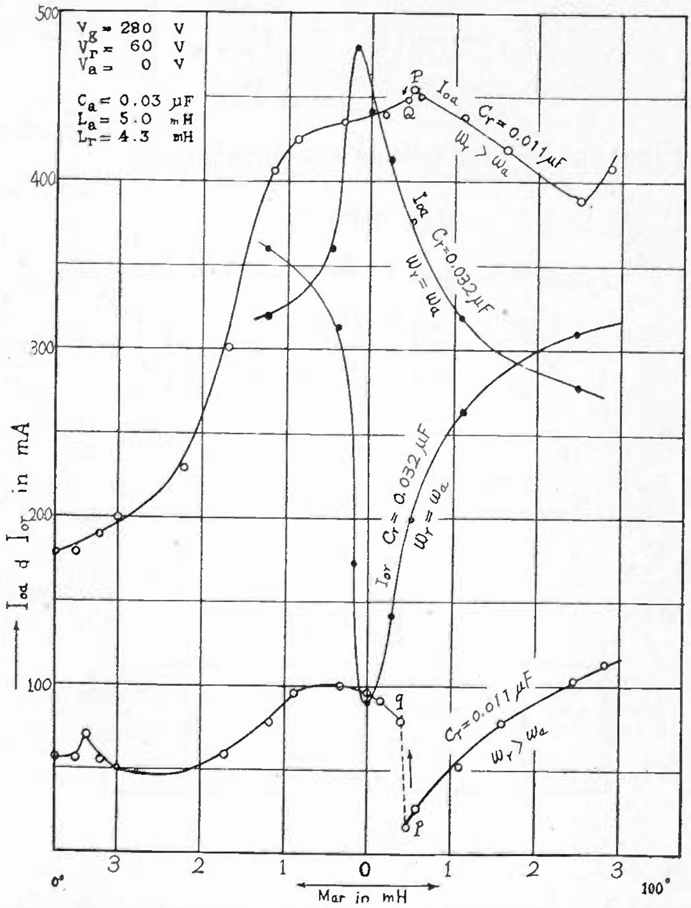

Fig. 15

VI. Theory of the Duodynatron Oscillator

The current relations between four electrode circuits of a duodynatron are shown in Fig. 16, in which, V_a, V_g, V_r, denote anode, outer-grid, and inner-grid voltages, respectively. I_a, I_g, I_r, denote anode,

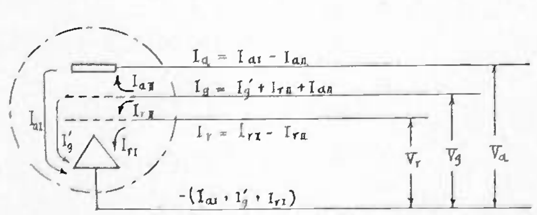

Fig. 16

outer-grid, and inner-grid currents, respectively. Suffix I refers to primary electron current. Suffix II refers to secondary electron current.

Referring to Fig. 16, we put

$$\left.\begin{aligned} I_{a\mathrm{I}} &= \Psi_{\mathrm{I}}(V_r,\ V_g,\ V_a) \\ I_{a\mathrm{II}} &= \Psi_{\mathrm{II}}(V_r,\ V_g,\ V_a) \\ I_{r\mathrm{I}} &= \Theta_{\mathrm{I}}(V_r,\ V_g,\ V_a) \\ I_{r\mathrm{II}} &= \Theta_{\mathrm{II}}(V_r,\ V_g,\ V_a) \end{aligned}\right\}. \tag{7}$$

When the instantaneous values are considered

$$\left.\begin{aligned} i_{a\mathrm{I}} &= \psi_{\mathrm{I}}(v_r,\ v_a) \\ i_{a\mathrm{II}} &= \psi_{\mathrm{II}}(v_r,\ v_a) \\ i_{r\mathrm{I}} &= \theta_{\mathrm{I}}(v_r,\ v_a) \\ i_{r\mathrm{II}} &= \theta_{\mathrm{II}}(v_r,\ v_a) \end{aligned}\right\}. \tag{8}$$

In this case,

$$\left.\begin{aligned} &\frac{\partial V_g}{\partial t} = 0 \\ \text{hence,}\quad & v_g = 0 \end{aligned}\right\}. \tag{9}$$

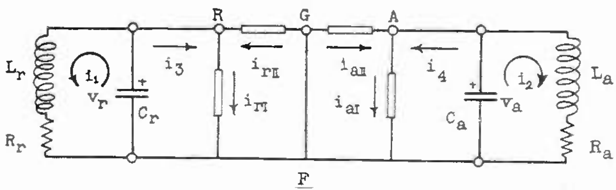

Fig. 17

According to the writer's experimental results, (8) can be approximately written in the form

$$\begin{aligned} i_{a\mathrm{I}} &= \frac{1}{r_{\mathrm{I}}}\ (-\ v_r + \gamma_{\mathrm{I}} v_a) \\ i_{a\mathrm{II}} &= \frac{1}{r_{\mathrm{II}}}\ (-\ v_r + \gamma_{\mathrm{II}} v_a) \\ i_{r\mathrm{I}} &= \frac{1}{\rho_{\mathrm{I}}}\ (v_r - \eta_{\mathrm{I}} v_a) \\ i_{r\mathrm{II}} &= \frac{1}{\rho_{\mathrm{II}}}\ (v_r - \eta_{\mathrm{II}} v_a). \end{aligned} \tag{10}$$

Hence, from the equivalent circuit of the duodynatron as shown in Fig. 17, we have simultaneous differential equations of the second order

$$\left.\begin{aligned} a_{11}\frac{d^2v_r}{dt^2} + b_{11}\frac{dv_r}{dt} + c_{11}v_r + b_{12}\frac{dv_a}{dt} + c_{12}v_a &= 0 \\ a_{22}\frac{d^2v_a}{dt^2} + b_{22}\frac{dv_a}{dt} + c_{22}v_a + b_{21}\frac{dv_r}{dt} + c_{21}v_r &= 0 \end{aligned}\right\} \tag{11}$$

where,

$$\left.\begin{aligned} a_{11} &= L_rC_r & a_{22} &= L_aC_a \\ b_{11} &= J_1L_r + R_rC_r & b_{22} &= J_2L_a + R_aC_a \\ c_{11} &= 1 + J_1R_r & c_{22} &= 1 + J_2R_a \\ b_{12} &= N_1L_r & b_{21} &= N_2L_a \\ c_{12} &= N_1R_r & c_{21} &= N_2R_a \end{aligned}\right\} \tag{12}$$

and,

$$\left.\begin{aligned} J_1 &= \left(\frac{1}{\rho_{II}} - \frac{1}{\rho_I}\right), & N_1 &= \left(\frac{\eta_I}{\rho_I} - \frac{\eta_{II}}{\rho_{II}}\right) \\ J_2 &= \left(\frac{\gamma_{II}}{r_{II}} - \frac{\gamma_I}{r_I}\right), & N_2 &= \left(\frac{1}{r_I} - \frac{1}{r_{II}}\right) \end{aligned}\right\}. \tag{13}$$

To get the solution of these differential equations, we may write v_r and v_a in the form

$$\left.\begin{aligned} v_r &= s_1K\epsilon^{\tau t} \\ v_a &= s_2K\epsilon^{\tau t} \end{aligned}\right\} \tag{14}$$

and from (11) and (14), we obtain

$$\left.\begin{aligned} F_{11}(\tau)s_1 + F_{12}(\tau)s_2 &= 0 \\ F_{21}(\tau)s_1 + F_{22}(\tau)s_2 &= 0 \end{aligned}\right\} \tag{15}$$

where,

$$F(\tau) = a\tau^2 + b\tau + c. \tag{16}$$

Equation (15) is soluble when

$$\begin{vmatrix} F_{11}(\tau) & F_{12}(\tau) \\ F_{21}(\tau) & F_{22}(\tau) \end{vmatrix} = 0. \tag{17}$$

From these relations, the following biquadratic equation can be derived,

$$H_0\tau^4 + H_1\tau^3 + H_2\tau^2 + H_3\tau + H_4 = 0 \tag{18}$$

where,

$$
\left.\begin{aligned}
H_0 &= a_{11}a_{22} \\
H_1 &= \begin{vmatrix} a_{11} & b_{11} \\ -a_{22} & b_{22} \end{vmatrix} \\
H_2 &= \begin{vmatrix} a_{11} & c_{11} \\ -a_{22} & c_{22} \end{vmatrix} + \begin{vmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{vmatrix} \\
H_3 &= \begin{vmatrix} b_{11} & c_{12} \\ b_{21} & c_{22} \end{vmatrix} + \begin{vmatrix} c_{11} & b_{12} \\ c_{21} & b_{22} \end{vmatrix} \\
H_4 &= \begin{vmatrix} c_{11} & c_{12} \\ c_{21} & c_{22} \end{vmatrix}
\end{aligned}\right\} . \tag{19}
$$

Hence, four roots of τ are generally obtained from (18), for three different conditions.

Case A. When all roots of τ are complex numbers let,

$$
\left.\begin{aligned}
\tau &= \delta + j\omega \\
j &= \sqrt{-1}
\end{aligned}\right\} \tag{20}
$$

and,

$$
\left.\begin{aligned}
s_1 &= + |s_1| \epsilon^{j\xi_1} \\
s_2 &= - |s_2| \epsilon^{j\xi_2}
\end{aligned}\right\} \tag{21}
$$

then the solutions of the differential equation in this case, are

$$
\left.\begin{aligned}
v_r = &\ |s_{1\mathrm{I}}| K_{\mathrm{I}} \epsilon^{\delta_{\mathrm{I}} t} \sin(\xi_{1\mathrm{I}} + \kappa_{\mathrm{I}} + \omega_{\mathrm{I}} t) \\
&+ |s_{1\mathrm{II}}| K_{\mathrm{II}} \epsilon^{\delta_{\mathrm{II}} t} \sin(\xi_{1\mathrm{II}} + \kappa_{\mathrm{II}} + \omega_{\mathrm{II}} t) \\
v_a = &- |s_{2\mathrm{I}}| K_{\mathrm{I}} \epsilon^{\delta_{\mathrm{I}} t} \sin(\xi_{2\mathrm{I}} + \kappa_{\mathrm{I}} + \omega_{\mathrm{I}} t) \\
&- |s_{2\mathrm{II}}| K_{\mathrm{II}} \epsilon^{\delta_{\mathrm{II}} t} \sin(\xi_{2\mathrm{II}} + \kappa_{\mathrm{II}} + \omega_{\mathrm{II}} t)
\end{aligned}\right\} . \tag{22}
$$

Case B. When two roots are complex numbers, and the other two are real numbers.

$$
\begin{aligned}
v_r &= s_{1\mathrm{I}} K_{\mathrm{I}} \epsilon^{\tau_{\mathrm{I}} t} + s_{1\mathrm{II}} K_{\mathrm{II}} \epsilon^{\tau_{\mathrm{II}} t} + s_1 K \epsilon^{\delta t} \sin(\xi_1 + \kappa + \omega t) \\
v_a &= s_{2\mathrm{I}} K_{\mathrm{I}} \epsilon^{\tau_{\mathrm{I}} t} + s_{2\mathrm{II}} K_{\mathrm{II}} \epsilon^{\tau_{\mathrm{II}} t} + s_2 K \epsilon^{\delta t} \sin(\xi_2 + \kappa + \omega t) .
\end{aligned} \tag{23}
$$

Case C. When all roots are real numbers the condition for the maintenance of oscillation is obviously

$$
\delta = 0. \tag{24}
$$

In Case A, we get

$$\left.\begin{aligned} v_r &= |s_{1\mathrm{I}}|K_{\mathrm{I}} \sin(\xi_{1\mathrm{I}} + \kappa_{\mathrm{I}} + \omega_{\mathrm{I}} t) \\ &\quad + |s_{1\mathrm{II}}|K_{\mathrm{II}} \sin(\xi_{1\mathrm{II}} + \kappa_{\mathrm{II}} + \omega_{\mathrm{II}} t) \\ v_a &= -\,|s_{2\mathrm{I}}|K_{\mathrm{I}} \sin(\xi_{2\mathrm{I}} + \kappa_{\mathrm{I}} + \omega_{\mathrm{I}} t) \\ &\quad - |s_{2\mathrm{II}}|K_{\mathrm{II}} \sin(\xi_{2\mathrm{II}} + \kappa_{\mathrm{II}} + \omega_{\mathrm{II}} t) \end{aligned}\right\}. \tag{25}$$

When the oscillation frequency of the inner-grid circuit is quite different from that of the anode circuit, there must be the following relations:

$$|s_{1\mathrm{I}}| \gg |s_{2\mathrm{I}}|, \qquad |s_{1\mathrm{II}}| \ll |s_{2\mathrm{II}}|$$

or,

$$|s_{1\mathrm{I}}| \ll |s_{2\mathrm{I}}|, \qquad |s_{1\mathrm{II}}| \gg |s_{2\mathrm{II}}|.$$

When oscillating at the same frequency, equations are simplified by putting

$$\omega_{\mathrm{I}} = \omega_{\mathrm{II}}.$$

Other solutions for this case are

$$s_{1\mathrm{I}} = s_{2\mathrm{I}} = 0$$

or,

$$s_{1\mathrm{II}} = s_{2\mathrm{II}} = 0.$$

Putting $\delta = 0$ in (23), the undamped oscillating voltages are now,

$$\left.\begin{aligned} v_r &= |s_1|K \sin(\xi_1 + \kappa + \omega t) \\ v_a &= |s_2|K \sin(\xi_2 + \kappa + \omega t) \end{aligned}\right\}. \tag{27}$$

When ω_{I} and ω_{II} differ slightly, v_r and v_a can be expressed by means of Taylor's series.

Conditions for the existence of roots in the biquadratic equation can be examined from a textbook of mathematics. So, further discussion of this subject is omitted. Putting $\tau = j\omega$ into (18), the following equation may be deduced,

$$\omega^2 = \frac{H_3}{H_1} = \tag{28}$$
$$\frac{(J_1L_r + R_rC_r)(1 + J_2R_a) + (J_2L_a + R_aC_a)(1 + J_1R_r) - N_1N_2(R_rL_a + R_aL_r)}{L_rC_r(J_2L_a + R_aC_a) + L_aC_a(J_1L_r + R_rC_r)}$$

which represents the oscillation angular frequency provided that this root exists.

If, R_r and R_a are negligibly small compared with other constants, the equation may be reduced to

$$\omega^2 = \frac{J_1 L_r + J_2 L_a}{L_r L_a (J_1 C_a + J_2 C_r)}. \tag{29}$$

As a special case, if $L_r = L_a = L$ and $C_r = C_a = C$

$$\omega^2 = \frac{L(J_1 + J_2)}{L^2 C (J_1 + J_2)} = \frac{1}{LC}. \tag{30}$$

This gives only a single frequency and is the same as the frequency equation of an ordinary triode oscillator.

If M_{ar} is taken into account, the equation will become so complicated that we are probably unable to solve it.

ACKNOWLEDGMENT

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A SCREEN-GRID VOLTMETER WITHOUT EXTERNAL LEAK*

BY

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Summary—*A highly sensitive screen-grid voltmeter using type 24 (or 32) tubes without grid leak is described. Static and transrectification characteristics of single and twin tubes and of tube circuits including a load resistance are shown. The voltmeter is independent of frequency from low audio to ultra-high frequencies. The useful range of the twin (full wave rectifying) circuit is from 0.05 to 20 volts root-mean-square; this can be extended to higher voltages as desired by suitable choice of blocking condensers. The range from 0.1 to 10 volts has a linear calibration curve. Over this range the transrectification factor* $\partial I_t/\partial e_g = 2.05\times 10^{-3}$. *The input admittance of the device is very low. Applications of the circuit for use as a peak voltmeter, as a resonance indicator, as a radio detector, and as a high-frequency intensity meter are briefly discussed.*

IN AN earlier paper[1] a screen-grid voltmeter and its application as a resonance indicator were described. The operation of the device was explained in terms of leaky-grid rectification with the comment that at very high frequencies there is usually enough leakage to make the use of a grid leak unnecessary. More recent studies have revealed that with screen-grid tubes of the 24 and 32 types a grid leak is not at all essential to their functioning as highly sensitive detectors in the wide frequency band from below 60 cycles to above 100 megacycles. An explanation for the excellent rectifying and demodulating action of most tubes of this type when their control grids are isolated to direct currents may be made in terms of secondary emission from the control grid and the presence of positive ions. The combined effect is to discharge the grid much as with a high resistance leak. It is the purpose of this paper to show typical static and transrectification characteristics of single and two tube voltmeters of this type, and to discuss briefly a few applications.

I. CHARACTERISTICS OF 24 TYPE TUBE

A. Static Characters

(a). $I_p - E_d$, $I_d - E_d$, $I_t - E_d$ characteristics with E_p as a parameter and with the control grid isolated were obtained using the circuit of Fig. 1. The curves are shown in Fig. 2.

* Decimal classification: R243.1. Original manuscript received by the Institute, June 27, 1933; revised manuscript received by the Institute, December 12, 1933.

[1] R. King, PROC. I.R.E., vol. 18, p. 1388; August, (1930); A. Hund, "High-Frequency Measurements," p. 157, McGraw-Hill, (1933).

(b). $I_p - E_g$, $I_t - E_g$, $I_g - E_g$ characteristics for a fixed value of E_p and with E_d adjusted to give maximum I_p were obtained using the circuit of Fig. 1 modified by substituting a microammeter for the

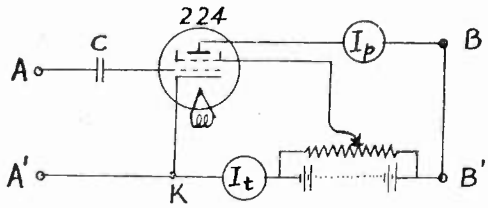

Fig. 1—Circuit for single tube voltmeter.

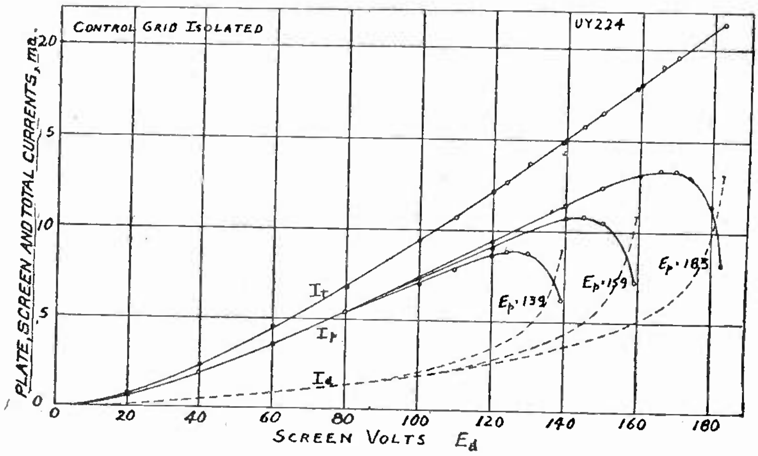

Fig. 2—Static characteristics of 24 tube with control grid isolated.

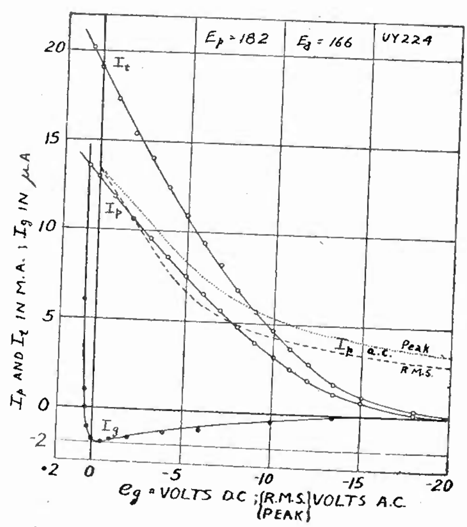

Fig. 3—Solid lines: static characteristics; broken lines: peak and root-mean-square transrectification characteristics.

blocking condenser C. E_g was connected between AA'. The resulting curves are shown in Fig. 3. It was observed that the plate current was the same with the grid isolated as with a positive grid voltage $E_g = 0.45$ volt. This means that in the case of the tube examined, the isolated grid built up a positive potential of 0.45 volt. Different tubes of the same type vary somewhat in this isolated grid voltage.

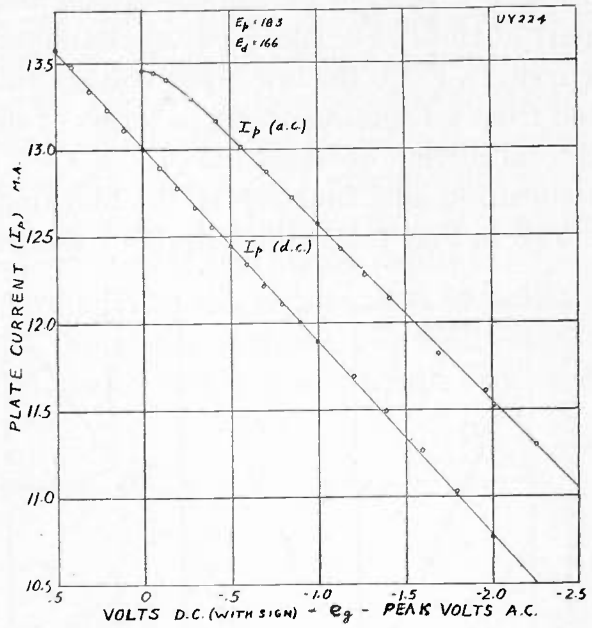

Fig. 4—Low voltage range of two of the characteristics of Fig. 3 replotted to bring out the relation between the direct-current and peak alternating-current curves.

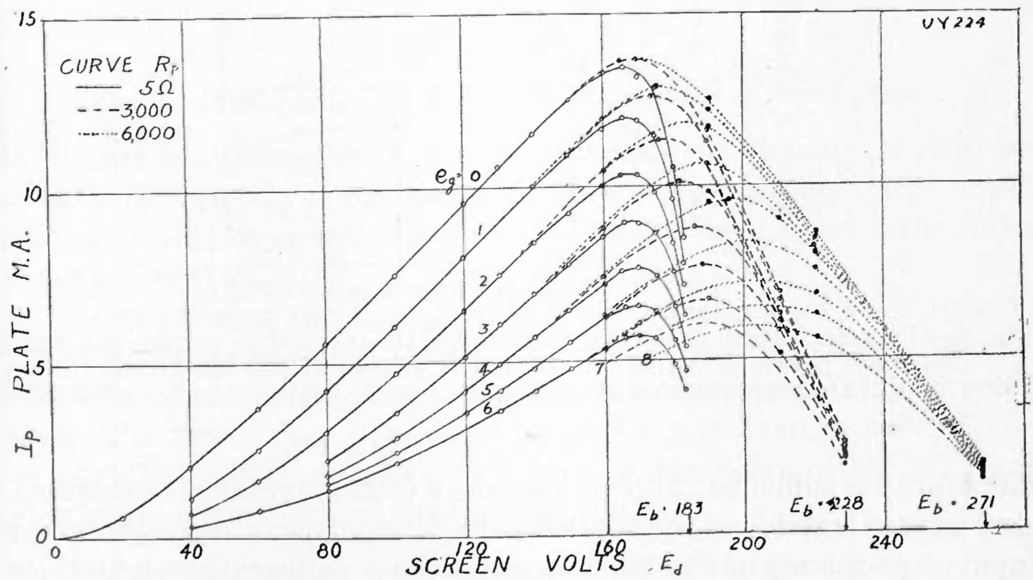

Fig. 5—Transrectification characteristics with indicated load resistance in plate circuit. e_g in volts root-mean-square.

B. Transrectification Characteristics

(a). $I_p - E_d$ charactersitics for a fixed E_p and with the alternating grid voltage e_g as parameter are reproduced in Fig. 5. The 60-cycle voltage was applied across AA' (Fig. 1).

(b). $I_b - e_g$, $I_t - e_g$, $I_d - e_g$ characteristics with E_p fixed and with E_d adjusted to give maximum I_p are shown in Fig. 6. The $I_p - e_g$ curve is also shown in Fig. 3 with e_g in root-mean-square and in peak volts. Fig. 4 shows a part of the $I_p - e_g$ (peak volts) characteristic compared with the static curve. In Fig. 6 the low input voltage range of I_p and I_t are shown plotted from a common origin in terms of plate current decrease (ΔI_p) and total current decrease (ΔI_t).

(c). Transrectification as a function of the blocking capacitance C at 60 cycles is shown in Fig. 7. At this frequency a capacitance of 0.05

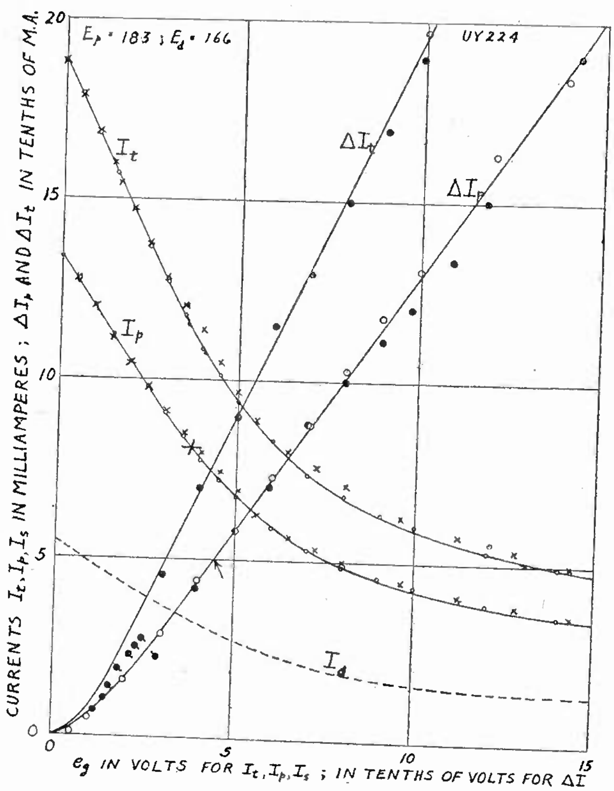

Fig. 6—Transrectification characteristics. On the I_p and I_t curves the small circles are points taken at 60 cycles, the crosses at 750 kilocycles. On the ΔI_p and ΔI_t curves the circles are at 60 cycles, the solid dots at 750 kilocycles.

microfarad is sufficient; at 750 kilocycles 0.0002 microfarad is adequate. Any desired fraction of the applied voltage may be impressed across the input capacitance of the tube by selecting a sufficiently small value of C. The response to changes in e_g becomes sluggish for C greater than 0.1 microfarad.

(d). Transrectification as a function of frequency is indicated in Fig. 6. Here the calibration points at 60 cycles and at 750 kilocycles are shown. The agreement is excellent considering the difficulties encountered in making a precise radio-frequency calibration. The calibration was rechecked at 5000 cycles with equally good agreement. In consequence one may conclude that transrectification is independent of frequency over the audio- and radio-frequency band. It may be added that at a frequency of 75 megacycles, where a direct calibration is extremely difficult if at all possible, this same calibration characteristic is probably correct. This conclusion is based on the following observations. It has been pointed out[2] that at this high frequency the transrectification characteristic $(I_p - e_g)$ is a straight line up to input voltages which produce a decrease in plate current of 5 milliamperes. As the input voltage is increased to still higher values the $I_p - e_g$ curve flattens out to the same value at 75 megacycles as at 60 cycles. It seems reasonable to suppose, therefore, that the two characteristics which start at the same point, which continue as straight lines as the input voltage

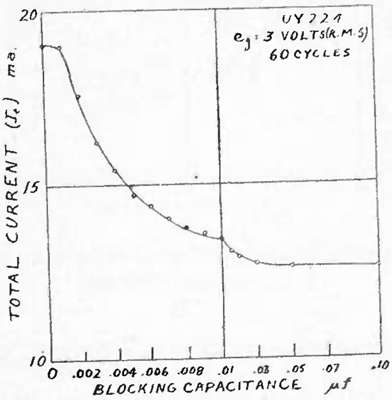

Fig. 7—I_t as a function of blocking capacitance at 3 volts input.

is increased to produce the first 5-milliampere deflection in I_p, and which eventually approach the same value for increasingly large input voltages, must be essentially, if not identically the same over the entire input voltage range.

To summarize, the $I_p - e_g$ and $I_t - e_g$ characteristics have a useful range for values of e_g extending from below 0.05 volt to above 10 volts root-mean-square; they are linear over the range from 0.1 to 5 volts. Over this linear part typical transrectification factors are: $\partial I_p/\partial e_g = 1.42 \times 10^{-3}$; $\partial I_t/\partial e_g = 2.05 \times 10^{-3}$, for the conditions of Fig. 6. These characteristics are independent of frequency.

C. Circuit Characteristics with Resistance in the Plate Circuit

(a). $I_p - E_d$ characteristics with e_g and R as parameters are shown in Fig. 5 in broken lines. Curves are given with $R = 3000$ and 6000 ohms. It is significant to note that with E_d less than the value which gives maximum I_p, the plate current is independent of both e_p and R.

[2] R. King, Proc. I.R.E., vol. 21, p. 1178; August, (1933).

This is a consequence of the extremely low plate conductance of the screen-grid tube.

(b). $I_p - e_0$ characteristics with E_d as parameter for fixed values of R and E_d are shown in Fig. 8.

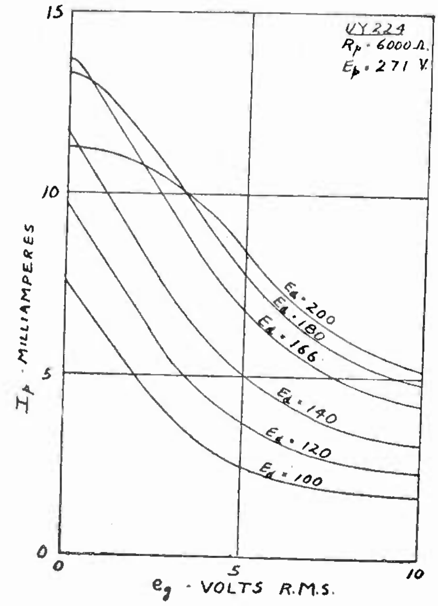

Fig. 8—Transrectification characteristics of a single tube circuit with 6000 ohms in the plate circuit.

II. Characteristics of Two Tubes in Push-Pull

Fig. 9 shows the circuit used to connect two 24 type tubes in a push-pull arrangement using no grid leak. The input is connected to AA';

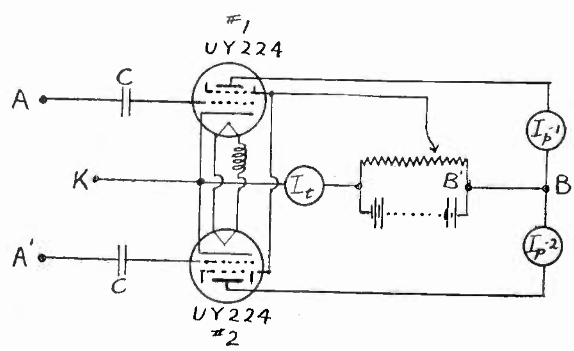

Fig. 9—Circuit for two tubes in push-pull arrangement.

it may be center-tapped to the cathode K. Transrectification characteristics for this circuit are shown in Figs. 10 and 11. In both figures characteristics are also given for the individual tubes, since it is desirable that these be as nearly alike as possible. Upon comparing corresponding characteristics it is evident that those of the two-tube circuit are like those of the single tube circuit but with the coördinates doubled. In other words the transrectification factor is the same in each case, but the input voltage range for two tubes is twice that for one tube.

Characteristics with a resistance in the plate circuit are not shown for the push-pull arrangement, since they hardly differ from the curves for a single tube with coördinates doubled. In the two-tube circuit the input capacitances of the two screen-grid tubes are in series so that smaller blocking capacitances may be used than for one tube alone.

The following may well be advantages of the two-tube circuit over the single tube one: 1. Twice the input voltage range with the same sensitivity; 2. full wave rectification; 3. smaller input capacitance; 4. completely symmetrical input circuit.

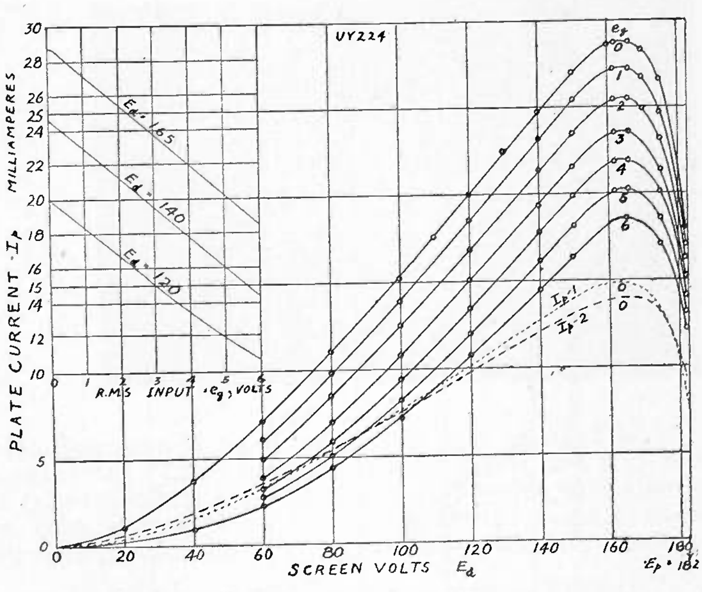

Fig. 10—Transrectification characteristics of two tubes with e_g in volts root-mean-square. The broken lines are the static characteristics for the separate tubes; their ordinates when added give the $e_g = 0$ curve for the two tubes.

III. Applications

A. A Vacuum Tube Voltmeter

Using either single or twin arrangements the $I_p - e_g$ or $I_t - e_g$ characteristics may be used advantageously as calibration curves for a vacuum tube voltmeter. Or, if desired, the voltage drop across a load resistance in the plate circuit may serve as a measure of the input voltage. In either case the device has practically the same sensitivity from low audio to ultra-high frequencies. Since it is a peak voltmeter it is subject to wave form correction for nonsinusoidal voltages.[3] This correc-

[3] W. B. Medlam and U. A. Oschwald, *Exp. Wireless and Wireless Eng.*, vol. 8, p. 596, (1926).

tion is considerably less in the full wave rectifying circuit. The general features of the screen-grid voltmeter without external leak are much like those of the same device with grid leak as described.[1] In the one case the blocking capacitance is made sufficiently small so that primary grid electrons and positive ions are effective in discharging the grid; in the other case the blocking capacitance is so large that electrons and ions cannot adjust the grid potential to variations of the input voltage sufficiently rapidly, so that a grid leak is required. At low frequencies the grid-leak arrangement is probably to be preferred since operation does not depend upon the somewhat variable grid emission and gas content of different tubes. At ultra-high frequencies where resistance is definitely a function of frequency, and where the reactance

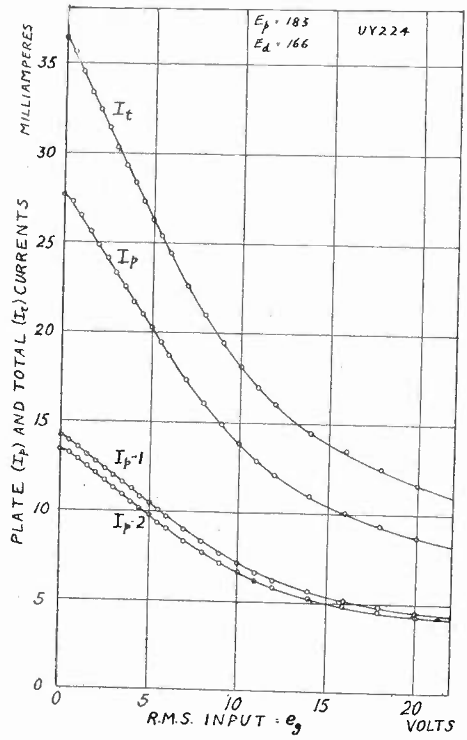

Fig. 11—Transrectification characteristics; I_p and I_t for two tubes; the curves I_p-1 and I_g-2 are the individual curves which when combined give the I_p curve.

of grid-leak mountings and connections becomes significant, the circuit without grid leak may have very great advantages.

B. A Symmetrical Resonance Indicator for Parallel Wires.

As a resonance indicator the symmetrical push-pull voltmeter is very much superior to the single tube circuit. Not only is its linear range greater and its input capacitance smaller, but asymmetrical capacitance to earth is eliminated. Connected across the ends of a parallelwire system with suitably chosen blocking capacitances, it serves as a very high impedance voltmeter even at frequencies as high as 100 megacycles.

C. A High-Frequency Field Intensity Meter

As a field intensity meter at ultra-high frequencies the single tube circuit is useful with a vertical antenna connected directly to A (Fig. 1) and with A' grounded. The two-tube circuit may be used with one half of the antenna erected vertically upward from one control grid and the other half vertically downward from the other control grid.

D. As a Detector and Demodulator.

Both circuits function excellently as radio detectors with properly selected tubes. Tests were made at broadcast wavelengths and at 4

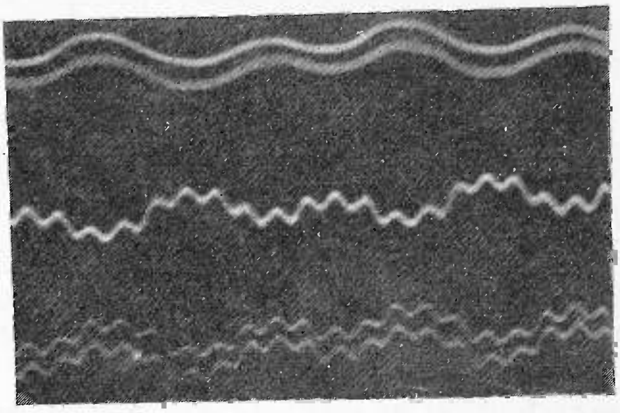

Fig. 12—Oscillograms of modulating and demodulated waves. The carrier was 75 megacycles. The smooth oscillograms at the top are the modulating wave (above) composed of 100 and 200 cycles with the demodulated wave below. The two waves at the bottom are the same as those at the top with 1000 cycles in addition. The center shows the two lower waves superposed.

meters with a single stage audio amplifier. With only a small loop neighboring stations were received with excellent quality and loud speaker volume. On Fig. 6 the arrow at $\Delta I_p = 0.5$ milliampere indicates the deflection due to the carrier of the local station using the loop. One end of the loop with its parallel tuning condenser was connected to A (Fig. 1); the other end was left disconnected since the capacitance of the loop to ground proved a very effective return circuit. Even for the low voltage input indicated by the arrow in Fig. 6, the transrectification characteristic is essentially linear so that little distortion takes place.[4]

Demodulation tests were made using a modulated 75-megacycle carrier and the single tube detector without amplifier. With the transmitter at the far end of the laboratory the detector input voltage is

[4] C. B. Aiken, Proc. I.R.E., vol. 21, p. 601; April, (1933).

indicated by the large cross in Fig. 6. Since the total variation in plate current with the modulated carrier was not over 1 milliampere, detection was practically linear. From this deflection the modulation was easily computed to be about 16 per cent. A typical oscillogram is shown in Fig. 12.

As a general conclusion it may be said that the screen-grid voltmeter without external leak has a frequency range, an input voltage range, and a sensitivity which compare favorably with other types of tube voltmeters. Whether it is to be preferred to a simpler triode type voltmeter with leaky grid depends upon particular requirements and conditions.

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HARMONIC PRODUCTION AND CROSS MODULATION IN THERMIONIC VALVES WITH RESISTIVE LOADS*

By

D. C. Espley

(Communication from the Staff of the Research Laboratories, The General Electric Company, Limited, Wembley, England)

Summary—*The harmonic and modulation components in the anode current of a thermionic valve, operating with a resistive load, are obtained in terms of coefficients of the power series by which the load characteristic is expressed, and amplitudes of the voltages applied to the control grid. An example is given in which it is shown that the presence of a large signal on the grid of an output pentode greatly increases the distortion of a small signal of different frequency.*

I

IT HAS been shown[1] that the load characteristic of a valve can be represented as a power series

$$i_a = a_0' + a_1'k + a_2'k^2 + \cdots + a_s'k^s = F(k) \tag{1}$$

in which $k = e_g/\Delta e_g$ and $a_0'a_1'a_2' \cdots a_s'$ are obtainable directly in terms of points on this load characteristic. e_g is the variable grid voltage measured from the working point, and it is assumed that the $E_a - i_a$ characteristics of the valve are available, as is usual, at equal intervals Δe_g of grid voltage. (i_a is used throughout for the anode current to avoid confusion with the Bessel coefficients I_n, etc.)

If we apply to the grid a voltage

$$(v_1 \cos \omega_1 t + v_2 \cos \omega_2 t + \cdots + v_r \cos \omega_r t)$$

then the components of the anode current i_a will be given by expansion of the expression

$$i_a = a_0' + a_1' \frac{(v_1 \cos \omega_1 t + \cdots + v_r \cos \omega_r t)}{\Delta e_g}$$
$$+ a_2' \frac{(v_1 \cos \omega_1 t + \cdots + v_r \cos \omega_r t)^2}{(\Delta e_g)^2} + \cdots. \tag{2}$$

If for simplicity we write

* Decimal classification: R135. Original manuscript received by the Institute, December 11, 1933.

[1] D. C. Espley, "The calculation of harmonic production in thermionic valves with resistive loads," Proc. I.R.E., vol. 21, pp. 1439–1446; October, (1933).

$$\frac{v_1}{\Delta e_g} = V_1$$

$$\frac{v_2}{\Delta e_g} = V_2 \text{ etc.}$$

$$\omega_1 t = \theta_1$$

$$\omega_2 t = \theta_2 \text{ etc.}$$

then putting

$$k = (V_1 \cos \theta_1 + \cdots V_r \cos \theta_r)$$

$$i_a = a_0' + a_1'(V_1 \cos \theta_1 + \cdots + V_r \cos \theta_r) + a_2'(V_1 \cos \theta_1 + \cdots + V_r \cos \theta_r)^2 + \cdots. \quad (3)$$

It is possible to expand all the brackets and replace all the powers of cosines by their expansions in unit powers of cosines of multiple angles, and then to pick out the cosine of $(n_1\theta_1 + n_2\theta_2 + \cdots + n_r\theta_r)$ which will give the modulation product of frequency $(n_1\omega_1 + n_2\omega_2 + \cdots + n_r\omega_r)/2\pi$. Such a process is extremely laborious and the result is obtained much more directly in the following way.[2] We can use Maclaurin's expansion.

$$i_a = F(0) + \sum_{n=1}^{\infty} \frac{1}{n!}(V_1 \cos \theta_1 + \cdots + V_r \cos \theta_r)^n \frac{d^n}{dk^n} F(0). \quad (4)$$

It will be seen that

$$\left.\begin{aligned} \frac{d}{dk} F(0) &= a_1' \\ \frac{d^2}{dk^2} F(0) &= 2a_2' \\ &\cdots\cdots\cdots \\ \frac{d^n}{dk^n} F(0) &= n!a_n' \end{aligned}\right\}. \quad (5)$$

Our final object is to obtain the multiple Fourier series for i_a and this can be obtained directly from the symbolic form of Maclaurin's theorem

$$F(V_1 \cos \theta_1 + \cdots + V_r \cos \theta_r) = \epsilon^{V_1 \cos \theta_1 (d/dk)} \epsilon^{V_2 \cos \theta_2 (d/dk)} \cdots \epsilon^{V_r \cos \theta_r (d/dk)} F(0). \quad (6)$$

The operators of the form

[2] A. C. Bartlett, "The calculation of modulation products," *Phil. Mag.*, vol. 16, pp. 845–847; October, (1933).

$$\epsilon^{V \cos \theta (d/dk)}$$

can be transformed using the Sonine expansion

$$\epsilon^{z \cos \theta} = I_0(z) + 2 \sum_{n=1}^{\infty} I_n(z) \cos n\theta \tag{7}$$

in which $I_n(z)$ is the modified Bessel coefficient

$$I_n(z) = \sum_{p=0}^{\infty} \frac{1}{p!(n+p)!} \left(\frac{z}{2}\right)^{n+2p}. \tag{8}$$

Thus,

$$\epsilon^{V \cos \theta (d/dk)} = I_0\left(V \frac{d}{dk}\right) + 2 \sum_{n=1}^{\infty} \cos n\theta \, I_n\left(V \frac{d}{dk}\right). \tag{9}$$

This expansion for the operators enables us to transform (4), for the anode current, into the required multiple Fourier series

$$\begin{aligned} i_a &= F(V_1 \cos \theta_1 + \cdots + V_r \cos \theta_r) \\ &= \left\{ I_0\left(V_1 \frac{d}{dk}\right) + 2 \sum_{n=1}^{\infty} \cos n_1\theta_1 \, I_{n_1}\left(V_1 \frac{d}{dk}\right) \right\} \\ &\times \left\{ I_0\left(V_2 \frac{d}{dk}\right) + 2 \sum_{n=1}^{\infty} \cos n_2\theta_2 \, I_{n_2}\left(V_2 \frac{d}{dk}\right) \right\} \\ &\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots \\ &\times \left\{ I_0\left(V_r \frac{d}{dk}\right) + 2 \sum_{n=1}^{\infty} \cos n_r\theta_r \, I_{n_r}\left(V_r \frac{d}{dk}\right) \right\} F(0). \end{aligned} \tag{10}$$

The following important results appear directly from (10).

(a) *The constant mean value* of the anode current is

$$I_0\left(V_1 \frac{d}{dk}\right) I_0\left(V_2 \frac{d}{dk}\right) \cdots I_0\left(V_r \frac{d}{dk}\right) F(0) \tag{11}$$

(b) *The general modulation term.*

The coefficient of $\cos n_1\theta_1 \cos n_2\theta_2 \cdots \cos n_r\theta_r$ is

$$2^r I_{n_1}\left(V_1 \frac{d}{dk}\right) I_{n_2}\left(V_2 \frac{d}{dk}\right) \cdots I_{n_r}\left(V_r \frac{d}{dk}\right) F(0). \tag{12}$$

(c) *Fundamental frequency terms* ($n_q = 1$) *and simple harmonics* of the form $\cos n_q\theta_q$ have a coefficient

$$\frac{2 I_{n_q}\left(V_q \frac{d}{dk}\right)}{I_0\left(V_q \frac{d}{dk}\right)} \times I_0\left(V_1 \frac{d}{dk}\right) I_0\left(V_2 \frac{d}{dk}\right) \cdots I_0\left(V_r \frac{d}{dk}\right) F(0). \tag{13}$$

II. The Case with Two Frequencies—General Terms

We have

$$i_a = F(V_1 \cos \theta_1 + V_2 \cos \theta_2). \tag{14}$$

(a) *Components of the form* $\cos n_1\theta_1$ have a coefficient.

$$2I_0\left(V_2\frac{d}{dk}\right) I_{n_1}\left(V_1\frac{d}{dk}\right)F(0). \tag{15}$$

Expanding the two Bessel coefficients according to (8) we may write (15) as

$$2\left\{1 + \left(\frac{V_2\frac{d}{dk}}{2}\right)^2 + \frac{1}{4}\left(\frac{V_2\frac{d}{dk}}{2}\right)^4 + \frac{1}{36}\left(\frac{V_2\frac{d}{dk}}{2}\right)^6 + \cdots\right\}$$

$$\times \left\{\frac{1}{n_1!}\left(\frac{V_1\frac{d}{dk}}{2}\right)^{n_1} + \frac{1}{(n_1+1)!}\left(\frac{V_1\frac{d}{dk}}{2}\right)^{n_1+2}\right.$$

$$\left. + \frac{1}{2(n_1+2)!}\left(\frac{V_1\frac{d}{dk}}{2}\right)^{n_1+4} + \cdots\right\}F(0). \tag{16}$$

After collecting terms (16) becomes

$$2\left\{\frac{d^{n_1}}{dk^{n_1}}\left(\frac{V_1^{n_1}}{2^{n_1}n_1!}\right) + \frac{d^{n_1+2}}{dk^{n_1+2}}\left(\frac{V_1^{n_1+2}}{2^{n_1+2}(n_1+1)!} + \frac{V_1^{n_1}V_2^2}{2^{n_1+2}n_1!}\right)\right.$$

$$+ \frac{d^{n_1+4}}{dk^{n_1+4}}\left(\frac{V_1^{n_1+4}}{2^{n_1+4}(n_1+2)!2} + \frac{V_1^{n_1+2}V_2^2}{2^{n_1+4}(n_1+1)!}\right.$$

$$\left.\left. + \frac{V_1^{n_1}V_2^4}{2^{n_1+4}n_1!4}\right) + \cdots\right\}F(0). \tag{17}$$

Knowing that

$$\left(\frac{d^nF(k)}{dk^n}\right)_{k=0} = n!a_n'$$

we may write the coefficient

$$2\left\{\frac{n_1!}{2^{n_1}}\left(\frac{V_1^{n_1}}{n_1!}\right)a'_{n_1} + \frac{(n_1+2)!}{2^{n_1+2}}\left(\frac{V_1^{n_1+2}}{(n_1+1)!} + \frac{V_1^{n_1}V_2^2}{n_1!}\right)a'_{n_1+2}\right.$$

$$+ \frac{(n_1+4)!}{2^{n_1+4}}\left(\frac{V_1^{n_1+4}}{(n_1+2)!2} + \frac{V_1^{n_1+2}V_2^2}{(n_1+1)!} + \frac{V_1^{n_1}V_2^4}{n_1!4}\right)a'_{n_1+4}$$

$$\left. + \cdots\cdots\cdots\cdots\cdots\cdots\right\}. \tag{18}$$

It will be clear that by substituting n_2 for n_1 and interchanging V_1 and V_2 equation (18) will then represent the coefficient of multiples of θ_2 instead of multiples of θ_1. The fundamental, in either case, is obtained by putting $n=1$, and the harmonic production with a single frequency applied will be given by equation (18) (with respect to V_1 or V_2) after putting either V_2 or V_1 zero.

(b) *The modulation component* $\cos n_1\theta_1 \cos n_2\theta_2$ has a coefficient

$$4I_{n_1}\left(V_1\frac{d}{dk}\right) I_{n_2}\left(V_2\frac{d}{dk}\right)F(0). \tag{19}$$

After expansion this becomes

$$\begin{aligned}
4\Bigg\{&\frac{1}{2^{n_1+n_2}}\left(\frac{V_1^{n_1}V_2^{n_2}}{n_1!n_2!}\right)\frac{d^{n_1+n_2}}{dk^{n_1+n_2}}\\
&+\frac{1}{2^{n_1+n_2+2}}\left(\frac{V_1^{n_1+2}V_2^{n_2}}{(n_1+1)!n_2!}+\frac{V_1^{n_1}V_2^{n_2+2}}{n_1!(n_2+1)!}\right)\frac{d^{n_1+n_2+2}}{dk^{n_1+n_2+2}}\\
&+\frac{1}{2^{n_1+n_2+4}}\left(\frac{V_1^{n_1+4}V_2^{n_2}}{(n_1+2)!n_2!2!}+\frac{V_1^{n_1+2}V_2^{n_2+2}}{(n_1+1)!(n_2+1)!}\right.\\
&\left.+\frac{V_1^{n_1}V_2^{n_2+4}}{n_1!(n_2+2)!2!}\right)\frac{d^{n_1+n_2+4}}{dk^{n_1+n_2+4}}+\cdots\Bigg\}F(0)
\end{aligned} \tag{20}$$

$$\begin{aligned}
=4\Bigg\{&\frac{(n_1+n_2)!}{2^{n_1+n_2}}\left(\frac{V_1^{n_1}V_2^{n_2}}{n_1!n_2!}\right)a'_{n_1+n_2}\\
&+\frac{(n_1+n_2+2)}{2^{n_1+n_2+2}}!\left(\frac{V_1^{n_1+2}V_2^{n_2}}{(n_1+1)!n_2!}+\frac{V_1^{n_1}V_2^{n_2+2}}{n_1!(n_2+1)!}\right)a'_{n_1+n_2+2}\\
&+\frac{(n_1+n_2+4)}{2^{n_1+n_2+4}}!\left(\frac{V_1^{n_1+4}V_2^{n_2}}{(n_1+2)!n_2!2!}+\frac{V_1^{n_1+2}V_2^{n_2+2}}{(n_1+1)!(n_2+1)!}\right.\\
&\left.+\frac{V^{n_1}V_2^{n_2+4}}{n_1!(n_2+2)!2!}\right)a_{n_1+n_2+4}+\cdots\Bigg\}.
\end{aligned} \tag{21}$$

III. The Case of Two Frequencies—Particular Terms (Simple Practical Results)

All expressions are given with respect to V_1 but, as mentioned above, interchange of V_1 and V_2 will give the corresponding expressions for V_2.

(a) *The fundamental component with a single frequency applied.*

$$\cos\theta_1\left\{V_1a_1'+\frac{3V_1^3a_3'}{4}+\frac{10V_1^5a_5'}{16}+\frac{35V_1^7a_7'}{64}+\cdots\right\}. \tag{22}$$

(b) *The fundamental component with two frequencies applied.*

$$\cos\theta_1\left\{V_1a_1' + \frac{1}{4}(3V_1^3 + 6V_1V_2^2)a_3' + \frac{1}{16}(10V_1^5 + 60V_1^3V_2^2 + 30V_1V_2^4)a_5' + \frac{1}{64}(35V_1^7 + 420V_1^5V_2^2 + 630V_1^3V_2^4 + 140V_1V_2^6)a_7' + \cdots\right\}. \tag{23}$$

(c) *The second harmonic component with a single frequency applied.*

$$\cos 2\theta_1\left\{\frac{V_1^2a_2'}{2} + \frac{4V_1^4a_4'}{8} + \frac{15V_1^6a_6'}{32} + \frac{56V_1^8a_8'}{128} + \cdots\right\}. \tag{24}$$

(d) *The second harmonic component with two frequencies applied.*

$$\cos 2\theta_1\left\{\frac{V_1^2a_2'}{2} + \frac{1}{8}(4V_1^4 + 12V_1^2V_2^2)a_4' + \frac{1}{32}(15V_1^6 + 120V_1^4V_2^2 + 90V_1^2V_2^4)a_6' + \cdots\right\}. \tag{25}$$

(e) *The third harmonic component with a single frequency applied.*

$$\cos 3\theta_1\left\{\frac{V_1^3a_3'}{4} + \frac{5V_1^5a_5'}{16} + \frac{21V_1^7a_7'}{64} + \cdots\right\}. \tag{26}$$

(f) *The third harmonic component with two frequencies applied.*

$$\cos 3\theta_1\left\{\frac{V_1^3a_3'}{4} + \frac{1}{16}(5V_1^5 + 20V_1^3V_2^2)a_5' + \frac{1}{64}(21V_1^7 + 210V_1^5V_2^2 + 210V_1^3V_2^4)a_7' + \cdots\right\}. \tag{27}$$

IV. General Considerations

If the case arises in which multiples of the frequencies, or multiples of combinations of the frequencies, have any common values

$$\{n_1\theta_1 \pm n_2\theta_2 \pm \cdots \pm n_r\theta_r\} = \pm\{n_1'\theta_1 \pm n_2'\theta_2 \pm \cdots \pm n_r'\theta_r'\} \tag{28}$$

in which the positive numbers $n_1n_2 \cdots n_r$ and $n_1'n_2' \cdots n_r'$ can have any values limited only by the conditions

$$0 \leqq (n_1 + n_2 + \cdots + n_r) \leqq S$$
$$0 \leqq (n_1' + n_2' + \cdots + n_r') \leqq S \qquad (29)$$

where S is the highest power in (1), then terms will appear whose coefficients may not be included in such coefficients as (11), (12), etc.

For example, it will be seen from the coefficient of the general modulation term $\cos n_1\theta_1 \cos n_2\theta_2 \cdots \cos n_r\theta_r$ that the value of the constant component (11) of the anode current is not determined explicitly, as other constant terms can arise if $\{n_1\theta_1 \pm n_2\theta_2 \pm \cdots \pm n_r\theta_r\} = 0$ and in which $n_1 n_2 \cdots n_r$ have all possible values limited only by the condition that

$$2 \leqq (n_1 + n_2 + \cdots + n_r) \leqq S. \qquad (30)$$

If three frequencies are applied the analysis should cover the case of demodulation of a modulated signal, but this ambiguity exists concerning the coefficient of difference frequencies if these frequencies come within the condition defined above.

The normal modulated signal

$$V(1 + m \cos \beta) \cos \alpha \qquad (31)$$

may be written as

$$V_1 \cos \theta_1 + V_2 \cos \theta_2 + V_3 \cos \theta_3 \qquad (32)$$

in which θ_2 represents the carrier frequency, and θ_1 and θ_3 the side frequencies. It is clear that $n(\theta_1 + \theta_3)$ will always be equal to $2n(\theta_2)$ and hence $\cos\{n\theta_1 - 2n\theta_2 + n\theta_3\}$ will always be a constant term, the amplitude of which does not appear in the coefficient

$$I_0\left(V_1 \frac{d}{dk}\right) I_0\left(V_2 \frac{d}{dk}\right) I_0\left(V_3 \frac{d}{dk}\right) F(0). \qquad (33)$$

The first order modulation arises from the products $\cos \theta_1 \cos \theta_2$ and $\cos \theta_2 \cos \theta_3$. These can be calculated separately, just as if the frequencies were incommensurable, and the coefficients of the same frequencies in the expansion should then be added. Either of these terms is of the form

$$\cos \theta_1 \cos \theta_2 \{a_2' b_2 + a_4' b_4 + a_6' b_6 + \cdots\} \qquad (34)$$

in which a_2' a_4' a_6' etc., are the coefficients of the polynomial (1).

It has been shown that $\cos\{n\theta_1 - 2n\theta_2 + n\theta_3\}$ will always be a constant term, and hence its introduction into the $(\theta_1 \pm \theta_2)$ term will always produce this same frequency. This also applies to $(\theta_2 \pm \theta_3)$.

A constant term could also be produced from $\cos\{n_1\theta_1 - n_2\theta_2\}$ or $\cos\{n_2\theta_2 - n_3\theta_3\}$ and similar terms, but in most cases of modulated waves the carrier frequency is much greater than the side band range and hence

$$(\theta_1 - \theta_2) \ll \theta_2 \quad \text{and} \quad (\theta_2 - \theta_3) \ll \theta_2.$$

Thus one or more of the numbers $n_1 n_2$ and n_3 would have to be large and the first term in the coefficients would probably require a term beyond the last in the polynomial expression for i_a. The number of points available on the load characteristic is usually comparatively small and, if $(n_1+n_2)>S$ and $(n_2+n_3)>S$, neither of the terms can appear in the calculations. This follows from the coefficients which are, of the form

$$\left\{a'_{n_1+n_2}b_{n_1+n_2} + a'_{n_1+n_2+2}b_{n_1+n_2+2} + \cdots\right\}. \tag{35}$$

It then follows that, with comparatively low orders of modulation, the only other components of frequency

$$\frac{1}{2\pi t}(\theta_1 - \theta_2) \quad \text{and} \quad \frac{1}{2\pi t}(\theta_2 - \theta_3)$$

are the terms of this frequency to which the zero frequency

$$\frac{1}{2\pi t}(n\theta_1 - 2n\theta_2 + n\theta_3)$$

has been added.

These terms are then given by either

$$\cos (n+1)\theta_1 \cos (2n+1)\theta_2 \cos n\theta_3$$

or,

$$\cos n\theta_1 \cos (2n+1)\theta_2 \cos (n+1)\theta_3$$

which have coefficients of the form

$$\left\{a'_{4n+2}b_{4n+2} + a'_{4n+4}b_{4n+4} + \cdots\right\}. \tag{36}$$

Thus if the polynomial for i_a is a quadratic $n=0$ and there will be only the original two terms of frequency $(\theta_1-\theta_2)/2\pi t$ or $(\theta_2-\theta_3)/2\pi t$. No other components will arise until a sixth degree polynomial is reached (i.e., seven points on the load characteristic, and $n=1$) when two new terms of frequency $(2\theta_1-3\theta_2+\theta_3)/2\pi t$ or $(\theta_1-3\theta_2+2\theta_3)/2\pi t$ must be added to give the complete coefficient of the simple difference frequency.

V. Numerical Example

Fig. 1 shows the characteristics of a pentode valve and seven points are available on the load characteristic. The seven values of the anode current at equally spaced intervals of grid voltage are as follows:

$$
\begin{aligned}
i_7 &= 48 \text{ ma} \\
i_6 &= 43 \text{ “} \\
i_5 &= 33 \text{ “} \\
i_4 &= 23 \text{ “} \\
i_3 &= 14 \text{ “} \\
i_2 &= 7 \text{ “} \\
i_1 &= 3 \text{ “}
\end{aligned}
$$

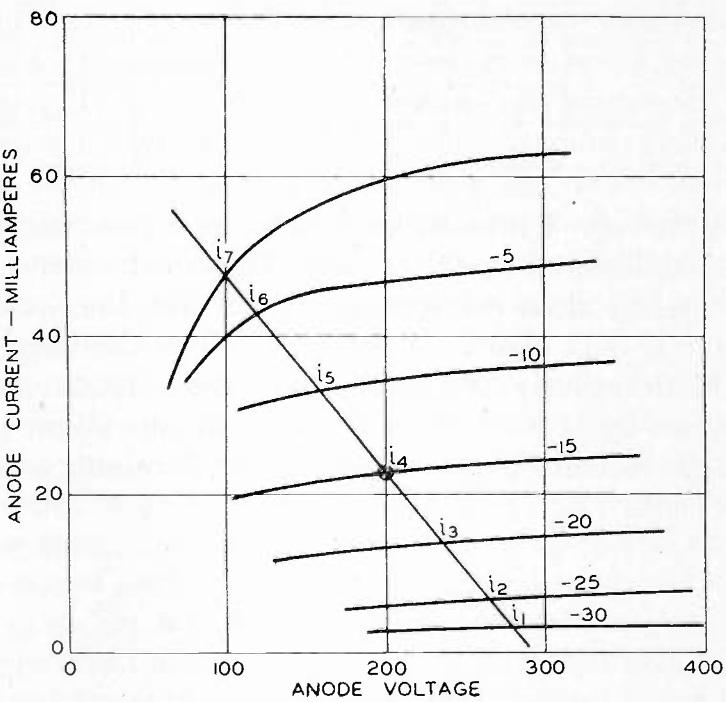

Fig. 1—Typical load line drawn on pentode characteristics

The coefficients of the polynomial (1) were worked out using the method referred to above.

$$
\begin{aligned}
a_0' &= +\ 23.0 \\
a_1' &= +\ \ 9.6 \\
a_2' &= +\ \ 0.45 \\
a_3' &= -\ \ 0.833 \\
a_4' &= +\ \ 0.0278 \\
a_5' &= -\ \ 0.0167 \\
a_6' &= -\ \ 0.00556.
\end{aligned}
$$

If two voltages $v_1 \cos \omega_1 t$ and $v_2 \cos \omega_2 t$ are applied to the control grid, then, if $v_1 = 2.5$ volts peak and $v_2 = 12.5$ volts peak, the substitution of the above coefficients and $V_1=0.5$ and $V_2=2.5$ into (21) to (27) will give the results tabulated in the Table.

TABLE I

Term	Amplitudes (milliamperes peak)		
	V_1 only applied	V_2 only applied	V_1 and V_2 applied
$\cos\theta_1$	4.71		3.23
$\cos\theta_2$		22.0	21.67
or, $\left.\begin{matrix}\cos(\theta_1+\theta_2)\\ \cos(\theta_1-\theta_2)\end{matrix}\right\}$			0.33
$\cos 2\theta_1$	0.065		0.876
$\cos 2\theta_2$		1.314	1.169
$\cos 3\theta_1$	−0.00276		−0.0191
$\cos 3\theta_2$		−0.834	−0.915

These figures are of interest in showing how great can be the distortion of a small signal by a large one. Thus in the above case when the small signal V_1 alone is impressed on the grid, the percentage second harmonic is only about 1.4 per cent. When the larger signal V_2 is applied simultaneously, the amplitude of the output from the small signal is reduced by 31 per cent; it carries more than 20 per cent modulation from the second signal, and the second harmonic is now greater than 25 per cent.

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NOTE ON AN IONIZED GAS MODULATOR FOR SHORT RADIO WAVES*

By

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OSCILLATORS which are largely dependent on applied voltages rather than circuit constants for their oscillation frequency, such as Barkhausen-Kurz and magnetron oscillators, have been widely used for the generation of ultra-short waves. A number of systems for modulating these oscillators has been proposed. In general, amplitude modulation of an oscillator, whose frequency is so dependent on external factors, entails a considerable amount of attendant frequency modulation. The system of modulation which is described in this paper has for one of its main objects the elimination of the frequency modulation which accompanies the amplitude modulation in these transmitters.

If a continuous unmodulated source of high-frequency electric waves is operating, the frequency changes are limited to those dependent on the normal stability of the circuit. If the radiation from this source is formed into a beam and this beam is intercepted by some material whose absorption for the waves can be varied, and which reflects a limited amount back on the source so that the reaction is small, the frequency modulation will be limited to that taking place due to the change of phase of the wave in passing through the modulator.

It has been found that a gas changes its absorption and index of refraction for electromagnetic waves when its degree of ionization is changed. This phenomenon can be readily applied to the modulation of high-frequency electric waves. A diagram of apparatus which was used to do this is shown in Fig. 1. The modulators investigated consisted of spherical glow discharge tubes of dimensions comparable to several wavelengths, and containing noble gases, such as neon, argon, and helium. A magnetron oscillator, T, generating waves of length approximately nine centimeters served as a source. The radiation was beamed with a parabolic reflector. The ionic gas modulator, M, was placed in the beam somewhere between transmitter and receiver. A direct-current polarizing voltage was applied across the tube so as to ionize the gas, and modulation was obtained by increasing and decreasing the ion density above and below the mean value by means of an

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alternating current from a source, S, superposed on this direct current. The direct current varied with different tubes, from 3 to 30 milliamperes. The relative parts played by changing the paths of the waves, and by direct absorption in the tube, in determining the amplitude modulation at the receiver, depends on the shape of the gas discharge. In most of the tubes which have been used so far, both these effects were taking place.

The quality of the modulation obtained with this apparatus is equal to that obtained by more direct means. It is found that the adjustment of the oscillator, which for direct modulation may be somewhat critical or difficult to maintain, is quite easily maintained when no modulation is applied to it directly.

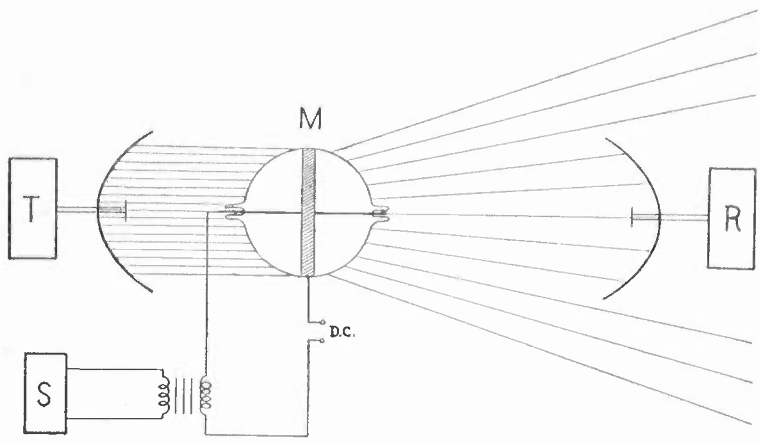

Fig. 1

An estimate of the amount of frequency modulation which accompanies the amplitude modulation can be obtained from some of the known absorption and refraction constants of an ionized gas.

If t represents the phase time required for the electromagnetic wave to pass through the ionized gas, l is the length of path in the ionized gas, n the index of refraction (assumed constant through the path), and c the velocity of light, then $t = ln/c$. If the subscript 0 refers to conditions when only the direct-current potential is applied to the gas tube,

$$t_0 = \frac{ln_0}{c} \quad \text{and} \quad t - t_0 = \frac{l}{c}(n - n_0).$$

Therefore the phase shift due to the change in index of refraction from n to n_0 equals

$$\frac{2\pi f_0 l}{c}(n - n_0) = \frac{2\pi l}{\lambda}(n - n_0)$$

where f_0 is the frequency of the unmodulated wave, and λ is the wavelength in air. Under practical conditions for a 10-centimeter wave, almost 100 per cent modulation can be obtained using a path length of 20 centimeters in the gas, with an accompanying change in index of refraction of 0.3 above and below that in the gas ionized with the direct current.[1] This leads to a phase shift of approximately π for maximum modulation. Since the frequency shift $\Delta f =$ phase shift $\times$ the modulation frequency, this represents a frequency modulation equal to $\pi \times$ the modulating frequency. This frequency modulation is not too severe for a great many purposes. This analysis assumes that all the amplitude modulation is due to absorption of the beam and that dispersion, refraction, and reflection are not playing an important part. If, as may be the case, these effects increase the amplitude modulation, the frequency modulation will be further reduced since a smaller change of index of refraction, or ion density, will be required in order to obtain a large per cent of amplitude modulation. In the tubes which have been used, the larger part of the operation of the modulator appears to be due to absorption, although evidence is at hand to show that reflection, refraction, and scattering also play a part.

[1] Keck and Zenneck, *Hochfrequenz. und Electroakustik*, vol. 40, p. 153, (1932).

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AUDIO-FREQUENCY MEASUREMENT BY THE ELECTRICALLY-EXCITED MONOCHORD*

By
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Summary—*A flexible stretched string is resonant at an infinite number of frequencies, which are successive multiples of the lowest or fundamental frequency. It has been suggested that an electrically-excited stretched steel wire might therefore be used to measure audio frequencies by giving multiples and submultiples of the frequency of a tuning fork. In practice, however, the resonant frequencies are far from being exact multiples of the fundamental. In this paper it is shown that this is due to the imperfect flexibility of the steel wire. A simple expression is given for the resonant frequencies, together with experimental evidence of its accuracy. The expression involves the elastic modulus of the steel of the wire, and a convenient method of determining this quantity is described.*

INTRODUCTION

THE electrically-excited monochord[1] furnishes a most convenient means of measuring audio frequencies, by giving multiple and submultiple frequencies of any standard tuning fork. The apparatus is inexpensive, and both simple and quick in use. It consists of a single stretched steel wire, mounted on a sounding board, and excited by a magnet system similar to that of a moving-iron loud speaker. It is well known that the free vibrations of a uniform and flexible stretched string of length l are given by

$$\begin{aligned} y = {} & A_1 \sin \frac{\pi x}{l} \sin (\omega t + B_1) + A_2 \sin \frac{2\pi x}{l} \sin (2\omega t + B_2) \\ & + A_3 \sin \frac{3\pi x}{l} \sin (3\omega t + B_3) + \cdots \end{aligned} \tag{1}$$

where y is the displacement at time t of a point distant x from one end of the string; the constants A and B depending on the nature of the initial disturbance. It will thus be seen that the string is resonant at frequencies of $\omega/2\pi$, $2\omega/2\pi$, $3\omega/2\pi$, etc. If alternating current flows through the exciting magnet coils, and if the frequency of the alternating current be varied continuously through the audio-frequency range, the string will be thrown into violent vibration as the frequency of the alternating current passes through the values $\omega/2\pi$, $2\omega/2\pi$, $3\omega/2\pi$, etc.,

* Decimal classification: R210. Original manuscript received by the Institute, September 14, 1933.

[1] J. H. O. Harries, PROC. I.R.E., vol. 17, p. 1316; August, (1929).

and if the initial tension of the string has been adjusted to make the fundamental resonant frequency, $\omega/2\pi$, equal to that of a standard tuning fork, then the second, third, fourth, etc., harmonic resonant frequencies (i.e. $2\omega/2\pi$, $3\omega/2\pi$, $4\omega/2\pi$) are known, and the frequency calibration of the source of alternating current is established. Submultiples of the frequency of a standard tuning fork may, of course, be obtained by adjusting the tension of the monochord string so that, say, its fifth resonant frequency, $5\omega/2\pi$, is equal to that of the fork. Its lower resonant frequencies then give frequencies of 1/5, 2/5, 3/5, and 4/5, respectively, of that of the fork. As has been pointed out by Harries[1] the labor of counting up to, say, the thirtieth resonance, may be avoided by counting the nodes in the string. When the string is vibrating in its fundamental mode only (as when it is excited by a sinusoidal current of frequency $\omega/2\pi$) its form at any one instant is given by

$$y = K \sin \frac{\pi x}{l} \cdot$$

Similarly, when vibrating in its nth mode its form at any one instant is given by

$$y = K \sin \frac{n\pi x}{l}$$

showing that there are $(n-1)$ nodes (i.e. points which remain at rest) in the string. On touching these points with the finger the vibration of the string is unaffected, whereas if the string be touched at any other point its vibration is checked. Thus if the finger (or, better still, a pencil) be drawn along the string, and the points counted at which the sound reappears, the number, or "order," of the mode of vibration is at once determined.

Deviations Due to Rigidity

Unfortunately, however, when the monochord is used in practice it is found that the frequencies of the modes of vibration deviate considerably from the harmonic scale of the above theory. In an extreme case the frequency of the fourth mode might be 10 per cent in excess of the calculated frequency, and that of the twentieth mode as much as 100 per cent in excess. This deviation can be observed in two ways. First, on exciting, say, the fifth and tenth modes in succession (by supplying the exciting magnet coils with alternating current of suitable frequencies) and comparing the pitch of two notes produced, it will be observed that the interval between the two notes is greater than an octave; i.e. that the frequency of the tenth mode is more than twice that of the fifth. A trained ear can estimate the interval between the notes, and thus the approximate ratio of the frequencies may be found. Second, if a second harmonic is present in the periodic force exerted by the magnets on the wire (as is always the case to some extent) and if the frequency, $\omega/2\pi$, of the applied alternating current be such as to excite the first mode of vibration of the string, then there will also be a periodic pull on the wire of frequency $2\omega/2\pi$ due to the second harmonic. Now if the frequency of the second mode of vibration were double that of the fundamental mode (as in the above simple theory) this double frequency pull should excite the second mode simultaneously with the first or fundamental mode. It is found, however, that the second mode is not excited until the frequency of the alternating current is increased slightly (say to $(\omega+\delta\omega)/2\pi$) indicating that the frequency of the second mode is greater than twice that of the fundamental, being in fact $2(\omega+\delta\omega)/2\pi$ instead of $2\omega/2\pi$.

If a source of alternating current of variable frequency (e.g. a valve oscillator) be connected to the exciting magnet coils of the monochord, and the frequency be continuously increased through the audio-frequency range, a succession of notes will be heard of frequencies equal to the frequencies of the successive modes of vibration of the string. The effect of a weak second harmonic in the periodic pull of the magnet is to introduce a second series of notes, much fainter than the first; e.g. when the frequency of the alternating current reaches $(\omega+\delta\omega)/2\pi$ a faint note of frequency $2(\omega+\delta\omega)/2\pi$ will be heard, due to the exciting of the second mode of vibration by the weak second harmonic. By noting the displacement of these 'phantom resonances' from the corresponding loud notes (in terms of condenser-dial divisions if a tuned valve-oscillator is used as source of alternating current) the frequencies of the modes of vibration may be estimated. Using this method in conjunction with the aural comparison of pitch already mentioned, it has been found that

(1) the deviation from the harmonic scale of the above simple theory increases rapidly with the order of the mode of vibration,

(2) the deviation decreases as the tension of the wire increases, and

(3) the deviation increases as the diameter of the wire increases. All these facts point to lack of flexibility of the wire as the cause of the departure from the theory. The following modified theory, taking into account the rigidity of the wire, has been shown (in a manner to be described) sufficient to define the behavior of the string, and an easily applied correction has been formulated.

Correction for Rigidity

The method of the correction is that of Lord Rayleigh,[2] who shows that for the purpose of the correction we may assume the vibration to be of the same *type* as for a flexible string; viz,

$$y = \phi_1 \sin\frac{\pi x}{l} + \phi_2 \sin\frac{2\pi x}{l} + \phi_3 \sin\frac{3\pi x}{l} + \cdots = \sum_1^\infty \phi_n \sin\frac{n\pi x}{l},$$

where ϕ_n is a function of the time t. Let ρ be the mass per unit length of the string, and T its initial tension. The kinetic energy of the string at any instant is given by[3]

$$\text{K.E.} = \frac{1}{2}\int_0^l \rho\left(\frac{dy}{dx}\right)^2 dx.$$

and since we may assume the vibrations to conform to the above type, we have, after substituting and integrating,

$$\text{K.E.} = \sum_1^\infty \frac{1}{4}\rho l \phi_n{}^2. \tag{2}$$

If the string were perfectly flexible the potential energy at any instant would be given by[3]

$$\begin{aligned}\text{P.E.} &= \frac{1}{2}T\int_0^l \left(\frac{dy}{dx}\right)^2 dx \\ &= \sum_1^\infty \frac{1}{4}Tl\cdot\frac{n^2\pi^2}{l^2}\cdot\phi_n{}^2.\end{aligned} \tag{3}$$

Adding (2) and (3) and solving, we should have evaluated ϕ_n for the case of a perfectly flexible string, viz., $\phi_n = A_n \sin(n\omega t + B_n)$ as in (1). Note that

$$\frac{\omega}{2\pi} = \frac{1}{2l}\sqrt{\frac{T}{\rho}}.$$

In order to evaluate ϕ_n for the case of a string or wire which is not perfectly flexible, a term representing the potential energy due to bending must be added to (3)

$$\text{P.E. due to bending} = \frac{1}{2}\int_0^l \frac{M^2}{EI}dx$$

[2] Lord Rayleigh, "Theory of Sound," vol. 1. par. 137.
[3] Lord Rayleigh, "Theory of Sound," vol. 1, par. 122.

where M is the bending moment at a point distant x from one end of the wire, I the second moment of area of the section at that point about an axis through its centroid, and E Young's modulus for the material of the wire. If R be the radius of curvature at a point distant x from one end of the wire,

$$\text{P.E. due to bending} = \frac{1}{2}\int_0^l \frac{EI}{R^2}\,dx.$$

Since dy/dx is always very small for a vibrating string or wire we may write $R = 1/(d^2y/dx^2)$. Thus, for a uniform wire,

$$\text{P.E. due to bending} = \frac{1}{2}\int_0^l EI\left(\frac{d^2y}{dx^2}\right)^2 dx = \sum_1^\infty \frac{1}{4}EI\cdot\phi_n{}^2 l\cdot\frac{n^4\pi^4}{l^4}\,.$$

$$\text{Total P.E.} = \sum_1^\infty\left(\frac{1}{4}Tl\cdot\frac{n^2\pi^2}{l^2}\cdot\phi_n{}^2 + \frac{1}{4}EI\cdot\phi_n{}^2 l\,\frac{n^4\pi^4}{l^4}\right)$$

$$= \sum_1^\infty \frac{1}{4}Tl\cdot\frac{n^2\pi^2}{l^2}\cdot\phi_n{}^2\left(1 + \frac{EI}{T}\,\frac{n^2\pi^2}{l^2}\right).$$

Or, writing C for the expression $EI\pi^2/Tl^2$,

$$\text{total P.E.} = \sum_1^\infty \frac{1}{4}Tl\cdot\frac{n^2\pi^2}{l^2}\,\phi_n{}^2(1 + n^2C). \tag{4}$$

Comparing this with (3) we see that the effect of rigidity is to introduce the factor $(1+n^2C)$. Adding (4) and (2) and solving we have

$$\phi_n = A_n \sin\,(n\omega t\sqrt{1 + n^2C} + B_n)$$

where ω has the same value as in the flexible case. Thus instead of being resonant at frequencies of $\omega/2\pi$, $2\omega/2\pi$, $3\omega/2\pi$, $4\omega/2\pi$, etc., the string is resonant at frequencies of

$$\frac{\omega}{2\pi}\sqrt{1 + C},\quad \frac{2\omega}{2\pi}\sqrt{1 + 4C},\quad \frac{3\omega}{2\pi}\sqrt{1 + 9C},\quad \frac{4\omega}{2\pi}\sqrt{1 + 16C},\ \text{etc.},$$

$$\text{where } C = \frac{EI}{T}\cdot\frac{\pi^2}{l^2}\,.$$

In short, the resonant frequencies are given by

$$f = nf_0\sqrt{1 + n^2C}. \tag{5}$$

where f represents the frequency of the *nth* mode, and

$$f_0 = \frac{\omega}{2\pi} = \frac{1}{2l}\sqrt{\frac{T}{\rho}}\,.$$

With sufficient accuracy for our purposes f_0 may be taken as the fundamental frequency of the monochord.

Experimental Verification

It has been ascertained experimentally that for values of n up to 30, equation (5) is followed with an accuracy which is within the limits

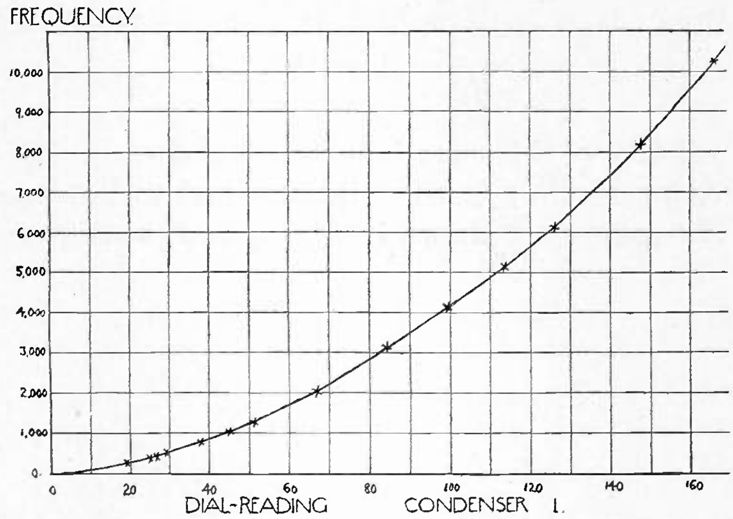


Fig. 1—Frequency calibration of oscillator.

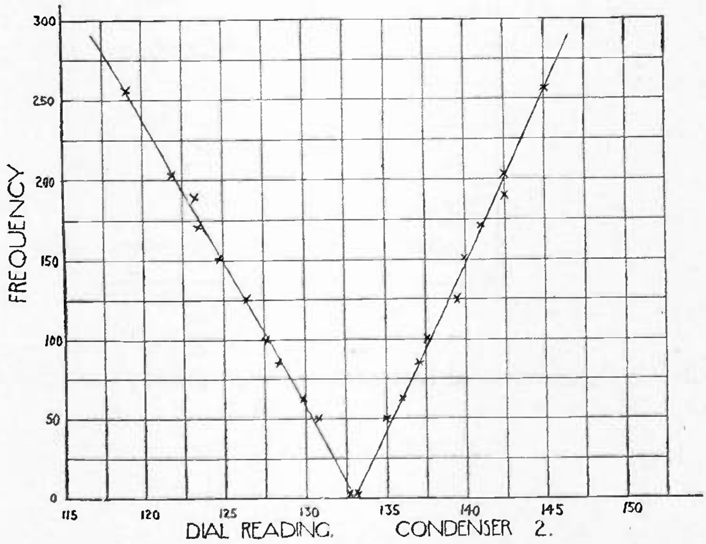


Fig. 2—Frequency calibration of oscillator.

of experimental error, and may be relied upon to within one per cent. Using a frequency-calibrated heterodyne oscillator, the frequencies of the first thirty monochord resonances were determined, for each of a series of values of initial tension. The calibration of the oscillator was effected by comparison with a large number of standard tuning forks, and the calibration curves are shown in Figs. 1 and 2. The exciting magnet coils of the monochord were supplied with alternating current from the oscillator, and the oscillator frequency was adjusted to resonance with each of the modes of vibration of the monochord in turn. The order, n, of the mode of vibration was determined by counting the nodes in the wire and its frequency, f, established by observing the dial reading of the oscillator. Equation (5) may be written

$$\frac{f^2}{n^2 f_0{}^2} - 1 = n^2 C.$$

To test the validity of this equation, the expression on the left-hand side is plotted against n^2, f_0 being taken equal to fundamental frequency, f_1. The graphs are shown in Fig. 3, and prove to be straight

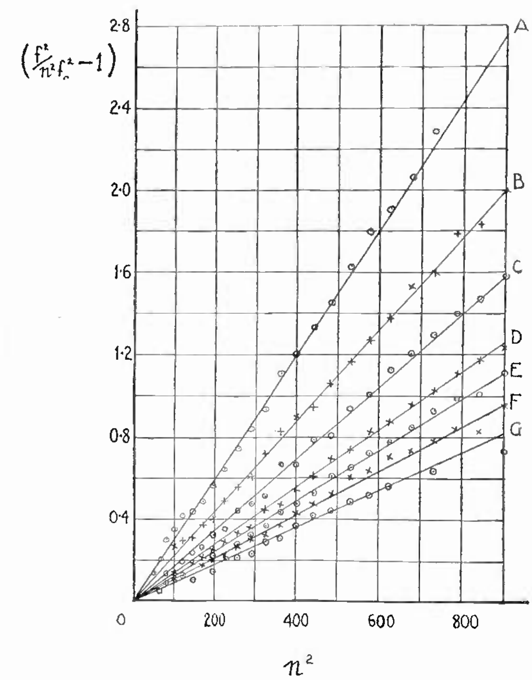


Fig. 3—A	Fundamental	frequency	59.8	cycles	per	second
B	"	"	71.0	"	"	"
C	"	"	79.4	"	"	"
D	"	"	89.3	"	"	"
E	"	"	94.5	"	"	"
F	"	"	101.3	"	"	"
G	"	"	113.6	"	"	"

lines as predicted. The values of C determined from these graphs are given below, and will be seen to agree closely with the values calculated from the equation

$$C = \frac{EI}{T} \cdot \frac{\pi^2}{l^2}.$$

Fundamental frequency	59.8	71.0	79.4	89.3	94.5	101.3	113.6
C from graph	0.00304	0.0022	0.00174	0.0014	0.00124	0.00107	0.00088
C by calculation	0.00302	0.00214	0.00171	0.00135	0.00121	0.00105	0.000838

It should be noted that for values of C about 0.001, an error of 10 per cent in the value of C causes an error of only about 0.05 per cent in the calculated value of the frequency of the tenth mode, and

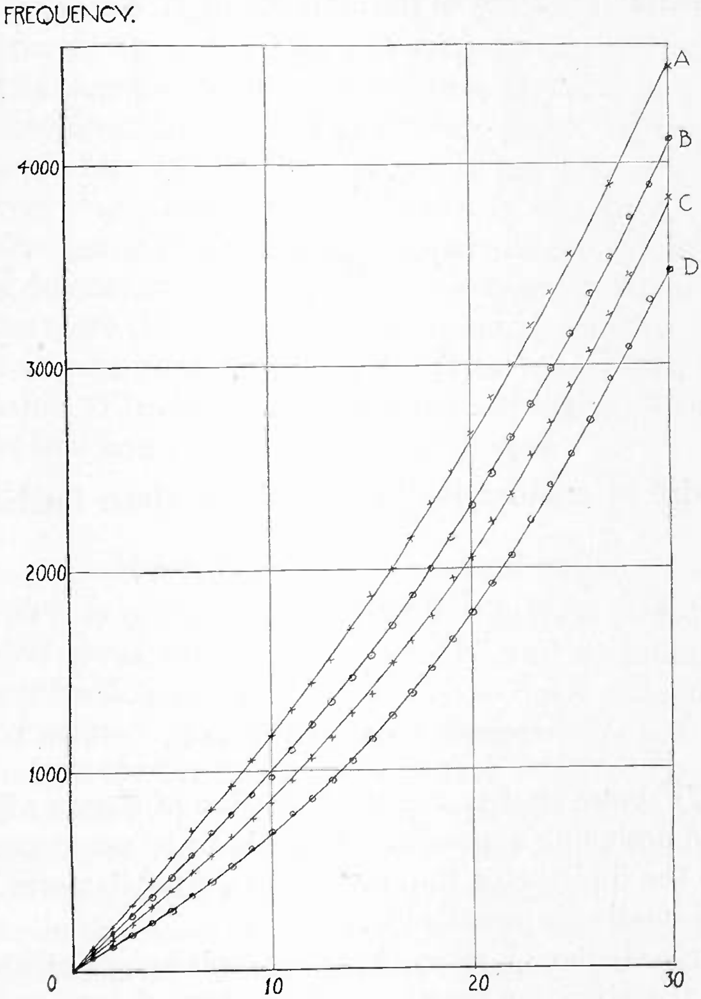


Fig. 4—A Fundamental frequency 113.6 cycles per second.
B " " 94.5 " " "
C " " 79.4 " " "
D " " 59.8 " " "

an error of only 1.7 per cent in that of the twentieth mode. The curves shown in Fig. 4 are the curves of frequency calculated from (5) using the calculated value of C. The points shown on them are the frequencies of the monochord resonances as determined by comparison with the calibrated oscillator.

The above tests were carried out on a monochord using a meter length of 1/16-inch diameter wire. Similar results were obtained with a smaller model, using a foot length of 0.028-inch diameter wire, the value of C being 0.0031 at a fundamental of 256 cycles.

Design and Use

With sufficient accuracy for our purposes we may take f_0 equal to the fundamental frequency of the monochord, f_1. Hence,

$$f_1 = \frac{1}{2l}\sqrt{\frac{T}{\rho}}$$

giving,

$$T = 4l^2 f_1^2 \rho.$$

We have,

$$C = \frac{EI}{T} \cdot \frac{\pi^2}{l^2}.$$

Thus for wire of circular section and for a given fundamental frequency

$$C \text{ varies as } d^2/l^4. \tag{7}$$

Also,

$$\text{tensile stress varies as } l^2. \tag{8}$$

Equation (7) shows that to reduce the value of C to a minimum the procedure in designing a monochord should be to use as great a length as possible, the limit being imposed by (8). The diameter of the wire should be as small as is practicable.

With perfectly flexible wires the behavior is independent of whether the ends of the vibrating length are defined by clamps (in which case the terminal directions are also fixed) or by bridges (terminal directions free); but in the case of wires which are not perfectly flexible this is not so. The above theory correcting for the rigidity of the wire assumes terminal directions free, and the experimental verification has been carried out using a monochord in which the wire is stretched between two piano pins, the vibrating length being defined by ebonite bridges.

With the lengths and diameters used, however, it has been found that, within the limits of experimental error, the behavior is independent of whether clamps or bridges are used. Bridges are generally to be preferred on account of the greater ease of adjusting the tension for a given fundamental frequency. For fine adjustments of tension, a long bar with a square hole at one end, to fit the piano pin, is infinitely superior to the ordinary piano tuning key. In setting the monochord frequency by means of a tuning fork it is better to set the monochord frequency *near* to that of the fork, and to determine the difference by counting beats, than to attempt to set the frequency accurately to that of the fork.

It is important that the base and supports of the monochord shall be as rigid as possible, lest the frequencies of the modes of vibration should be modified by vibration of these parts. In this connection bridges are slightly inferior to clamps. It has been shown[4] that the effect of very high-frequency resonances in the base is slightly to lower the frequency of all modes *in the same ratio*, whereas the effect of very low-frequency resonances is to increase the frequency of some of the modes more than others. The aim should therefore be to increase the stiffness, and a good plan is to clamp the monochord securely to a bench or table. If reasonable precautions are taken all trouble due to vibration of base and supports can be eliminated.

Appendix

Measurement of Young's Modulus

Although it is usually safe to take $E=30\times 10^6$ pounds per square inch for steel wires without more ado, it may sometimes be necessary to measure this quantity. This may be done without further apparatus and without removing the wire from the monochord. The procedure is to tune the fundamental resonance to a given frequency, and to mark off a measured length, l_1, on the wire by two fine scriber marks. The tension is then reduced until the second mode has the given frequency, and the marked length of wire again measured, l_2. If p_1 and p_2 be respectively the stress in the wire originally and that after reducing the tension, and l_0 be the unstretched length of the marked section, then

$$l_1 = l_0(1 + p_1/E)$$

$$l_2 = l_0(1 + p_2/E)$$

giving,

[4] Lord Rayleigh, "Theory of Sound," vol. 1, par. 135.

$$E = \frac{p_1 l_2 - p_2 l_1}{l_1 - l_2} \cdot p_2.$$

Neglecting the correction for rigidity (even in an extreme case the resulting error in the value of E will be less than 1 per cent) we have

$$p_1 = 4p_2$$

$$\therefore \quad E = \frac{4l_2 - l_1}{l_1 - l_2} \cdot p_2$$

$$= \frac{4l_2 - l_1}{l_1 - l_2} \cdot \frac{L^2 f^2 \rho}{A}$$

where L is the whole length of the vibrating wire, A its cross-sectional area, and f the given frequency.

The accuracy of the method is chiefly limited by the accuracy of measuring $(l_1 - l_2)$, but by the use of two traversing microscopes (one at each scriber mark) sufficiently accurate results can be obtained.

Proceedings of the Institute of Radio Engineers
Volume 22, Number 6 *June, 1934*

BOOK REVIEW

The Physics of Electron Tubes, by R. L. Koller. Research Laboratory of the General Electric Company, Schenectady. Published by McGraw-Hill Book Co., Inc., New York. 199 pages. Price $3.00.

This book, one of the International Series in Physics, is based upon the lectures given by Dr. Koller at the Massachusetts Institute of Technology during the summer of 1931 and is concerned with the broad outlines of certain of the more fundamental physical phenomena of electron tubes. The book is not intended to be a compendium, but rather by means of the brief outline and the bibliography, to serve as a guide to study.

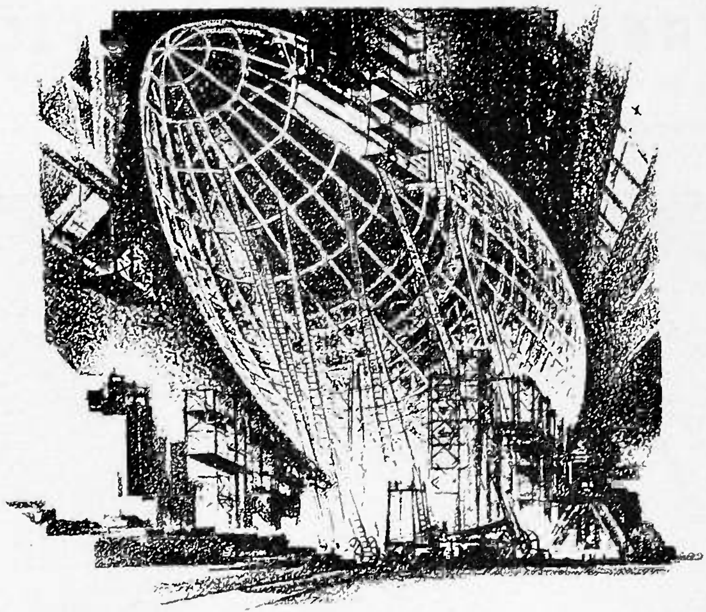
Approximately one hundred pages are concerned with electron emission—its determination and characteristics; most of the rest of the book is concerned with discharge in gases and with photo effects of various kinds.

Since Dr. Koller's book is restricted to certain of the more fundamental physical phenomena occurring in tubes, many subjects important to the research or development engineer, are omitted. This omission is in part compensated for by the rather inclusive lists of references at the ends of the various chapters.

The material is of high grade, well arranged and well illustrated. The book makes a valuable addition to the list of books available in the electronic field.

*B. E. Shackelford

* RCA Radiotron Company, Inc., Harrison, New Jersey.



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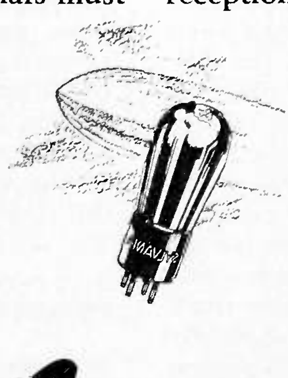

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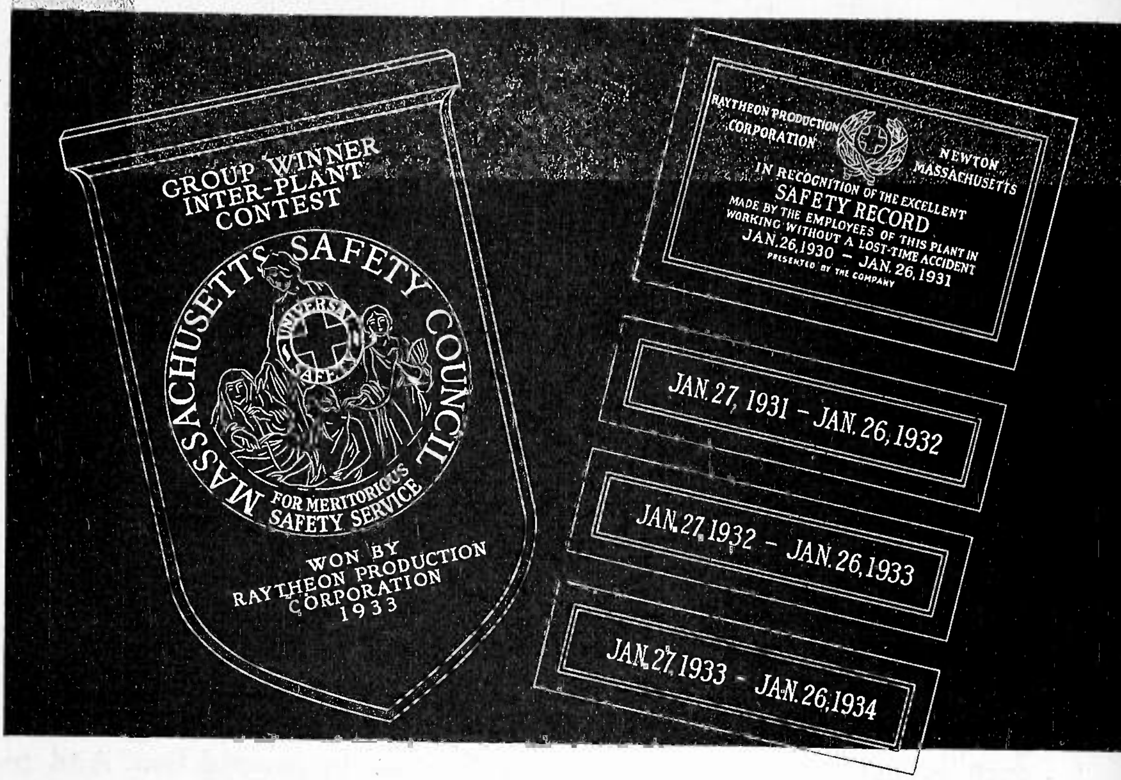

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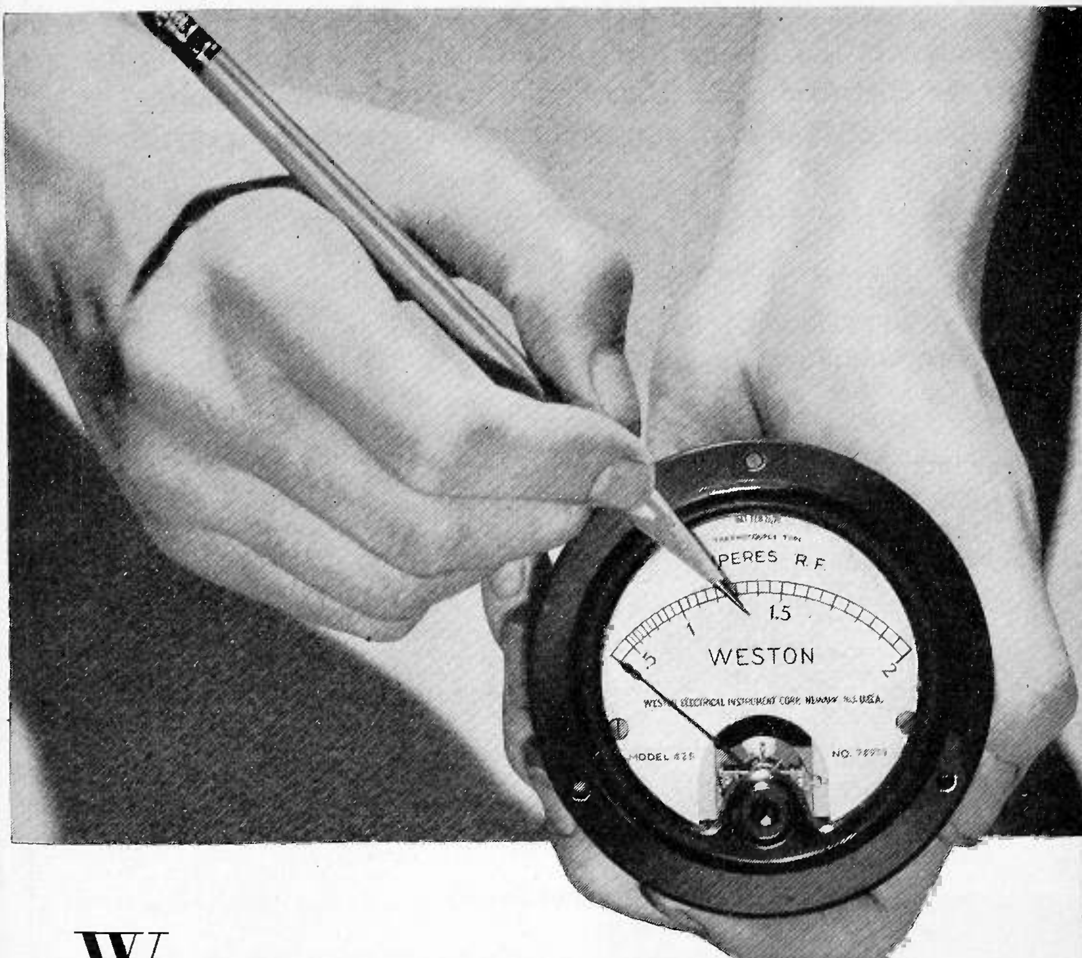

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