Proceedings of the I·R·E

DECEMBER 1940

VOLUME 28 NUMBER 12

Institute of Radio Engineers

New York Meeting—March 5, 1941

Joint Meeting with American Section, International Scientific Radio Union, Washington, D.C., May 2, 1941

Summer Convention—Institute of Radio Engineers—Detroit, Michigan June 23, 24, and 25, 1941

SECTION MEETINGS

ATLANTA
February 21

CLEVELAND
February 28

DETROIT
February 21

LOS ANGELES
March 18

PHILADELPHIA
March 6

PITTSBURGH
March 10

WASHINGTON
March 10

SECTIONS

ATLANTA—Chairman, P. C. Bangs; Secretary, Adolph Andersen, Federal Communications Commission, 411 Federal Annex, Atlanta, Ga.

BALTIMORE—Chairman, Ferdinand Hamburger; Secretary, G. J. Gross, Pennsylvania Water and Power Co., 1611 Lexington Bldg., Baltimore, Md.

BOSTON—Chairman, W. L. Barrow; Secretary, P. K. McElroy, General Radio Co., 30 State St., Cambridge, Mass.

BUENOS AIRES—Chairman, A. T. Cosentino; Secretary, G. J. Andrews, Cia. Standard Electric Argentina, Cangallo 1286, Buenos Aires, Argentina.

BUFFALO-NIAGARA—Chairman, B. E. Atwood; Secretary, Leroy Fiedler, 98 West Ave., Lockport, N. Y.

CHICAGO—Chairman, G. I. Martin; Secretary, R. A. Kay, RCA Institutes, Inc., 1154 Merchandise Mart, Chicago, Ill.

CINCINNATI—Chairman, J. M. McDonald; Secretary, W. L. Schwesinger, Crosley Corp., Cincinnati, Ohio.

CLEVELAND—Chairman, C. E. Smith; Secretary, W. G. Hutton, 7314 Dorothy Ave., Parma, Ohio.

CONNECTICUT VALLEY—Chairman, K. A. McLeod; Secretary, W. M. Smith, 250 Main St., Suffield, Conn.

DALLAS-FORT WORTH—Chairman, D. A. Peterson; Secretary, J. R. Sullivan, WRR, Dallas, Tex.

DETROIT—Chairman, M. Cottrell; Secretary, Paul Frincke, WJBK, 6559 Hamilton Ave., Detroit, Mich.

EMPORIUM—Chairman, R. K. Gessford; Secretary, C. W. Reash, Hygrade Sylvania Corp., Emporium, Pa.

INDIANAPOLIS—Chairman, Arthur Curtis; Secretary, B. V. K. French, P. R. Mallory & Co., E. Washington St., Indianapolis, Ind.

LOS ANGELES—Chairman, W. W. Lindsay, Jr.; Secretary, C. F. Wolcott, 1609 S. Western Ave., Los Angeles, Calif.

MONTREAL—Chairman, E. A. Laport; Secretary, W. A. Nichols, Canadian Broadcasting Corp., 1012 Keefer Bldg., Montreal, Que.

NEW ORLEANS—Chairman, G. H. Peirce; Secretary, D. W. Bowman, 8327 Sycamore St., New Orleans, La.

PHILADELPHIA—Chairman, C. M. Burrill; Secretary, R. L. Snyder, 103 Franklin Rd., Glassboro, N. J.

PITTSBURGH—Chairman, R. E. Stark; Secretary, M. G. Jarrett, Bell Telephone Company of Pennsylvania, Pittsburgh, Pa.

PORTLAND—Chairman, Marcus O'Day; Secretary, Earl Schoenfeld, Route 1, Box 3070, Portland, Ore.

ROCHESTER—Chairman, H. C. Sheve; Secretary, O. L. Angevine, Jr., Stromberg-Carlson Telephone Manufacturing Co., Rochester, N. Y.

SAN FRANCISCO—Chairman, L. J. Black; Secretary, Karl Spangenberg, Stanford University, Calif.

SEATTLE—Chairman, K. H. Ellerbeck; Secretary, R. J. Gleason, Pacific Alaska Airways, Inc., 3314 White Bldg., Seattle, Wash.

TORONTO—Chairman, H. S. Dawson; Secretary, L. C. Simmonds, 301 King St., E., Toronto, Ont.

WASHINGTON—Chairman, M. H. Biser; Secretary, C. M. Hunt, WJSV, 817 Earle Bldg., Washington, D. C.

Proceedings of the I·R·E

Published Monthly by
The Institute of Radio Engineers, Inc.

VOLUME 28 — *December, 1940* — NUMBER 12

Entered as second-class matter October 26, 1927, at the post office at Menasha, Wisconsin, under the Act of February 28, 1925, embodied in Paragraph 4, Section 538 of the Postal Laws and Regulations. Publication office, 450 Ahnaip Street, Menasha, Wisconsin. Subscription, $10.00 per year; foreign, $11.00.

THE INSTITUTE

The Institute of Radio Engineers serves those interested in radio and allied electrical-communication fields through the presentation and publication of technical material. In 1913 the first issue of the PROCEEDINGS appeared; it has been published uninterruptedly since then. Almost 2000 technical contributions have been included in its pages and portray a currently written history of developments in both theory and practice.

STANDARDS

In addition to the publication of submitted papers, many thousands of man-hours have been devoted to the preparation of standards useful to engineers. These comprise the general fields of terminology, graphical and literal symbols, and methods of testing and rating apparatus. Members received a copy of each report. A list of the current issues of these reports follows:

Standards on Electroacoustics, 1938
Standards on Electronics, 1938
Standards on Radio Receivers, 1938
Standards on Radio Transmitters and Antennas, 1938.

MEETINGS

Meetings at which technical papers are presented are held in the twenty-four cities in the United States, Canada, and Argentina listed on the inside front cover of this issue. A number of special meetings are held annually and include one in Washington, D. C., in co-operation with the American Section of the International Scientific Radio Union (U.R.S.I.) in April, which is devoted to the general problems of wave propagation and measurement technique, the Summer Convention which is held late in June in a place other than New York City, the Rochester Fall Meeting in co-operation with the Radio Manufacturers Association in November, which is devoted chiefly to the problems of broadcast-receiver design, the Pacific Coast Convention, which is held each year during the summer months and is supervised by a joint committee of our four Pacific Coast sections, and the Annual Convention, which will be held in New York City during January.

MEMBERSHIP

Membership has grown from a few dozen in 1912 to more than five thousand. Practically every country in the world in which radio engineers may be found is represented in our membership roster. Approximately a quarter of the membership is located outside of the United States. There are several grades of membership, depending on the qualifications of the applicant. Dues range between $3.00 per year for Students and $10.00 per year for Members. PROCEEDINGS are sent to each member without further payment.

PROCEEDINGS

The contents of each paper published in the PROCEEDINGS are the responsibility of the author and are not binding on the Institute or its members. Material appearing in the PROCEEDINGS may be reprinted or abstracted in other publications on the express condition that specific reference shall be made to its original appearance in the PROCEEDINGS. Illustrations of any variety may not be reproduced, however, without specific permission from the Institute.

Papers submitted to the Institute for publication shall be regarded as no longer confidential. They will be examined by the Papers Committee and Board of Editors to determine their suitability for publication. Suggestions on the mechanical form in which manuscripts should be prepared may be obtained from the Secretary.

SUBSCRIPTIONS

Annual subscription rates for the United States of America, its possessions, and Canada, $10.00; to college and public libraries when ordering direct, $5.00. Other countries, $1.00 additional.

The Institute of Radio Engineers, Inc.
Harold P. Westman, Secretary
330 West 42nd Street
New York, N.Y.

Two-Signal Cross Modulation in a Frequency-Modulation Receiver*

HAROLD A. WHEELER†, FELLOW, I.R.E.

Summary—*Cross modulation of a desired signal by an undesired signal is caused by amplitude modulation on the undesired signal and appears as amplitude modulation of the desired signal. Therefore, a frequency-modulation receiver is insensitive to cross modulation, except to the extent that the undesired signal has amplitude modulation and the receiver is incidentally sensitive to amplitude modulation. It is shown that cross modulation is not caused by frequency modulation on the undesired signal and does not appear as frequency modulation of the desired signal. There still may be a kind of beat-note interference if the signals are in adjacent frequency channels. Also the desired signal may be attenuated if the undesired signal is strong enough to overload the receiver before it is filtered out.*

FREQUENCY-modulation receivers have a different behavior with respect to the cross modulation which has been experienced in amplitude-modulation receivers. The amount of residual cross modulation is determined by the incidental amplitude modulation rather than the essential frequency modulation. This distinction can be proved in simple terms, and may influence the usual compromise between gain and selectivity in the coupling from the antenna to the first tube in a frequency-modulation receiver.

Cross modulation of one signal by another usually occurs in the first tube of an amplitude-modulation receiver if there is too little selectivity or too great a voltage ratio from the antenna to the first grid. It is caused by a strong undesired signal modulating the transconductance in the first tube and thereby modulating the amplitude of all other signals passing through this tube. The undesired signal may be filtered out in subsequent selective circuits, but its modulation rides through as cross talk on the carrier of the desired signal.[1]

Amplitude cross modulation was experienced the most in the receivers around 1928 which had the "antenna-coupling tube," the first tube of the receiver,

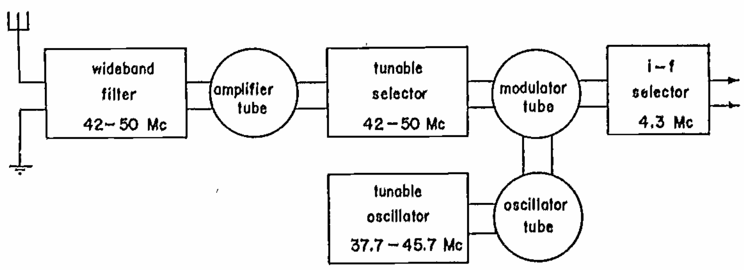


Fig. 1—The introductory circuits of a frequency-modulation receiver.

receiving the antenna voltage of all signals directly on its grid without any preceding selection. Sometimes the modulation of several strong undesired signals would ride through on the carrier of the desired signal. This could not be cured by subsequent selectivity, even though the desired signal might be as strong as the others and separated therefrom by a large difference of frequency.

In order to investigate any corresponding effect in a frequency-modulation receiver, a set is outlined which has no selectivity ahead of the first tube, at least none to separate signals in the frequency-modulation band (42 to 50 megacycles). Fig. 1 shows the outline of the first part of such a set. The antenna is coupled to the first tube, a radio-frequency amplifier, by a fixed

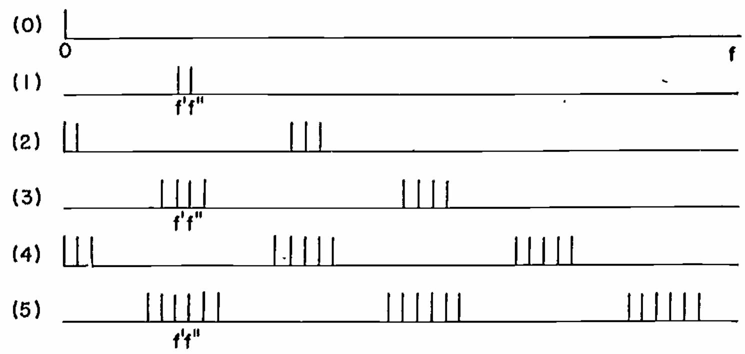


Fig. 2—Frequency terms caused by successive derivatives, with two signals.

wide-band filter which passes all signals in the frequency-modulation band. This tube is coupled to the frequency changer by a tunable selector which favors the desired signal against others and especially assists the wide-band filter in providing the necessary attenuation at the image frequency, in this case 8.6 megacycles below the signal frequency.

The amplitude cross modulation is caused by certain types of curvature in the transconductance of the first tube, known as the third derivative and higher odd-order derivatives. These are the same types of curvature which are associated with overloading and limiting action. They cannot be canceled out in a push-pull amplifier, as can the second and higher even-order derivatives. Of course the transconductance itself is the first derivative or slope of output current with input voltage. To minimize these effects was the purpose of the gradual-cutoff or variable-mu tubes, but the offending curvatures cannot be reduced much further in a practical tube without excessive reduction of its amplifying ability.[2]

The interference effects caused by variation of the transconductance are most readily classified in terms of the derivatives of output current with respect to input voltage. Fig. 2 shows such a classification of the frequency components resulting from two signals of

* Decimal classification: R161.5×R414. Original manuscript received by the Institute, October 31, 1940. Presented, New York meeting, February 5, 1941.

† Hazeltine Service Corporation, Little Neck, L. I., N. Y.

[1] This is one cause of the cross-talk interference for which a two-signal test is specified in the I.R.E. Standards on Radio Receivers, 1938, section IV-E-18 p. 48, "Two-signal cross-talk interference."

[2] S. Ballantine and H. A. Snow, "Reduction of distortion and cross-talk in radio receivers by means of variable-mu tetrodes," Proc. I.R.E., vol. 18, pp. 2102–2127; December, 1930.

slightly different frequencies applied to the first tube. Each row gives the components caused by one derivative. The zero-order derivative gives the zero-frequency or direct-current component. The first derivative or transconductance gives the output components corresponding to the input signals, at frequencies f' and f''. The second derivative gives the second harmonics, the rectified direct-current component, and the sum and difference frequencies. The higher derivatives give a greater number of spurious components.

Most of the spurious components have frequencies far removed from the signal frequencies, so they are filtered out by subsequent selectors. The odd-order derivatives, however, give components at the signal

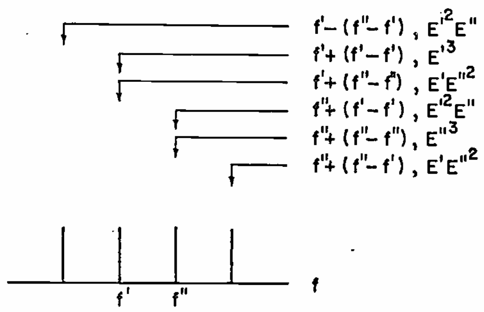


Fig. 3—The near-by frequency components caused by the third derivative.

frequencies and near-by frequencies. Some of these components are not rejected by the selectors, so they deserve further study. Since the first derivative gives only the signal-frequency components, the third derivative is the lowest to give spurious components of near-by frequencies.

Fig. 3 shows the spurious components given by the third derivative, at or near the signal frequencies f' and f''. Each component is labeled with the triple combination of frequencies which determines its frequency. Each is labeled also with the voltage factors which determine its amplitude, in powers of the respective signal voltages E' and E''. For example, there are two spurious terms at the signal frequency f'. One is proportional to E'^3 and represents overloading on this signal alone. The other is proportional to $E'E''^2$ and represents the cross modulation of the amplitude E' of this signal by the amplitude E'' of the other signal. These spurious components cannot be rejected by subsequent selectors, if the signal f' is to be received, because they have the same frequency.

The purpose of Figs. 2 and 3 is to show how all spurious components at either of the signal frequencies have exactly the same frequency so they do not cause interference in the form of frequency modulation. Also they have the same phase. Cross modulation appears only in the form of spurious amplitude modulation of one signal in response to amplitude modulation of the other.

In some cases, however, the modulation of the signal frequencies causes near-by components to swing into the pass band of a selector tuned to either signal. The effect is an interference similar to that caused by an undesired signal in the same channel, except that it can occur only during modulation of both signals.

Fig. 4 shows how this kind of interference can occur between two frequency-modulated signals on adjacent channels. The diagram is based on frequency modula-

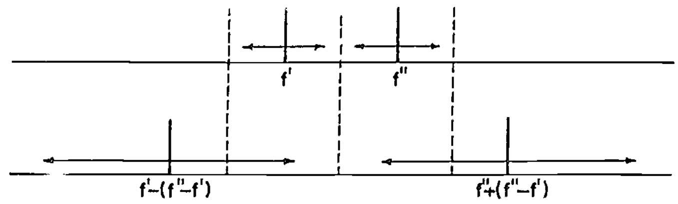


Fig. 4—Noise caused by triple cross products of frequency-modulated signals in adjacent channels.

tion over three fourths of the channel width; for example, ± 75-kilocycle modulation in a channel 200 kilocycles wide. The arrows show the maximum frequency modulation. The dotted lines show the channel width, which is approximately the band width of the subsequent selectors in the receiver. The triple products on near-by frequencies, shown in the lower row, have frequency modulation from both signals. Their maximum modulation is three times that of each signal alone, and occurs only when the two signal frequencies are simultaneously modulated by the maximum amount in opposite directions. When the desired signal is unmodulated, so the reception is most sensitive to noise, this type of interference does not occur. Therefore it is of secondary importance.

If the two signals are separated by one or more intervening channels, Fig. 5 shows how this type of interference cannot occur. As long as each signal frequency is modulated only within its own channel width, the near-by spurious components cannot swing into either signal channel.

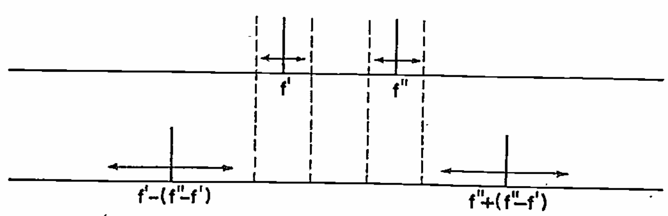


Fig. 5—Freedom from noise with frequency-modulated signals in nonadjacent channels.

This type of adjacent-channel interference is likely to be serious only if the desired signal is very much weaker than the undesired signal. In such a case, it is unlikely that the desired signal could be selected free of interference, even if there were no distortion in the first tube. Also this question has little effect on the design, because even a simple tunable selector between the antenna and the first tube could exercise little selection between signals on adjacent channels.

There is another possible source of similar interference if the derivatives of a very high even order are appreciable. Referring to Fig. 2, the group of spurious components around zero frequency, and that around twice the signal frequencies, might widen out to overlap in the region of the signal frequencies. The order of derivative required to give such interference is at least $2f'/(f''-f')$. For example, in the frequency-modulation service on 42 to 50 megacycles, this kind of interference could be caused only by derivatives of the twelfth and higher orders. It is unlikely that the effects of such high-order derivatives would be appreciable, so this source of interference probably deserves little attention.

As a result of this study, it appears that distortion in the first tube does not cause interference between two frequency-modulated signals on nonadjacent channels. If the undesired signal has incidental amplitude modulation, and to the extent that the receiver is responsive to amplitude modulation, there might be a small amount of interference by the usual amplitude cross modulation. This residual interference can be minimized to a negligible amount by designing for uniform response from the antenna to the first grid, and by including in the receiver a carrier limiter or a balanced frequency detector or both.

Although frequency cross modulation of one signal by another is not caused by distortion in the first tube, the amplitude of the desired signal is affected by that of the undesired signal. A moderate change of amplitude is unimportant, so it is necessary to investigate only the cases in which there may be a great effect on the desired-signal amplitude. All cases are divided in two classes by whether the amplitude of the desired signal is increased or decreased.

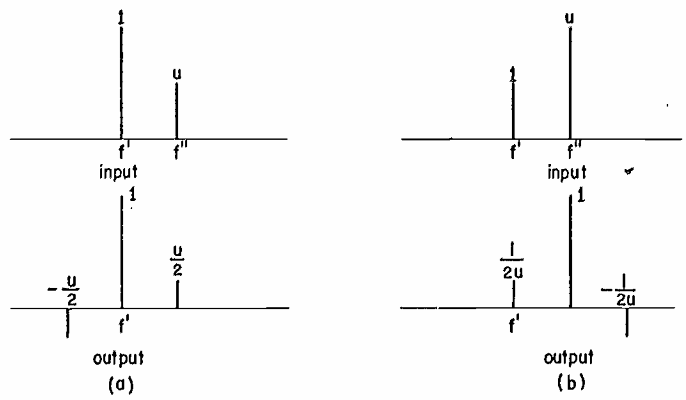


Fig. 6—The masking of a desired signal f' by an undesired signal f'' in a perfect limiter.
(a) Undesired signal much weaker ($u \ll 1$).
(b) Undesired signal much stronger ($u \gg 1$).

The undesired signal is not likely to cause more than a slight increase in the desired-signal amplitude, unless the first tube is biased near cutoff. In this case, the only detrimental effect would be the greater susceptibility to amplitude cross modulation, which is of secondary importance.

There may be a decrease in the desired-signal amplitude, caused by the undesired signal saturating the limiter and thereby masking the desired signal. This reduction of the desired-signal output can be derived readily on the assumption of a limiter which holds uniform the envelope of the two signals together. A simple description of this effect for two signals of much different amplitude is shown in Fig. 6, and the general relation in Fig. 7.

One extreme case, shown in Fig. 6(a), is that in which the desired signal f' is much stronger than the undesired f''. The relative amplitude of the undesired signal is denoted by u, in this case much less than one. In the composite signal input, the undesired weaker signal appears as a single sideband of the desired stronger signal. It is well known that such a single sideband represents composite amplitude and phase modulation in equal amounts. The limiter removes the amplitude modulation and leaves only the phase modulation. The output contains the stronger signal and a skew-symmetrical pair of sidebands representa-

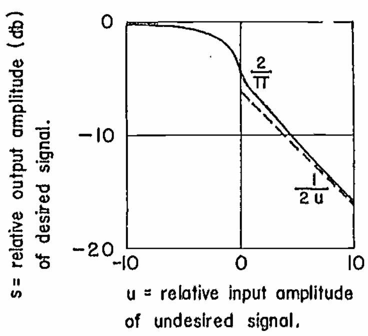


Fig. 7—The masking of a desired signal by an undesired signal in a perfect limiter.

tive of the residual phase modulation.[3] The amplitude of the desired signal, the center component of the output, is determined only by the limiter level, taken as unit amplitude in the diagram, unaffected by the minor components incidental to the phase modulation. Since the phase modulation in the output is the same as that in the input, the total relative amplitude of the sidebands, u in this case, is maintained. Therefore each sideband in the output has the relative amplitude $u/2$, as shown in Fig. 6(a). Since only the desired signal is selected, the presence of the weak undesired signal has no effect on the desired-signal output of the limiter.

The other extreme case, shown in Fig. 6(b), is that in which the undesired signal is much the stronger and therefore saturates the limiter. The relative amplitude u is much greater than one. Following the same line of reasoning, with the two signals interchanged, the desired signal appears in this case as one of the sidebands in the output. Since the principal output component is reduced to unit amplitude by the limiter, each of the minor components is reduced to an amplitude of $\frac{1}{2}u$, much less than the unit amplitude of the limiter output for one signal alone. One of these minor components is the desired-signal output, so the effect of the strong undesired signal is to reduce the desired-signal amplitude by the factor $\frac{1}{2}u$.

The general case, for any relative amplitude of the two signals, is not difficult to solve. The limiter function is performed by reducing to unit amplitude the envelope of the composite output. By trigonometric

[3] Edwin H. Armstrong, "A method of reducing disturbances in radio signaling by a system of frequency modulation," Proc. I.R.E., vol. 24, pp. 689–740, May, 1936. (This effect is described on p. 705.)

transformations, the output is expressed in terms of sine and cosine components of the desired-signal frequency, each amplitude modulated at the beat frequency. The average amplitude of each of these components is evaluated by integration, to obtain the signal-frequency output components. Their quadratic sum is the amplitude of the desired-signal output. It is denoted s and is found to be expressible in terms of elliptic integrals.[4] It is plotted in Fig. 7, and is asymptotic to the extreme cases described in Fig. 6. The reduction of the desired-signal amplitude is not severe unless the undesired signal is much the stronger. Even then it reduces both signal and noise so it has little effect on their ratio.

The background noise caused by overlapping of the adjacent groups of components in Fig. 2 is present in the output of a perfect limiter. It has the greatest amplitude if the two signals are nearly equal, because the limiter function then involves the greatest ratio of maximum over minimum amplitude of the composite signal envelope. With practical limiters, this noise is unlikely to be appreciable.

Qualitative tests have verified the above theoretical predictions with respect to the freedom from cross modulation of frequency and the attenuation of a desired signal in the presence of a strong undesired signal. Each phenomenon was tested with two signals having a frequency separation sufficient to permit selection of the desired signal in the subsequent selectors of the receiver.

The cross-modulation test was made with two frequency-modulated signals coupled to the first grid, a moderately weak desired signal to which the receiver was tuned and an undesired signal as strong as 3 volts. The first tube was a sharp-cutoff pentode of greatest susceptibility to amplitude cross modulation. The receiver had no limiter just preceding the frequency detector. A small amount of cross modulation was detected under some conditions. It was found to be caused by amplitude modulation of the undesired signal, caused by a slope in the filter coupling this signal to the first grid. Either of two expedients was found to remove the cross modulation. First, the filter could be adjusted to remove the slope. Second, the receiver could be carefully tuned to balance out the amplitude modulation in the frequency detector. A limiter would have further reduced the cross talk by removing most of the incidental amplitude modulation.

The masking attenuation test was made with similar conditions except that the undesired signal was unmodulated and even stronger, up to 10 volts, more than sufficient to saturate the first tube as a limiter. The desired signal was attenuated by as much as 10 decibels by the masking effect, but the character of the signal was unchanged and there was only slightly greater noise output relative to the signal.

As an indication of the amount of residual cross talk from the incidental amplitude cross modulation, a representative case may be described. The antenna and wide-band filter coupling all signals to the first grid may have a maximum slope of 1 decibel per megacycle, or about 0.1 decibel for a maximum deviation of 75 kilocycles. This means that the incidental amplitude modulation of the undesired signal would not exceed 1 per cent. The frequency detector may be balanced within 10 per cent by reasonably careful tuning, or the same insensitivity to amplitude modulation may be obtained by a good limiter just ahead of the detector. In this case, the relative output from residual cross modulation in the frequency-modulation receiver is only 1/1000 as great as it would be in an amplitude-modulation receiver under similar conditions of signal strength and sharp cutoff in the first tube.

The principal interference noted in these tests was not the slight cross modulation but rather the selective response at the few critical frequencies at which the superheterodyne receiver happened to be sensitive to interference. Spurious response at such points is reduced by the total amount of selectivity ahead of the frequency changer. This selectivity need not be ahead of the first tube, if it is simply an amplifier.

It is concluded that distortion in the first tube is unlikely to cause interference between two frequency-modulated signals in nonadjacent channels within a frequency band whose width is much less than its minimum frequency. If the first tube acts as a limiter, and if the undesired signal is strong enough to saturate this limiter, the result is a reduction of the desired-signal amplitude, but this effect is small under conditions likely to be met in practice. Therefore, it appears that the coupling from antenna to first grid need not furnish selectivity between signals in the frequency-modulation band of 42 to 50 megacycles, although it should attenuate signals outside of this band. This conclusion remains to be verified in practice for more than two signals, that is, for more than one strong undesired signal.

[4] $$s = \frac{2}{\pi}\int_0^{\pi/2} \frac{1 + u\cos 2x}{\sqrt{1 + u^2 + 2u\cos 2x}}\,dx = \frac{2}{\pi}\,\frac{E(k) + K(k)\cos a}{1 + \cos a}$$

or $(u \leqq 1)$

$$s = \frac{2}{\pi}E(u)$$

in which K and E are the complete elliptic integrals of first and second kinds, and

$$a = 2\ \mathrm{antitan}\ \sqrt{u}, \quad u = \tan^2\frac{a}{2}, \quad k = \frac{2\sqrt{u}}{1+u} = \sin a.$$

For $u = 0, 1, \infty$:

$$s = 1, \quad \frac{2}{\pi}, \quad \frac{1}{2u}.$$

Jahnke-Emde, "Tables of Functions," 1933; pp. 124–151.
Bierens de Haan, "New Tables of Definite Integrals," 1867/1939; tables 53(1), 57(1)(4)(5), 67(5)(7), 68(22)–(25); the notation F' and E' is used instead of K and E.

A Method of Measuring the Magnetic Properties of Small Samples of Transformer Laminations*

HORATIO W. LAMSON†, FELLOW, I.R.E.

Summary—*This paper describes equipment developed for measuring the magnetic properties of small pieces of transformer laminations when energized with a 60-cycle magnetic field with or without superimposed direct-current polarization. A Maxwell bridge circuit was used in such a manner that the adjustable amount of magnetic gradient H which was applied to the sample remained unchanged by any manipulation of the bridge parameters in making a balance, thus permitting a direct-reading calibration of H. Bridge parameters were direct reading in dynamic permeability for a specified cross-sectional area and in a factor Δ from which the power loss per unit volume could be computed. Second-order correction factors are evaluated in the Appendix.*

A visual null-balance detector of the directional type was devised to be sensitive uniquely to either the reactive or the resistive balance control of the bridge. This was accomplished by the use of a polarized modulation bridge together with a phase-shifting network and facilitated balancing operations to a considerable degree. The necessary sensitivity and selectivity were obtained with a degenerative amplifier terminated in a current-limiting network which eliminated all necessity of monitoring an adjustable gain control.

The Problem

THE magnetic properties of laminated ferromagnetic materials, used in the assembly of transformer and inductor cores, are of some interest to communication engineers. A standardized method for the alternating-current testing of such materials at relatively low flux densities has been described.[1] This procedure, however, requires the use of over two pounds of the test material cut in rather large strips, 28 by 3 centimeters.

The author believes that some desire exists for a testing technique which permits the use of much smaller samples, such as might be obtained from small transformer laminations. It is also of interest to study the magnetic properties of such samples when functioning at extremely small flux densities approaching initial permeability, and to subject them to an independent but simultaneous direct-current magnetization. Accordingly, equipment intended for 60-cycle measurements and designed to be as convenient and direct-reading as possible, has been developed for this purpose.

Fundamental Considerations

In order to simplify the electromagnetic relationships involved, the ferromagnetic path will be considered to be homogeneous in character, to be continuous with an effective length of l centimeters, to have a constant cross-sectional area of A square centimeters, and to contain, with uniform distribution, all of the flux ϕ produced, thus having no magnetic leakage. If such a path, initially unmagnetized, is circumscribed N times by a solenoidal circuit carrying a steady current of I amperes, there will be created at each point in the path a gradient H of magnetomotive force

$$H = \frac{0.4\pi NI}{l} \tag{1}$$

measured in oersteds, and a flux density or magnetic induction

$$B = \frac{\phi}{A} \tag{2}$$

measured in gausses.

Regarded as a vector, the flux density has two component parts

$$\mathbf{B} = c\mathbf{H} + 4\pi\mathbf{J} \tag{3}$$

wherein the field intensity $\mathbf{H}$ is independent of matter and has for its scalar value the gradient H, so that c is a constant with the dimensions gauss/oersted. All phenomena of ferromagnetism depend upon the second component, the magnetization $\mathbf{J}$, which is positive in ferro- and paramagnetic media, negative in diamagnetic media, zero in all nonmagnetic media, and which is not a linear function of the applied gradient.

If the three vectors $\mathbf{B}$, $\mathbf{H}$, and $\mathbf{J}$ are mutually parallel, then c has a value of unity and the ratios of the scalar values

$$\mu = \frac{B}{H} \tag{4}$$

and

$$\kappa = \frac{J}{H} \tag{5}$$

define the static or direct-current permeability μ and susceptibility κ of the specimen material and are related by the equation

$$\mu = 1 + 4\pi\kappa. \tag{6}$$

The magnetic parameters B, J, μ, and κ define intrinsic properties of the material and, together with H, are independent of the particular geometry of the path.

The nonlinear increase of the magnetization J with applied gradient H and its ultimate attainment of a constant saturation value cause the B-versus-H curves of ferromagnetic materials to start with a relatively small value of slope which, at first, increases with H but finally recedes to unity when complete saturation is reached. The static permeability corresponding to any given gradient H_x, see Fig. 1, is not the slope of the

* Decimal classification: R282.3. Original manuscript received by the Institute, August 21, 1940. Presented, Fifteenth Annual Convention, Boston, Mass., June, 28, 1940.

† General Radio Company, Cambridge, Mass.

[1] American Society of Testing Materials Standards (A34–39) 1939 edition, part I, pages 523 et seq.

B-versus-H curve, at that point, but rather the slope of a secant drawn from the origin to the point. Consequently, the μ-versus-H curve starts at some initial value μ_o, increases to a maximum value μ_{max}, and finally recedes to progressively lower values as a saturation of J is approached. Thereafter any further increase of B can be produced only by an increase of the field intensity H.

If the steady current in the solenoid is replaced by a sinusoidal alternating current the instantaneous values

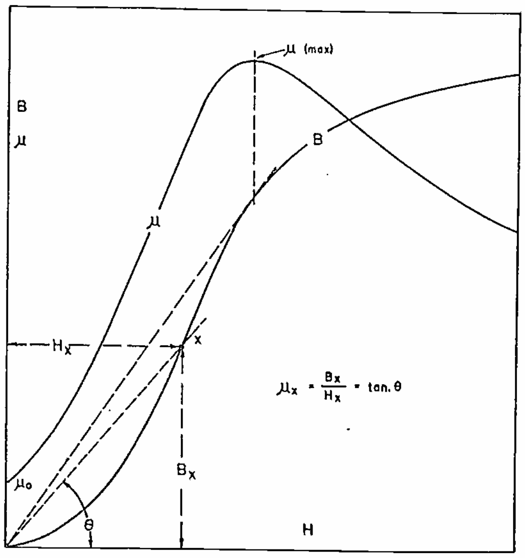


Fig. 1—Static (direct-current) permeability.

of B will follow the contour of the well-known hysteresis loop and the magnetic parameters will acquire a dynamic significance. The dynamic or alternating-current permeability μ_d may then be determined as a function of the electromotive force induced in the solenoid for any specific peak value of applied gradient H. This dynamic permeability may be regarded as a proper time average of the instantaneous slopes (dB/dH) at all points around the periphery of the hysteresis loop throughout a complete cycle. While the dynamic and static values of μ may differ somewhat from each other, they both undergo essentially similar variations as the gradient is increased.

If the specimen material is brought by direct current to any specific point on the static B-versus-H curve and then subjected to a cyclic excursion by a superimposed alternating current, the instantaneous B values will trace a displaced hysteresis loop and the magnetic parameters will have an incremental significance. The incremental permeability μ_i may likewise be evaluated in terms of the induced electromotive force for any specific values of static and peak dynamic gradient.

It is evident that (1) the static value of μ_{max} will be the slope of the static B-versus-H curve at the corresponding value of H; (2) as the dynamic excursion is reduced to zero, the incremental permeability approaches the slope of the static B-versus-H curve at the static H value; (3) for any given dynamic excursion, dynamic and incremental permeabilities become equal as the static gradient approaches zero; (4) dynamic permeability approaches initial (static) permeability as the dynamic excursion is diminished to zero. Statement (2) shows that a close approximation of all direct-current magnetic data could be obtained by actual alternating-current incremental measurements of μ using very small excursions and subsequently developing the static B-versus-H curve in terms of incremental data.

The present equipment is intended to measure either the dynamic or the incremental permeability of the specimen material in terms of an applied dynamic and polarizing static gradient. From such data the corresponding values of the parameters B, J, and κ may be computed.

When the magnetic phenomena is analyzed, the gradient H may be considered to be the initial magnetic effort applied to the specimen, while the flux density B may be regarded as the final result attained. A study of the dynamic or incremental behavior of these two quantities when they are subjected to a cyclic variation with time becomes simplified and more significant if one of them undergoes a purely sinusoidal variation, since both cannot vary in a sinusoidal manner simultaneously.

If the solenoid winding around the specimen is energized by a sinusoidal electromotive force, then the flux-density variation will be sinusoidal, provided that the reactance of the solenoid constitutes essentially the entire impedance of the circuit. The current in the solenoid and hence the dynamic gradient will have superimposed upon them the odd harmonics of the fundamental frequency, due to the curvature of the B-versus-H curve of the specimen (variation of μ). This hypothesis is utilized in the American Society of Testing Materials method of testing.

The author has used the alternative condition wherein a sinusoidal electromotive force is applied across the solenoid in series with a linear impedance whose value is large compared with the nonlinear impedance of the solenoid. In this case the current in the solenoid, controlled essentially by the linear impedance, must be sinusoidal and produce a sinusoidal variation of the magnetic gradient. The flux density and, accordingly, the back electromotive force induced in the solenoid will now have odd harmonic components in their variation, due to the nonlinear characteristics of the specimen.

The measurement of the solenoid by any suitable impedance bridge would give its inductance L and its equivalent series resistance R, representing core and copper losses, both of these factors L and R being dependent upon the current through the solenoid and its frequency. From these data the magnetic properties of the specimen material and the power loss in it may be evaluated.

Assuming no leakage, the linkage of this electromagnetic system will equal the product ϕN (maxwell-turns), so that the inductance of the solenoid will be

$$L = \frac{\phi N}{10^8 I} = \frac{0.4\pi N^2 \mu_d A}{10^8 l} \qquad (7)$$

henrys. In reality, the second member of (7) defines the fundamental or static inductance while the third member gives the dynamic or effective inductance with an alternating current. This permits the dynamic permeability of the material, for any given peak gradient, to be determined in terms of the dynamic inductance of the solenoid and the geometrical factors N, A, and l of the system

$$\mu_d = \frac{10^8 Ll}{0.4\pi N^2 A}. \qquad (8)$$

If an effective alternating current of I amperes, having a frequency of ω radians per second, is passed through the solenoid the latter will have a total power dissipation in watts

$$P_t = I^2 R. \qquad (9)$$

If r is the alternating-current resistance of the solenoid winding, which is responsible for copper loss, then the magnetic core loss, due to hysteresis and eddy currents in the ferromagnetic material, will be

$$P_m = I^2(R - r). \qquad (10)$$

Discounting any temperature coefficient, r is independent of I and has a unique value for any given frequency. However, at frequencies such as 60 cycles per second where the skin effect is negligible, r may be taken to be the direct-current resistance of the solenoid winding.

The Measurement of μ_d and P_m

A Maxwell bridge, indicated in Fig. 2, was used for measuring the parameters L and R of the solenoid. The bridge contained a fixed resistor R_1, a fixed capacitor

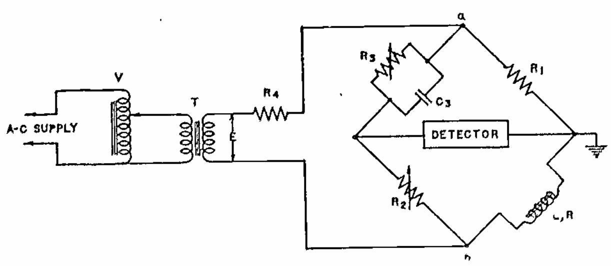


Fig. 2—The Maxwell bridge adapted for dynamic (alternating-current) measurements.

C_3, and two independent, adjustable resistors R_2 and R_3, and was energized through a Variac V, a shielded isolation transformer T, and a fixed resistor R_4. The value of the latter was made large compared with the impedance across the corners a–b of the bridge to assure a sinusoidal current in the solenoid. When the bridge is balanced the Kirchhoff equations give

$$L = (R_1 C_3) R_2 \qquad (11)$$

$$R = R_1 \left(\frac{R_2}{R_3}\right). \qquad (12)$$

By the proper choice of R_1 and C_3 (see Appendix) the sinusoidal current in the solenoid becomes essentially the ratio of E to R_4; hence,

$$H = \left[\frac{0.4\pi N}{l R_4}\right] E \qquad (13)$$

wherein the coefficient of E is a fixed quantity which permits the Variac control adjusting E to be calibrated directly in oersteds.

Equations (8) and (11) give

$$\mu_d = \left[\frac{10^8 l R_1 C_3}{0.4\pi N^2 A}\right] R_2 \qquad (14)$$

so that the control adjusting R_2 may be calibrated directly in terms of the dynamic permeability.

Finally, by combining (1), (10), (12), and (14) the magnetic power loss in watts per cubic centimeter may be written in the form

$$\frac{P_m}{lA} = H^2 \left(\frac{\mu_d}{\Delta} - K\right) \qquad (15)$$

wherein the control adjusting R_3 may be calibrated directly in terms of the divisor

$$\Delta = [0.4\pi C_3 \times 10^8] R_3 \qquad (16)$$

and the constant term

$$= \frac{lr}{0.16\pi^2 N^2 A} \qquad (17)$$

involves only known invariable factors.

With the bridge calibrated for any specific values of l, A, and N and with the value of r, and hence the term K, known, the Variac may be set to give directly any desired gradient H. Balancing the bridge will then give directly the corresponding values of the specimen permeability μ_d and the divisor Δ. The power loss in the specimen may then be computed by (15). Repeating for other values of gradient gives data for the μ-versus-H and P_m/lA-versus-H curves of the specimen material. Any other desired data may then be computed by the use of (2) to (6).

This method of measuring samples of laminated stock may readily be applied at frequencies other than 60 cycles per second as used by the author. In so doing, the evaluation of Q in terms of Δ by (18) in the Appendix and the corresponding data tabulating the several second-order correction factors would be modified correspondingly.

Applying a Unidirectional Magnetic Gradient

To subject the specimen to a simultaneous direct-current gradient, which may or may not exceed the peak dynamic value of H, the bridge was modified, as shown in Fig. 3, by the addition of a direct-current source E', which was either a battery or the well-filtered anode supply of the detector amplifier, and a control resistor R' which was kept large compared with the reactance of the solenoid (see the Appendix). To assure that all of the direct current I', measured by a milliammeter, passed through the solenoid, three blocking capacitors were required. The value of C_d depended upon the design of the null detector; the reactance of C_4 was small compared to R_4; while the

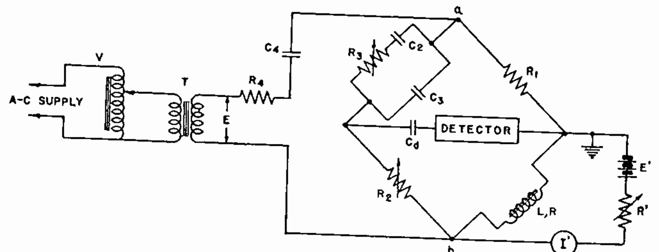


Fig. 3—The Maxwell bridge modified for incremental measurements.

capacitance of C_2 was sufficiently large compared to C_3 that it introduced no appreciable error into the bridge equations (see Appendix). The direct-current or static magnetic gradient H' applied to the specimen was then determined by substituting I' for I in (1) while, for any superimposed value of dynamic gradient, the bridge data now gave incremental values of permeability and loss.

The Test Yoke

It was considered desirable to have the test strip rectangular in form. Accordingly, the magnetic path shown in Fig. 4 was employed in which the strip had a nominal length of 3 inches (7.62 centimeters) and a nominal width of $\frac{3}{8}$ inch (0.95 centimeter). The test strip was symmetrically interposed between two stacks of return-path leaves which consisted of square laminations from which a central circular hole $1\frac{3}{4}$ inches (4.45 centimeters) in diameter had been removed. These leaves were made of mu-metal and subsequently were hydrogen annealed so that they possessed a high initial permeability. If four or more leaves were used on each side of the test strip, the total reluctance of the two-way return path was, in most cases, negligibly small compared to the reluctance of the test strip bridging the central hole, so that essentially all of the magnetomotive force created ($0.4\pi NI$) was consumed in establishing the flux in the test strip. The strip could then be considered to be the *entire* magnetic path for the purpose of applying the foregoing equations.

The validity of this assumption, together with the absence of appreciable magnetic leakage, was verified by comparing the μ-versus-H curves of a number of samples measured, first individually and, subsequently, in various parallel combinations.

The test strip could, of course, exceed 3 inches in length and project beyond the limits of the leaves. By progressively shortening the test strip, placed symmetrically, it was found that a reduction of length to 2.5 inches introduced a maximum error, at μ(maximum), of only 4 per cent. Analysis of such data indicated that, with the physical length of the strip 2.5 inches or any greater value, the mean flux path in the system was such that the mean magnetic length l of the strip was something in excess of the hole diameter and could be considered close to 1.90 inches (4.83 centimeters).

If desired, several test strips could be measured simultaneously in parallel, provided that at least 4 leaves were interposed between adjacent strips in assembling the yoke stack. The cross section A in the equations could be computed from the number, width, and average thickness of the strip laminations, or perhaps more accurately from their mass, density, and physical length.

With a unique value of l, the calibration of the gradient H is definitely determined. Since both l and A enter into the determination of μ and K, the calibration of the μ dial and the value of the term K can be definite only for a specified A. For any other value of A a correction factor (ratio of specified A to actual A) may be applied directly to the calibration of μ and the specified value of the constant K.

Fig. 5 shows a view of the assembled test yoke in which the stack of leaves and test strips were clamped between two substantial bakelite blocks having the contour of the leaves. It was found that clamping the stack approximately finger-tight served to reduce to a definite (negligible) minimum the air-gap reluctance between adjacent leaf and strip pieces. Further tightening produced no appreciable effect, which would indicate that finger-tightening minimized contact re-

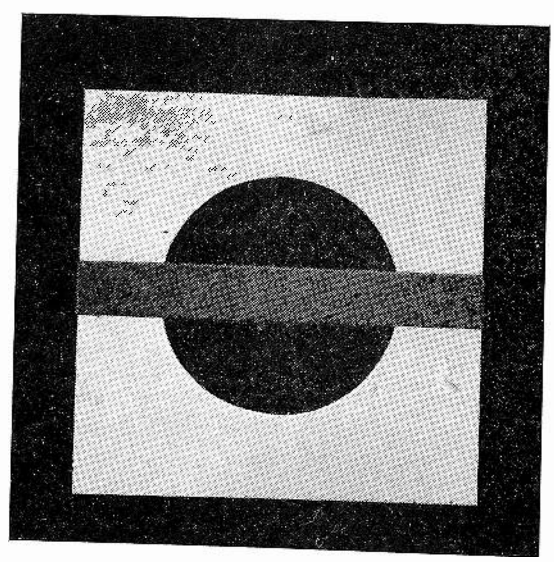


Fig. 4—Test strip and return-path leaf forming the magnetic circuit.

luctance without introducing undue strains in the material. The coil surrounding the test strip had an internal cross section 25/64 inch square and its winding ($N=400$ turns) was distributed over essentially the entire exposed length of the test strip.

The magnetic path of this test yoke was found, in practice, to be as satisfactory as the toroidal path used by some investigators. While the latter eliminates all possibility of air-gap reluctance, it requires special annular laminations of the test material and an individual winding of each specimen. Furthermore, rectangular strips permit the magnetization to be established in any desired direction with respect to the grain of the specimen, a feature impossible with a toroidal path.

In common with universal practice it was found desirable, after assembling and clamping the stack leaves and strips, to subject them to several applications of an increasing and decreasing alternating-current gradient, somewhat in excess of the contemplated maximum value to be used subsequently in obtaining test data for the specimen.

The Null-Balance Detector

The null detector used in balancing the Maxwell bridge required a high sensitivity for measurements at low magnetization and considerable selectivity at larger values of H in order to eliminate any response to the harmonics of electromotive force inherently introduced by the nonlinear characteristic of the ferromagnetic specimen. These two prime requisites were met, in part, by the use of a degenerative amplifier.[2] The detector system, Fig. 6, contained several features which were found to be very advantageous in making rapid and convenient measurements. While none of these elements is individually novel, it is believed that they were first used in combination in the present equipment.

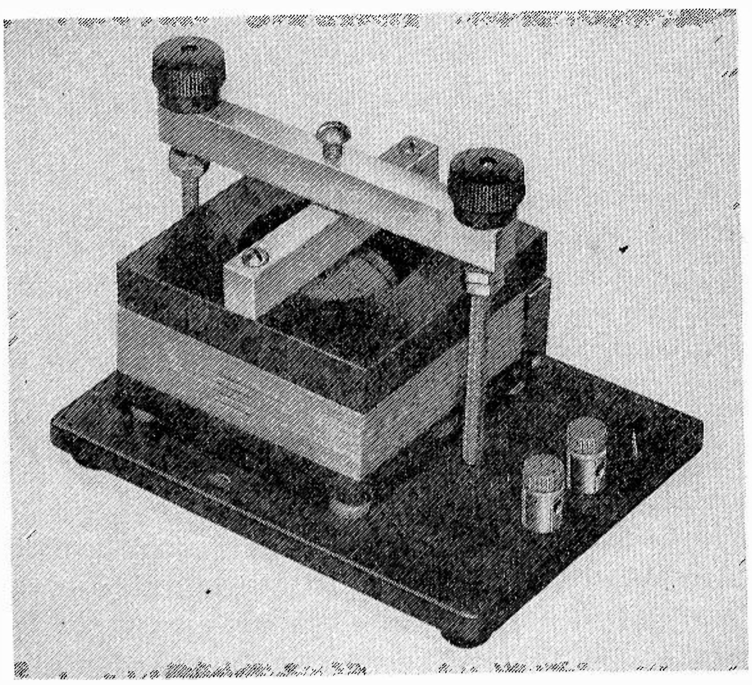


Fig. 5—Assembly of test yoke.

The visual balance indicator, a zero-center-scale galvanometer G, was inserted in a modulation bridge, which was polarized through a phase-shifting network permitting two alternative phases 90 degrees apart to be used. By the choice of one or the other of these phases, the indicator could be rendered uniquely sensitive to the manipulation of either the μ or the Δ control of the Maxwell bridge. This gave the detector a phase-selective feature, lacking which any impedance bridge must be balanced by an alternate adjustment of its two controls through successively diminishing minima, while the detector remains equally sensitive to signals due to off-balance of either control. This latter and customary operation is somewhat tedious, especially if the bridge has a sliding zero, which means that the

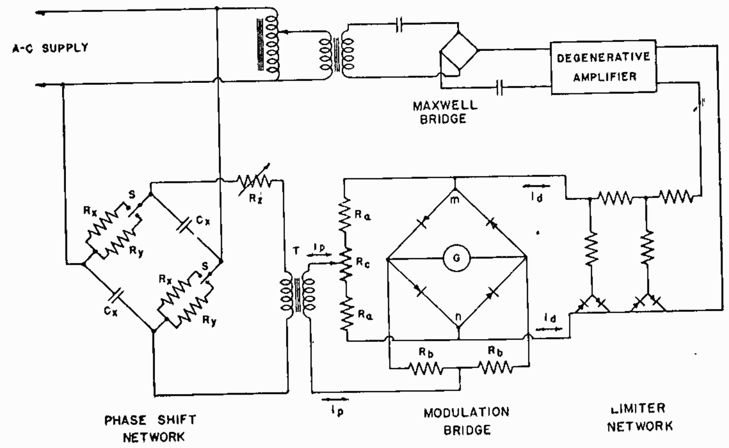


Fig. 6—The selective and directional null detector system.

settings of the bridge controls for successive minima do not coincide with their values at exact balance.

The modulation bridge also gave the indicator a directional feature by indicating, in a left or right motion of the galvanometer pointer, whether the particular bridge control to which it was sensitive was off-balanced by being too high or too low in value, and thereby giving a cue as to which way that control should be manipulated to secure a balance, after the manner of the galvanometer used with a direct-current Wheatstone bridge.

Finally, a special type of limiter network[3] gave the galvanometer its maximum sensitivity at balance but prevented it from ever going off-scale and thereby eliminated any need for monitoring the amplifier gain or for a galvanometer shunt while balancing the Maxwell bridge.

This null detector system could be adapted for use with any impedance bridge. It may be noted in passing that if the initial power source is sinusoidal in electromotive force, the modulation bridge will be quite immune to any even harmonics and will possess a selectivity of about 18 decibels against any third harmonic in the output of the impedance bridge. Furthermore, the modulation bridge, of itself, had an equivalent gain of the order of 40 decibels above the customary rectifier bridge employing the same elements. Except for measurements near initial permeability, the modulation bridge could be used directly without the degenerative amplifier, provided that a suppressor network, similar to that used in the amplifier but designed to eliminate the third harmonic, was substituted for the amplifier in Fig. 6.

Each arm of the modulation bridge (see Appendix) consisted of a copper-oxide rectifier directed as shown,

[2] H. H. Scott, "A new type of selective circuit and some applications," Proc. I.R.E., vol. 26, pp. 226–235; February, 1938.

[3] W. N. Tuttle, U. S. Patent 2,104,336.

so that detector signals gave no response unless these rectifier elements were properly polarized by a current from the same source which energized the Maxwell bridge. The phase-shifting network, described by Turner and McNamara[4] and previously used by the author,[5] contained two equal fixed capacitors C_x and two pairs of fixed resistors R_x and R_y whose selection, by the double-pole double-throw switch S, determined the two alternative phases of the current polarizing the modulation bridge and the selective sensitivity of the latter to the two controls of the Maxwell bridge. Since the parameters R_1 and C_3 of the latter were fixed, a single set of R_x and R_y values sufficed to render the detector essentially sensitive to either the μ or the Δ control.

The output from the phasing network was applied to the modulation bridge through an adjustable resistor R_z, an isolation transformer, two pairs of fixed resistors R_a and R_b, and an adjustable potentiometer R_c. Due to discrepancies in the current-versus-potential characteristics of the 4 rectifier elements, the modulation bridge was initially balanced for a zero galvanometer reading with zero-amplifier signals: first, by adjusting the polarizing current i_p by the resistor R_z with the points m and n shorted, and second, by adjusting R_c to rebalance with this short circuit removed. By keeping R_z large compared to the reactance of C_x, the magnitude of the polarizing current i_p was rendered independent of its phase.

The two resistive L pads comprising the limiter network contained, in each of their shunt arms, two copper-oxide rectifier elements in parallel but oppositely directed so that they introduced no rectification. The characteristics of these elements, however, caused the effective resistance of the shunt arms to diminish and, hence, the attenuation of the network to increase with the applied signal, thereby giving the useful limiter action.

Some Phenomena at Initial Permeability

The present paper is intended to describe a method of magnetic measurement, rather than to give a compilation and analysis of extensive data taken therewith. However, some representative data obtained at very low magnetic gradients, Fig. 7, may be of interest. It is well known that, starting at initial permeability, the value of μ increases to a maximum and subsequently decreases, see Fig. 1. It is not so generally realized, however, that as the gradient H approaches zero the slope of the μ-versus-H curve becomes zero, as permeability acquires its constant initial value. At these low values of H, therefore, the B-versus-H curves are strictly linear with a finite slope μ_o and the ferromagnetic material possesses no harmonic distortion due to varying permeability. To exhibit these interesting phenomena with ordinary silicon-steel transformer laminations, the author found that the applied dynamic gradient must be kept below a maximum value of about one millioersted.

Acknowledgment

The author wishes to thank his colleagues, Messrs. R. F. Field and P. K. McElroy, for helpful suggestions in the preparation of this paper.

Appendix

The direct-reading calibration of the Variac dial in terms of the magnetic gradient H, and the modification of the Maxwell bridge made to permit a simultaneous direct-current magnetization of the sample, introduced

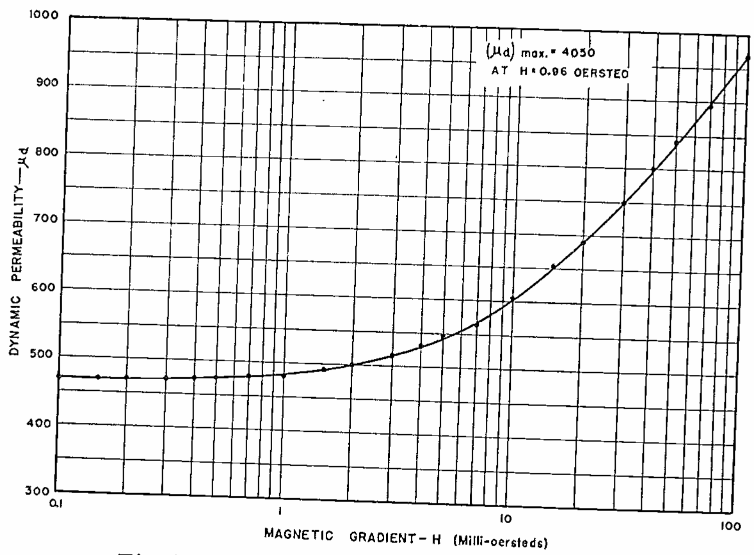


Fig. 7—Initial permeability in silicon steel.

three assumptions which were quite valid for most uses of this equipment, especially for relative measurements between different samples of the same general character. For more precise absolute measurements, however, the following correction factors may be applied, if desired, to the observed values of H, μ and the term μ/Δ occurring in the evaluation of the core loss.

It is most convenient to determine these correction factors in terms of the inductance L and the storage factor Q of the solenoid. A sufficiently accurate value of each factor may, however, be computed by using the apparent bridge data at balance, and then evaluating L by substituting the observed value of μ in (7). Q is determined most readily from the equation

$$Q = \frac{\omega\Delta}{0.4\pi \infty 10^8}. \tag{18}$$

which, at a frequency of 60 cycles per second, becomes

$$Q = \frac{3\Delta}{10^6} \tag{18'}$$

Subdivision of Total Bridge Current

If I is the total bridge current and I_x is the current

[4] H. M. Turner and F. T. McNamara, "An electron tube wattmeter and voltmeter and a phase shifting bridge," Proc. I.R.E., vol. 18, pp. 1743–1747; October, 1930.

[5] Horatio W. Lamson, "An electronic null detector for impedance bridges," *Rev. Sci. Instr.*, vol. 9, pp. 272–275; September, 1938.

through R_1 and the solenoid, then when the bridge is balanced

$$I_x = \frac{z_{ab}}{z_x} I \qquad (19)$$

wherein z_{ab} is the scalar impedance across the corners a–b of the bridge, and z_x is the scalar impedance of R_1 and the solenoid in series. Expressed in terms of the inductance L and the storage factor Q of the solenoid and of the fixed parameters R_1 and C_3 of the bridge, the vector values of these two impedances are

$$Z_{ab} = \frac{Q^2R_1(1+\omega^2LC_3)+\omega Q(L+C_3R_1^2)+\omega^2C_3R_1L+j\omega Q^2(L-R_1^2C_3)}{\omega^2C_3^2R_1^2Q^2+(Q+\omega R_1C_3)^2} \qquad (20)$$

$$Z_x = R_1 + \frac{\omega L}{Q} + j\omega L. \qquad (21)$$

In the 60-cycle bridge used, the fixed parameters were chosen to be: $R_1 = 100$ ohms; $C_3 = 1$ microfarad. Consequently,

$$Z_{ab} = \frac{100Q^2(1+0.142L)+377Q(L+0.01)+14.2L+j377Q^2(L-0.01)}{0.00142Q^2+(Q+0.0377)^2} \qquad (20')$$

$$Z_x = 100 + \frac{377L}{Q} + j377L. \qquad (21')$$

The scalar values of these impedances and their ratios are then given, for specific values of L and Q, in Table I.

TABLE I

	$Q=1$	$Q=2$	$Q=5$	$Q=10$
L	1.0	1.0	1.0	1.0
z_{ab}	586	473	413	399
z_x	608	475	416	401
z_{ab}/z_x	0.963	0.995	0.994	0.995
L	0.1	0.1	0.1	0.1
z_{ab}	137.7	122.7	113.0	109.5
z_x	142.6	124.7	113.9	110.4
z_{ab}/z_x	0.967	0.985	0.993	0.995
L	0.01	0.01	0.01	0.01
z_{ab}	100.1	100.1	100.2	100.0
z_x	103.8	101.9	100.8	100.4
z_{ab}/z_x	0.966	0.983	0.993	0.996
L	0.001	0.001	0.001	0.001
z_{ab}	96.7	98.3	99.4	99.6
z_x	100.4	100.2	100.1	100.0
z_{ab}/z_x	0.963	0.981	0.993	0.996

These current subdivision ratios z_{ab}/z_x may be multiplied into the calibrated value of H to obtain the true gradient. However, the error introduced by assuming a unity ratio will be less than 2 per cent if the Q ratio of the solenoid exceeds 2 and will be only 3.7 per cent should the Q ratio fall to unity.

The Shunting Effect of R'

When a direct-current gradient is applied through a resistor R', the vector impedance of the solenoid arm of the bridge becomes

$$Z' = \frac{RR'(R+R')+\omega^2L^2R'+j\omega LR'^2}{(R+R')^2+\omega^2L^2}. \qquad (22)$$

Then defining the ratio

$$\beta = \frac{\omega L}{R'} \qquad (23)$$

it can be shown that the effective inductance of this arm has been reduced by the factor

$$\alpha = 1 + 2\frac{\beta}{Q} + \beta^2\left(1+\frac{1}{Q^2}\right) \qquad (24)$$

so that the observed value of μ should be increased by this factor.

The effective Q of this arm is likewise reduced by the factor

$$\gamma = 1 + \frac{\beta}{Q} + \beta Q. \qquad (25)$$

Since the R term in (10) is equal to the ratio $\omega L/Q$, the μ/Δ term in (15) may be corrected by multiplying by the factor α/γ. Choosing 10,000 ohms for the value of R', these factors, for specific values of L and Q, are given in Table II. Except for high inductances with low Q values, the correction to μ is negligible; while with high inductances having high Q values, an appreciable correction must be applied to the μ/Δ term unless a larger value of R' is used. These corrections are, of course, necessary only when making incremental measurements.

TABLE II

	$Q=1$	$Q=2$	$Q=5$	$Q=10$
L	1.0	1.0	1.0	1.0
β	0.0377	0.0377	0.0377	0.0377
α	1.078	1.040	1.017	1.009
γ	1.075	1.094	1.196	1.381
α/γ	1.003	0.951	0.850	0.730
L	0.1	0.1	0.1	0.1
β	0.00377	0.00377	0.00377	0.00377
α	1.008	1.004	1.002	1.0008
γ	1.008	1.009	1.020	1.0381
α/γ	1.000	0.995	0.983	0.963
L	0.01	0.01	0.01	0.01
β	0.000377	0.000377	0.000377	0.000377
α	1.0008	1.0004	1.0002	1.0001
γ	1.0008	1.0009	1.0020	1.0038
α/γ	1.000	0.9995	0.9982	0.9963

Error Due to Blocking Capacitor C_2

The introduction of the capacitor C_2 in series with R_3 modifies the simple Kirchhoff equations (11) and (12) at balance so that they become

$$L = R_1R_2C_3\delta \qquad (26)$$

and

$$R = \frac{R_1R_2}{R_3}\epsilon \qquad (27)$$

wherein the correction factors δ and ϵ are

$$\delta = 1 + \frac{2}{nQ^2 - 2 + Q\sqrt{n^2Q^2 - 4(n+1)}} \tag{28}$$

$$\epsilon = 1 - \frac{2}{nQ[nQ + \sqrt{n^2Q^2 - 4(n+1)}] - 2n} \tag{29}$$

and

$$n = \frac{C_2}{C_3}. \tag{30}$$

The scale reading of μ, therefore, may be multiplied by the factor δ, while the computed value of the term μ/Δ may be multiplied by the factor ϵ. Note that the bridge cannot now be balanced unless Q^2 exceeds the ratio $4(n+1)/n^2$.

If C_2 is made ten times C_3 the values of these correction factors, for specific values of Q, become

	$Q=1$	$Q=2$	$Q=5$	$Q=10$
δ	1.1292	1.0264	1.0040	1.0010
ϵ	0.9871	0.9974	0.9996	0.9999

These factors may be taken as unity except when low Q values cause an appreciable increase in δ.

For incremental measurements the complete correction to the μ dial reading should then be the product $\alpha\delta$ and the complete correction to the μ/Δ term is the ratio $\epsilon\alpha/\gamma$.

It may be noted that the equivalent series resistance of C_2 merely becomes a part of R_3. A high-quality capacitor is required for C_3 in order that its low dissipation factor may introduce no appreciable errors into the bridge equations.

Construction of the Modulation Bridge

The modulation bridge used by the author was initially constructed in the following manner. If the two alternating-current terminals of an ordinary rectifier type of bridge are tied together, a unit is formed whose extremities are the two direct-current terminals and whose mid-tap is the united alternating-current terminals. This unit will then consist of two rectifiers in series and like-directed, each rectifier actually being two copper-oxide elements in parallel. By the use of two such units the four-rectifier modulation bridge shown in Fig. 6 may be assembled.

Subsequently, a single-unit modulation type of bridge was procured for the purpose. An analysis of the functional operation of this bridge may be reserved for another paper.

Equivalent Electrostatic Circuits for Vacuum Tubes*

W. G. DOW†, MEMBER, I.R.E.

Summary—*A method of field analysis by conformal transformation is used to demonstrate that the electrostatic properties of a triode may be represented by three capacitances in star, whose magnitudes are related to tube geometry in simple fashion. The method is then extended to the construction of an equivalent electrostatic circuit for multigrid tubes.*

It is shown how there can be derived from the electrostatic circuit (a) good approximations to the potential distribution in various parts of the tube and around these first approximations more accurate determinations of the local potential distribution; (b) a simple and rational expression of the dependence of cathode-current flow upon the various electrode potentials, for both parallel-plane and cylindrical geometry; (c) the ordinary interelectrode capacitances, as far as these are affected by the structure of the active portions of the electrodes.

The application of this general method to structures having nonregular geometry and to other problems of current and potential distribution is discussed.

I. Introduction

THIS paper outlines a framework of ideas, regarding the internal operation of high-vacuum tubes, which the author has found useful both in instruction and analysis. Its chief merit is that of employing simple electric-circuit principles, familiar to all electrical engineers, for analyzing electronic phenomena in vacuum tubes. The underlying concepts made use of here probably were first employed by L. A. Hazeltine, in the form of unpublished lecture notes, in the early days of vacuum-tube analysis.

* Decimal classification: R131. Original manuscript received by the Institute, April 15, 1940; revised manuscript received, December 2, 1940. Presented, Fifteenth Annual Convention, Boston, Mass., June 28, 1940.

† University of Michigan, Ann Arbor, Mich.

II. The Equivalent Electrostatic Circuit of a Triode

The over-all electrostatic properties of any space-charge-free arrangement of three electrodes, whether in a vacuum or elsewhere, can be represented by three properly chosen condensers arranged in delta,[1] Fig. 1(a), or three other properly chosen condensers arranged in star, Fig. 1(b).

Typical interrelations between the star and delta capacitances in Fig. 1 are as follows:[2]

$$C_{cg} = \frac{C_{cG}C_{gG}}{C_{cG} + C_{gG} + C_{pG}} \tag{1}$$

$$C_{cG} = \frac{C_{cg}C_{gp} + C_{gp}C_{pc} + C_{pc}C_{cg}}{C_{gp}}. \tag{2}$$

Two other pairs of similar relations can obviously be obtained by symmetrical rearrangement of symbols.

Simple expressions for the star capacitances when no space charge is present are derivable from known electrostatic-field properties. As shown in Appendix I, sufficiently close approximate expressions, for a parallel-plane triode, Fig. 2, are as follows:

[1] I. Langmuir and K. T. Compton, "Electrical discharges in gases, part II," *Rev. Mod. Phys.*, Vol. 3, p. 206; April, 1931.

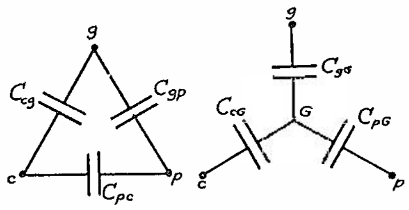
[2] R. R. Lawrence, "Principles of Alternating Currents," Second Edition, p. 312, McGraw-Hill Book Company, New York, N.Y.

$$C_{cG} = \frac{1}{4\pi a} \tag{3}$$

$$C_{pG} = \frac{1}{4\pi b} \tag{4}$$

$$C_{gG} = \frac{n}{2 \log \coth 2\pi n R}. \tag{5}$$

These capacitances are in statfarads per square centimeter. Obviously C_{cG} as so obtained is the capacitance that would exist between the cathode and a flat



(a) Delta (b) Star

Fig. 1—Triode electrostatic circuits.

electrode in the plane of the grid, and C_{pG} that between the plate and the same "grid-plane" electrode.

These facts lead to the introduction of the "equivalent grid-plane" concept.[3] The point G in the star electrostatic diagram, Fig. 1(b), represents, of course, a purely fictitious electrode. However, we may conveniently visualize it as a "make-believe" electrode, lying in the plane of the grid, but stopping just short of each grid wire. This concept leads to the form of the star diagram that is illustrated in Fig. 2.

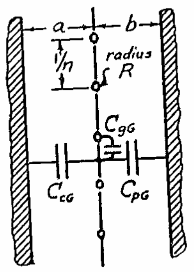


Fig. 2—Positional diagram of the star electrostatic circuit of a triode.

Thus C_{cG} and C_{pG} are the capacitances between an equivalent grid-plane electrode and the cathode and plate, respectively, while C_{gG} is that between the grid wires and the plane in which they lie.

These concepts may be extended to cylindrical triodes, C_{cG} and C_{pG} becoming capacitances between concentric cylinders, and C_{gG} being evaluated by multiplying a value per unit area, equation (5), by circumferential area $2\pi r_g$. The results are, in statfarads per centimeter of axial length,

$$C_{cg} = \frac{1}{2 \log r_g / r_c} \tag{6}$$

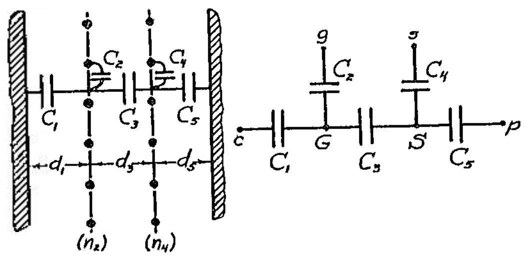
$$C_{pG} = \frac{1}{2 \log r_p / r_g} \tag{7}$$

[3] The equivalent grid-plane approach and a number of other concepts employed in this article were used extensively by B. J. Thompson and F. B. Llewellyn in their lectures at the University of Michigan Electronics Institute, July, 1937.

$$C_{gG} = \frac{2\pi n r_g}{2 \log \coth 2\pi n R}. \tag{8}$$

III. The Electrostatic Circuit for Multigrid Tubes[4]

The value of C_{gG} per unit area, equation (5), depends only on the structure of the grid itself, and is independent of the relative positions of grid plane, cathode, and plate. This dependence of C_{gG} *on grid structure only* suggests extension of the electrostatic diagram to multigrid tubes in the manner illustrated in Fig. 3 for a tetrode. Mathematical proof of this extension, using successive conformal transformations, is not difficult.



(a) Positional diagram (b) Schematic diagram

Fig. 3—Equivalent electrostatic circuit of a tetrode.

Fig. 3 represents a parallel-plane tetrode; the values of the C's in statfarads per square centimeter, are

$$C_1 = \frac{1}{4\pi d_1} \quad C_3 = \frac{1}{4\pi d_3} \quad C_5 = \frac{1}{4\pi d_5} \tag{9}$$

$$C_2 = \frac{n_2}{2 \log \coth 2\pi n_2 R_2} \quad C_4 = \frac{n_4}{2 \log \coth 2\pi n_4 R_4}. \tag{10}$$

Similar extensions of the cylindrical-tube relations to multigrid cylindrical structures are obvious.

IV. Use of the Electrostatic Circuit in Determining Potential Distribution

The electrostatic-circuit diagram provides an easy approach to the construction of potential-distribution diagrams for multigrid tubes, as illustrated for a tetrode in Fig. 4.

If the capacitances, and the potentials of grid, screen, and plate are known, the potentials E_G and E_S of the equivalent grid and screen planes can be calculated by ordinary electric-circuit methods, using the schematic circuit diagram, Fig. 3(b). Fig. 4(a) represents the next step in the construction of the corresponding potential-distribution diagram for a parallel-plane type of structure. This involves spotting the values of $E_c(=0)$, E_G, E_S, and E_p at the grid, screen, and plate positions on a space-potential diagram, then joining E_c to E_G, E_G to E_S, and E_S to E_p by straight lines.

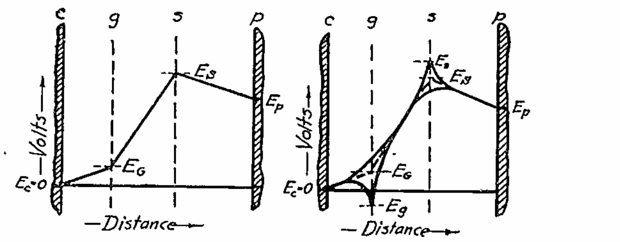
These straight lines represent the potential distribution reasonably correctly except in the near neighborhoods of the grid and screen. The reason for this is

[4] S. Koizumi, "On the amplification constants of multielectrode tubes," *Jour. I.E.E.*, (Japan) (abstracts), vol. 10, p. 18, 1934.

that for a considerable part of the regions midway between each electrode and its nearest neighbors the electric field is essentially uniform. In these uniform-field regions the equipotentials are parallel planes uniformly spaced, and the potential-distribution curves are therefore straight lines.

The extent to which the departure from field uniformity reaches out each way from grid and screen depends on the ratio of the grid-wire spacing $1/n$ to the interelectrode spacings, (d_1, d_3, d_5). If this ratio is small, the departure from uniformity is restricted to the very near neighborhood of the grids; if it is large, the departure from uniformity may reach out far enough to make the present analysis result in only a very rough approximation to the truth.

The nature of the potential distribution near grid and screen is illustrated in Fig. 4(b). Here the values



(a) Straight construction lines (b) Complete with fillets.
Fig. 4—Potential distribution in a tetrode (no space charge).

E_g and E_s of actual grid and screen potentials have been located, and "fillets" drawn showing potentials along the two extreme paths. At the grid location the upper, concave-upward fillet describes the potential distribution along a path midway between grid wires, and the two lower, concave-downward fillets describe that along a path through a grid wire.

The accurate mathematical determination of the between-grid-wire fillet[5] is obtained by writing the potential equation for a general point T along this path. This is done by use of (46) in Appendix I, letting $\phi' = \pi$; thus,

$$E_T - E_c = -2\tau_c \log s_c s_g \left| 1 + \frac{1}{r' s_g} \right| + 2\tau_p \log \left| \frac{r'}{s_g} + 1 \right|. \tag{11}$$

This equation describes the potential relative to the cathode, in statvolts. It is much more convenient to have a description relative to the equivalent grid-plane potential. An expression answering this requirement is obtained as follows:

Rewrite (11) in the following form:

$$E_T - E_c = -2\tau_c \log s_c - 2\tau_c \log \left| \frac{1}{r'} + s_g \right| + 2\tau_p \log \left| \frac{r'}{s_g} + 1 \right|. \tag{12}$$

[5] W. G. Dow, "Fundamentals of Engineering Electronics," John Wiley and Sons, New York, N. Y., 1937. Chapters II and V.

Now note that

$$E_G - E_c = 4\pi n a \tau_c = +2\tau_c \log \epsilon^{+2\pi n a} = -2\tau_c \log s_c. \tag{13}$$

Equation (13) comes from the fact that $-\tau_c$ is the charge in statcoulombs per grid section of cathode, so that $n(-\tau_c)$ is the charge per square centimeter of cathode area, likewise also the charge per square centimeter of area of the capacitance C_{cG}. But $C_{cG} = 1/4\pi a$ statfarads per square centimeter; (13) follows directly.

Subtraction of (13) from (12) leads to the following expression for potential along the fillet *midway between grid wires*,

$$E_T - E_G = -2\tau_c \log \left| \frac{1}{r'} + s_g \right| + 2\tau_p \log \left| \frac{r'}{s_g} + 1 \right|. \tag{14}$$

The corresponding expression for the potential distribution along a path through a grid wire is

$$E_T - E_G = -2\tau_c \log \left| \frac{1}{r'} - s_g \right| + 2\tau_p \log \left| \frac{r'}{s_g} - 1 \right|. \tag{15}$$

Of course this has no significance at points within the grid wire.

It is desirable to be able to determine the potential-distribution curve in the plane of the grid. Points for such a curve are obtainable by placing the general point T at Z-plane radius $r' = 1$ and subtracting (13). The mathematical expression for grid-plane potential distribution so obtained is

$$E_T - E_G = (\tau_p - \tau_c) \log \left| s_g + \frac{1}{s_g} - 2 \cos \phi' \right| - (\tau_p + \tau_c) \log s_g. \tag{16}$$

These equations for potential distribution near a grid may be put into ordinary (W-plane) form by using $\epsilon^{2\pi n x}$ for r', $2\pi n y$ for ϕ', and $\cosh 2\pi n R$ for s_g. The complete form, with discussions of certain detail points, is given in Appendix II. x represents the distance measured toward the plate from an origin at the center of a grid wire; y of course measures distance at right angles to x and to the grid wire.

Equations (14), (15), and (16) represent the potential distribution in the neighborhood of the grid entirely in terms of *local* quantities. These local quantities are (1) E_G; (2) the charge (per grid section) to the left of the grid $-\tau_c$; (3) the charge (per grid section) to the right of the grid τ_p; (4) $r' (= \epsilon^{2\pi n x})$; (5) $\phi' (= 2\pi n y)$; and (6) the grid-structure quantities n and $2nR$. Therefore the whole argument can be extended bodily to a screen, suppressor, or any other similar electrode for which the ratio of wire spacing to electrode spacing is small. Thus (14) and (15) permit construction of the fillets at grid and screen grids in such a geometry as that of Figs. 3 and 4. Of course $-\tau_c$ becomes the *total* charge per grid section to the left, τ_p that to the right, of the grid being considered. These τ's may be directly evaluated from the gradients to the left and right of the grid being considered.

For cylindrical geometry, the general procedure in arriving at a representation of the potential distribution is the same as given above for parallel planes, except that (a) logarithmic curves replace the straight lines between the grid positions, (b) the transformation equations to be used in connection with details near the grids are different, and (c) the charge per centimeter length of tube becomes $2\pi n r_g$ times that per grid section. (See Appendix II.) The logarithmic curves are obtained from the following equations: At radius r (actual, not Z plane) on the cathode side of the grid, expressed relative to E_G

$$E_T - E_G = -4\pi n r_g(-\tau_c)\log\frac{r}{r_g}. \qquad (17)$$

On the plate side of the grid

$$E_T - E_G = -4\pi n r_g(+\tau_p)\log\frac{r}{r_g}. \qquad (18)$$

Values for $-\tau_c$ and τ_p are obtained by placing T at adjacent equivalent electrode radii, inside and outside, respectively. Curves plotted according to these two equations take the place of the preliminary straight lines (Fig. 4(a)) used for the parallel-plane situation.

V. Determination of the Equivalent Grid-Plane Potential from the Electrostatic Circuit

When the cathode is heated, the space-charge-limited current that flows is determined by the location and potential of the equivalent grid plane. It is therefore important to consider the factors that control the equivalent grid-plane potential.

Referring to Fig. 1 again, one may write

$$(E_G - E_c)C_{cG} + (E_G - E_g)C_{gG} + (E_G - E_p)C_{pG} = 0. \qquad (19)$$

This can be solved for E_G, giving

$$E_G = \frac{E_cC_{cG} + E_gC_{gG} + E_pC_{pG}}{C_{cG} + C_{gG} + C_{pG}}. \qquad (20)$$

Potential is habitually measured relative to a zero value at the cathode, so that $E_c = 0$. Also, it is convenient to divide the numerator and the denominator by C_{gG}, and then rearrange into the following form:

$$E_G = \frac{E_g + \dfrac{E_p}{C_{gG}/C_{pG}}}{1 + \dfrac{1}{C_{gG}/C_{pG}} + \dfrac{1}{C_{gG}/C_{cG}}}. \qquad (21)$$

Now let

$\mu = C_{gG}C_{pG}$ (the usual "amplification factor"). (22)

Note also that

$$\frac{1}{C_{gG}/C_{cG}} = \frac{C_{pG}}{C_{gG}}\,\frac{C_{cG}}{C_{pG}} = \frac{1}{\mu}\,\frac{C_{cG}}{C_{pG}}. \qquad (23)$$

Thus the equivalent grid-plane potential for a triode can be expressed in the following generally useful form:

$$E_G = \frac{E_g + \dfrac{E_p}{\mu}}{1 + \dfrac{1}{\mu} + \dfrac{1}{\mu}\,\dfrac{C_{cG}}{C_{pG}}}. \qquad (24)$$

A similar procedure may be employed to determine an expression for the equivalent grid-plane potential of a tetrode, Fig. 3. The first step is to write expressions similar to (19) for both of the junction points G and S; thus,

$$(E_G - E_c)C_1 + (E_G - E_g)C_2 + (E_G - E_S)C_3 = 0 \qquad (25)$$

$$(E_S - E_G)C_3 + (E_S - E_s)C_4 + (E_S - E_p)C_5 = 0. \qquad (26)$$

These two equations may be solved simultaneously for E_G and E_S. The solution for E_G can be put into the following form:

$$E_G = \frac{E_g + \dfrac{E_s}{\mu_{gs}} + \dfrac{E_p}{\mu_{gp}}}{1 + \dfrac{1}{\mu_{gs}} + \dfrac{1}{\mu_{gp}} + \dfrac{1}{C_1/C_2}} \qquad (27)$$

where

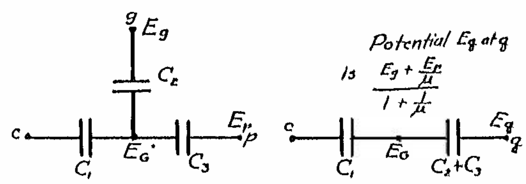
$$\mu_{gs} = \frac{C_2}{C_3}\left[\frac{C_3 + C_4 + C_5}{C_4}\right] \qquad (28)$$

$$\mu_{gp} = \frac{C_2}{C_3}\left[\frac{C_3 + C_4 + C_5}{C_5}\right]. \qquad (29)$$

The process just described can be extended to as many grids as desired.

VI. Reduction Formulas

The results obtained in the previous section can be arrived at by another method. For many purposes the star electrostatic circuit of the triode can be replaced by two condensers in series; see Fig. 5. The two condensers in series are C_1 and $C_2 + C_3$, respectively. If now a potential $E_q = [E_g + (E_p/\mu)]/[1 + (1/\mu)]$ is applied to the two in series the potential E_G will be as given by (21) and (24). To prove this it is only necessary to determine E_G as a fraction of the total voltage,



(a) Triode star circuit (b) Triode reduction

Fig. 5—Replacement of grid and plate capacitances by a single equivalent capacitance with an equivalent voltage.

using the fact that potential divides inversely as capacitance. Thus,

$$E_G = \frac{E_g + E_p\dfrac{C_2}{C_3}}{1+\dfrac{C_3}{C_2}}\cdot\frac{C_2+C_3}{C_1+C_2+C_3} = \frac{E_g + E_p\dfrac{C_3}{C_2}}{1+\dfrac{C_3}{C_2}+\dfrac{C_1}{C_2}}\cdot \quad (30)$$

which is identical with (27).

In an exactly similar way, for the tetrode of Fig. 3(b), the three capacitances C_3, C_4, and C_5 can be replaced by two in series. The two in series are of course C_3 and C_4+C_5, with a voltage $[E_s+(E_pC_4/C_5)]/[1+(C_4/C_5)]$ applied at the single terminal that replaces the s and p terminals. This can be reduced further to an appearance similar to Fig. 5(b) by a repetition of the process. This procedure is equivalent to the use of what some tube analysts call the q plane,[6] some the i plane.

VII. Space Charge, in Relation to Cathode Current and to its Effects on the Electrostatic-Circuit Capacitances[7,8]

Capacitance may be defined

(a) As the ratio of charge to potential, thus

$$Q = CE. \quad (31)$$

This is essentially a direct-current conception of capacitance for it permits defining C for direct-current conditions.

(b) As the ratio of current to rate of change of potential, thus

$$i = C\frac{de}{dt}\cdot \quad (32)$$

This is of course primarily an alternating-current conception of C.

(c) In terms of electrostatic-energy storage W, thus

$$W = \tfrac{1}{2}CE^2. \quad (33)$$

Some important uses of this definition have to do with mechanical-force relationships.

As long as (1) charge and potential are proportional to one another and (2) there is no space charge between the electrodes, these three defining equations are equivalent. In the case of complete space charge[8] (space-charge-limited current) charge and potential are proportional, yet the three defining equations given above are not equivalent. Thus the (1) and (2) qualifications just stated are distinct from one another.

Consider a simple parallel-plane diode, Fig. 6, with complete space charge. For any given current density there exists a definite potential-distribution curve ($E=kx^{4/3}$) terminating at a definite potential, and with definite slope, at the plate.

A very simple proof shows that a tangent to the potential-distribution curve at the plate intersects the zero-potential axis at three fourths of the distance from the plate to the cathode. This is true for all values of plate voltage. But the slope of a tangent at the plate is a direct measure of the charge on the plate surface. *Thus the charge on the plate is just the same as would exist in a space-charge-free condenser whose spacing is three fourths of the actual cathode-to-plate spacing*, if C is defined as the direct-current charge-to-potential ratio, equation (31).

For a parallel-plane triode with complete space charge, the direct-current potential-distribution curve

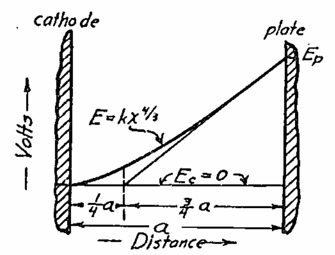


Fig. 6—Potential distribution in a parallel-plane diode under complete space-charge conditions.

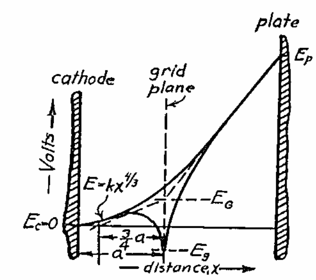


Fig. 7—Potential distribution in a triode under complete space-charge conditions.

in the neighborhood of the cathode is of course a 4/3-power curve like that in Fig. 6. However, in the neighborhood of and beyond the grid the electron velocity is ordinarily large enough so that space charge is negligible. Therefore the local grid-region potential distribution follows the pattern discussed in connection with Fig. 4. Thus the 4/3-power curve from the cathode merges into the grid-region situation in just the way the straight line from the cathode does in Fig. 4(b). This new complete-space-charge potential structure is illustrated in Fig. 7. Note that the line that is tangent to the 4/3-power curve at the grid plane intersects the zero potential line three fourths of the distance from grid plane to cathode.

It appears then, that a result of the presence of complete space charge is to make the direct-current potential distribution shape up as though the cathode were nearer to the grid than is actually the case, in the ratio of 3 to 4. This is, of course, the equivalent of increasing the capacitance C_{cG}, Fig. 1(b), in the ratio 4 to 3.

Thus in Fig. 1(b), under complete space-charge conditions, a new value E_G' of equivalent grid-plane potential is obtained by using a new value C_{cG}' in place of C_{cG}. Mathematically,

$$C_{cG}' = \frac{1}{3\pi a}\left\{\begin{array}{l}\text{statfarads per square centimeter}\\ \text{complete space charge}\\ \text{parallel planes}\\ \text{direct-current capacitance}\end{array}\right\}. \quad (34)$$

[6] Presented by B. J. Thompson at the Electronics Institute lectures, University of Michigan, July, 1937.

[7] B. D. I. Tellegan, "The effect of the emission current in a triode," *Physica*, vol. 50, p. 301, 1925.

[8] F. B. Llewellyn, "Operation of ultra-high-frequency vacuum tubes," *Bell Sys. Tech. Jour.*, vol. 14, p. 632, 1935.

Neither C_{cG} nor C_{pG} are appreciably affected by space charge, because of the relatively high electron velocity at and beyond the grid.

So, to determine the complete-space-charge current density for a triode, one first obtains a new value E_G' of direct-current equivalent grid-plane potential by putting C_{cG}' for C_{cG} in (24); thus,

$$E_G' = \frac{E_g + \dfrac{E_p}{\mu}}{1 + \dfrac{1}{\mu} + \dfrac{1}{\mu}\dfrac{C_{cG}'}{C_{pG}}}. \tag{35}$$

Any definite value of cathode-current density J corresponds to some definite 4/3-power curve from the cathode. The value of J may be expressed in terms of potential and distance at any point along this curve, but the most convenient point is E_G' at distance a, Fig. 7. From the usual form of the 3/2-power law for parallel planes,

$$J = 2.33 \times 10^{-6}\frac{E_G'^{3/2}}{a^2} \tag{36}$$

where a is the cathode-to-grid spacing.

Thus (35) and (36) in combination determine the direct-current component of cathode current for a parallel-plane triode. The analysis can be extended to multigrid tubes by the obvious expedient of replacing C_1 in, for example, Fig. 3, by a capacitance C_1' evaluated for complete space charge.

If the point G in Fig. 1(b) and the points G and S in Fig. 3(b) represented actual physical electrodes, their direct-current potentials would be determined by the various leakage resistances rather than by the direct-current capacitance values. However, these points represent entirely fictitious electrodes. Therefore the "resistances" from them to the actual electrodes are infinite, so that E_G and E_S may properly be thought of as being controlled by the direct-current values of the capacitances.

If any particular grid's potential is low enough to produce a virtual cathode the effect of space charge in that locality should be considered. One effect will be to increase the direct-current grid-plane-to-grid-plane capacitance in the same manner that C_{cG} is increased by space charge. The capacitance C_{gG}' between such a grid and the plane in which it lies will be somewhat increased by the grid's being more or less surrounded by space charge, as can be seen by a study of the effect of space charge on the potential distribution in the plane of the grid.

VIII. The Effect of Space Charge in Cylindrical Geometry

Equation (35) applies to cylindrical as well as to parallel-plane geometry, provided the correct interpretation is given to the modified capacitance C_{cG}'.

Langmuir's solution[1] for current-potential relations for complete space charge between concentric cylinders can be put into the form of an equation for potential distribution, and a direct-current value of the complete-space-charge capacitance obtained in the same general manner as in the parallel-plane case.[5] The result is as follows:

$$C_{cG}' = \frac{1}{3}\left[1 + \frac{r/r_c}{\beta^2}\frac{d\beta^2}{d(r/r_c)}\right]\left\{\begin{array}{l}\text{statfarads per centimeter, complete}\\ \text{space charge, cylindrical geometry,}\\ \text{direct-current capacitance, at } r = r_g.\end{array}\right\}. \tag{37}$$

A graph of C_{cG}' appears in Fig. 8. For values of r_g/r_c larger than about 20 the error involved in assuming that $C_{cG}' = \frac{1}{3}$ (0.370 micromicrofarad per centimeter) is less than 5 per cent. This capacitance of $\frac{1}{3}$ statfarad per centimeter is that which corresponds to a value $r_g/r_c = 4.48$ in the space-charge-free condition.

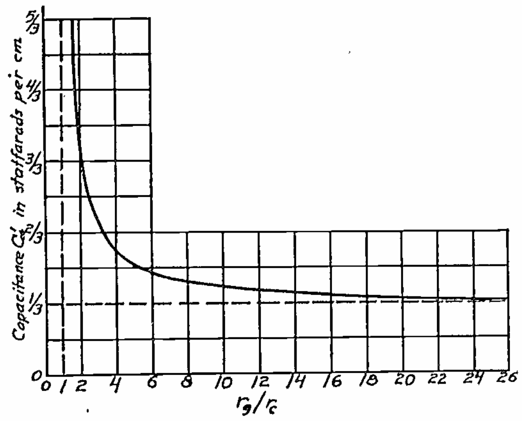


Fig. 8—Dependence on direct-current capacitance C_{cG}' (between cathode and grid cylinder) on r_g/r_c, for coaxial geometry under complete space-charge conditions.

The use of a suitable value of C_{cG}', obtained from Fig. 8, permits evaluation of the equivalent grid-plane potential in an ideal cylindrical tube. Values given by Fig. 8 are only valid in case the cathode or virtual cathode is smaller in diameter than the adjacent electrode. In case the cathode or virtual cathode is larger than the other electrode being considered, a different set of values, based on what Langmuir calls $-\beta^2$ is obtained. The dependence of current I per centimeter length of cathode is given by the usual β^2 equation,[5] as follows:

$$I = 2\pi \times 2.33 \times 10^{-6}\frac{E_G'^{3/2}}{r_g\beta^2}. \tag{38}$$

IX. Relationship of the Electrostatic Circuit to Interelectrode Capacitances

It is possible to determine, from the electrostatic-circuit diagram, the tube interelectrode capacitances, as far as they relate to the geometry of the electronically active portions of the electrode structure.

The general types of interrelations are given for a triode by (1) and (2), because the delta capacitances

C_{cg}, C_{pc}, and C_{gp} are in fact the interelectrode capacitances for a triode.

The general interelectrode-capacitance diagram for a pentode is illustrated in Fig. 9(a) and the corresponding electrostatic-circuit diagram in Fig. 9(b).

The arrangement of an alternating-current power source, voltmeter, ammeter, and set of connections shown dotted in Fig. 9 is designed to measure the reactance X_{cg1} due to C_{cg1}. Of course X_{cg1} is the ratio of source voltage to the ammeter current. No current other than that to C_{cg1} flows to electrode g_1, because g_2, g_3, and p are all held at the same potential as g_1.

Since the circuit of Fig. 9(b) does in fact represent all the important electrostatic relationships between the electrodes, any measurements on the circuit of Fig. 9(b) must give the same results as those on the circuit of Fig. 9(a). The results of such measurements as applied to the circuit of Fig. 9(b) are easily predicted (if C_1, C_2, C_3, etc., are known) by reduction of

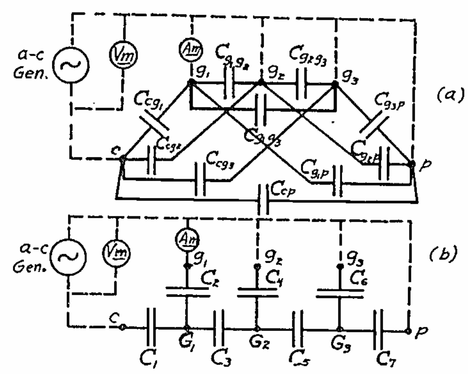


Fig. 9
(a) Interelectrode capacitances for a pentode, with circuit (dotted) for measuring C_{cg1}.
(b) Same measuring circuit as in (a), applied to the equivalent electrostatic circuit.

the circuit, using customary series-parallel methods. A similar method can be used for evaluating any of the other interelectrode capacitances in Fig. 9(a), if the C's of Fig. 9(b) are known.

In many devices the electrode shapes are not as regular as has been assumed. By judicious employment of combinations of equivalent electrostatic circuits in parallel, of flux-mapping and electrolytic-tank methods, of elliptical transformations, and various other graphical and mathematical devices, many types of tube geometry can be analyzed in terms of one or a set of equivalent electrostatic circuits. The behavior of converter and mixer tubes can be discussed very conveniently in terms of the general principles outlined above.

X. Comparisons with Other Methods of Analysis

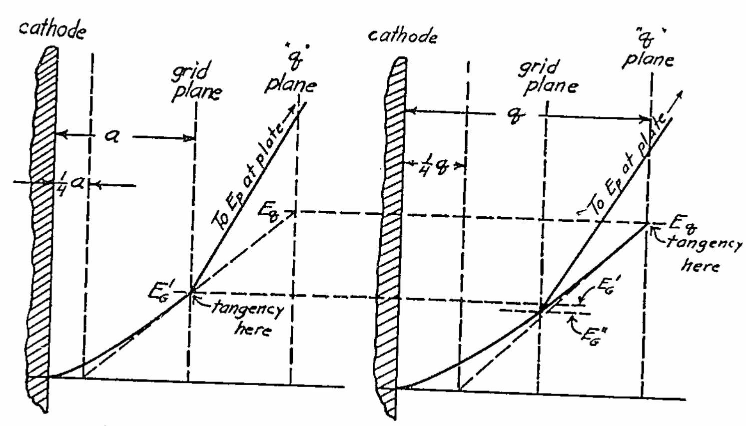
Fremlin[9] has recently described another method of relating triode geometry and potentials to cathode current. His procedure may be summarized as follows:

(a) It is assumed that cathode current is proportional to an equivalent voltage, $E_g+(E_p/u)$; this assumption has of course received wide *experimental* verification, and is expressed by (35) and (36) above.

(b) The proportionality factor involved is determined from the simple special case in which the actual grid potential lies on a 4/3-power potential-distribution curve extending from zero potential at the cathode to plate potential at the plate. Thus space charge behind the grid as well as in front of it is presumably accounted for.

Now if space charge between the grid and the plate is taken into account, the potential distribution between the grid and the plate, Fig. 7, ceases to be linear. Therefore C_{pG}, consequently also μ (related to C_{pG} by (22)), change from their space-charge-free values. The amount of the change depends on the current, because in this region *complete* space charge does not exist. Therefore μ is no longer a constant, but depends on the current. Thus Fremlin's analysis is inconsistent in that it ignores the effect on μ of space charge behind the grid in the basic (a) assumption, yet takes great pains to consider it in the subsequent (b) procedure. If a procedure analogous to Fremlin's is followed, except that space charge behind the grid is ignored, the results obtained agree with those given by the method used in the present paper.

Another method[3] of current prediction sometimes used consists in setting up an equivalent diode whose electrodes are the c and q terminals of Fig. 5(b), then determining the complete-space-charge current in that diode (the q-plane or i-plane diode) at potential E_q. The difference between the potential distributions assumed in the method using E_q and the one using E_G, (35) and (36), is illustrated in Fig. 10. The location and value of E_q is the same for both methods. In the E_G method, the 4/3-power potential curve is tangent, at the grid plane, to a straight line from the q point to a point at zero potential and distant $a/4$ from the cathode. In the E_q method, the 4/3-power potential curve is tangent, at the q plane, to a straight line from the q point to a point at zero potential distant $q/4$ from the cathode. The 4/3-power curve for the q method must cross the grid plane slightly below E_G, so that the E_q



(a) E_g method (b) E_q method
Fig. 10—Comparison of potential distributions for methods of current determination (plates omitted).

[9] J. H. Fremlin, "Calculation of Triode Constants," *Phil. Mag.*, vol. 27, no. 185, June, 1939.

method predicts a current slightly smaller than that predicted by the E_G method.

The Fremlin method, the E_q method, and the E_G method give numerical results that are, for customary geometries and potentials, closer to one another than any of them are to experimental values obtained from commercial tubes of suitable geometry. All involve some degree of approximation, in regard to the effect of space charge behind the grid. Therefore it seems to the writer that the choice between them should be based on simplicity and ease in understanding and use. From the instructional standpoint the E_G method is far simpler to present and more quickly grasped than either of the others. None of the methods are satisfactory if space charge behind the grid is important, as when E_g is high and either E_p low or grid-plate spacing large. Within the range of experimental verification of Fremlin's (a) assumption, all of them are satisfactory.

Fremlin,[9] also Glosius,[10], also Oertel[11] have recently analyzed the reduction in the value of μ at low plate currents, that becomes so prominent when the grid-cathode spacing is small. Fremlin's approach to this question, using infinite images, is essentially the same as one the present author has worked with, using finite Z-plane images, to give an equipotential cathode surface. The author's approach to that problem, however, would extend the E_G method by setting up an expression for an equivalent grid-plane potential which varies from point to point along the tube, thus permitting integration to determine the total current.

XI. Conclusions

The material presented above may be summarized as follows:

(a) A method of electric-field analysis based on conformal transformation has been used to demonstrate that the electrostatic properties of a triode may be represented by three capacitances in star, whose magnitudes are related to tube geometry in simple fashion.

(b) The simplicity of the dependence of the capacitance values on geometry permits immediate extension of the method to the construction of an equivalent electrostatic circuit for multigrid tubes.

(c) Good approximations to the potential distribution in various parts of tubes can be obtained directly from the electrostatic circuit, and more accurate determinations of local potential distribution can then be built around these first approximations.

(d) The electrostatic circuit permits setting up very simple and rational expressions for the dependence of equivalent grid-plane potential and cathode-current flow upon the various electrode potentials, for parallel-plane and cylindrical geometry.

(e) The electrostatic circuit can be directly employed to determine the ordinary interelectrode capacitances, as far as these are affected by the structure of the active portions of the electrodes.

(f) A comparison is made of the prediction of total cathode current by three methods, (1) using E_G, (2) using E_q, and (3) Fremlin's method.

Appendix I

Derivation of Triode Star Capacitances

If, in the star circuit, Figs. 1(b) and 2, there is no external circuit connection to the grid, there will, of course, be no charge on the grid. In the actual triode the grid wires, having no charge and occupying only a very small volume, will exert negligible effect on the capacitance between cathode and plate. The over-all cathode-to-plate capacitance will therefore be to all intents and purposes the same as if the grid were not there, that is, $1/4\pi(a+b)$ statfarads per square centimeter. But this must correspond to the over-all capacitance between c and p, Fig. 1(b), with the grid floating; therefore,

$$\frac{1}{\dfrac{1}{C_{cG}}+\dfrac{1}{C_{pG}}}=\frac{1}{4\pi(a+b)}. \tag{39}$$

Vogdes and Elder[12] showed by a method employing conformal transformation[5] that the geometric amplification factor[13] μ for a triode can be satisfactorily expressed by the following formula, if the screening fraction $2nR$ of the grid is not more than about 1/6:

$$\mu=\frac{2\pi nb-\log\cosh 2\pi nR}{\log\coth 2\pi nR}. \tag{40}$$

In the great majority of interesting cases, $\log\cosh 2\pi nR \ll 1$. Since $2\pi nb$ invariably is considerably larger than 1, the term $\log\cosh 2\pi nR$ can be neglected ordinarily. For example, if $2nR$ is 1/6, $\log\cosh 2\pi nR$ is about 0.12; even if the grid-plate spacing is only 1.5 times the grid-wire spacing (i.e., $nb=1.5$) the error introduced by neglecting $\log\cosh 2\pi nR$ in the numerator is slightly over 1 per cent, which makes it of no importance.

If the screening fraction is small, the factor $\log\coth 2\pi nR$ becomes approximately $\log 1/2\pi nR$. However, the error in μ introduced by making this approximation is a little over 10 per cent if the screening fraction is as large as the value 1/6 used in the preceding paragraph. Thus in a number of real cases this latter approximation leads to an appreciable error. The most generally useful formula for μ is, therefore,

[10] T. Glosius, "Calculation of the characteristics of triodes," *Hochfrequenz. und Elektroakustik*, vol. 52, pp. 88–93; September, 1938.

[11] L. Oertel, "On the theory of vacuum tubes in which the grid-cathode distance is small relative to the grid-wire spacing," *Die Telefunken-Rohre*, vol. 12, pp. 7–17; April, 1938.

[12] F. B. Vodges and F. R. Elder, "Formulas for the amplification constant for three-element tubes," *Phys. Rev.*, vol. 24, p. 683, 1924.

[13] F. Ollendorf, "Calculation of the amplification factor of narrow gratings," *Elektrotech. und Maschinenbau* (Vienna), no. 50, p. 585; December 16, 1934.

$$\mu = \frac{2\pi nb}{\log \coth 2\pi nR}. \tag{41}$$

These equations for μ are obtained[5,11] as the ratio $-E_p/E_g$ when the cathode charge is zero. But in that case

$$(E_p - E_G)C_{pG} + (E_g - E_G)C_{gG} = 0 \tag{42}$$

and $E_G = E_c = 0$. Therefore, with zero cathode charge

$$\frac{C_{gG}}{C_{pG}} = -\left.\frac{E_p}{E_g}\right|_{Q_c=0} = \mu. \tag{43}$$

That is, from (41) and (43)

$$\frac{C_{gG}}{C_{pG}} = \frac{2\pi nb}{\log \coth 2\pi nR}. \tag{44}$$

By similarly working with zero charge on the plate it can be shown that

$$\frac{C_{gG}}{C_{pG}} = \frac{2\pi na}{\log \coth 2\pi nR}. \tag{45}$$

Simultaneous solution of (39), (44), and (45) lead immediately to (3), (4), and (5) for the star capacitance values per square centimeter.

Appendix II

Potential-Distribution Equations

By adding and reducing (66), (67), and (68) in the author's book "Fundamentals of Engineering Electronics," the following general equation can be obtained for the potential at any point T, with coordinates x, y, in a triode:

$$E_T - E_c = -2\tau_c \log s_c s_g \sqrt{1 + \frac{1}{r'^2 s_g^2} - \frac{2}{r' s_g}\cos\phi'} + 2\tau_p \log \sqrt{1 + \frac{r'^2}{s_g^2} - \frac{2r'}{s_g}\cos\phi'} \tag{46}$$

where $(-\tau_c)$ is cathode charge per grid section, $+\tau_p$ plate charge per grid section, $s_g = \cosh 2\pi nR$, $s_c = \epsilon^{-2\pi na}$, $r' = \epsilon^{2\pi nx}$, and $\phi' = 2\pi ny$. The quantities n, R, and a are as indicated in Fig. 2. x and y are distances measured relative to an origin at grid-wire center. For a path midway between grid wires, $\phi' = \pi$; for a path through the center of a grid wire, $\phi' = 0$. For either of these two values, the quantities under the radicals become perfect squares, so that the entire expression reduces to the form of (14).

Equations (14) and (15), when converted into terms of x and the *slopes* $4\pi n\tau_c$ and $4\pi n\tau_p$ of the straight portions of the potential-distribution curves (Fig. 4(a)), take the following forms, applicable to parallel-plane geometry:

$$E_T - E_G = -4\pi n\tau_c\left[\frac{1}{2\pi n}\log\left|\epsilon^{-2\pi nx} \pm \cosh 2\pi nR\right|\right] + 4\pi n\tau_p\left[\frac{1}{2\pi n}\log\left|1 \pm \frac{\epsilon^{2\pi nx}}{\cosh 2\pi nR}\right|\right]. \tag{47}$$

The positive signs are used for the between-grid-wire path, negative signs for the through-grid-wire path. The potential is in statvolts and the τ's in statcoulombs per centimeter of length per grid section.

A special point whose potential is of interest is that in the grid plane, and midway between grid wires. To obtain a value of potential for this point, let $\phi' = \pi$ in (16). This gives, for the potential E_h at point h, where $x = 0$ and $y = 1/2n$.

$$E_h - E_G = -2\tau_c \log(1 + \cosh 2\pi nR) + 2\tau_p \log\left(1 + \frac{1}{\cosh 2\pi nR}\right). \tag{48}$$

The above equations can be converted directly into forms usable with cylindrical geometry by the following transforming relations:

$$\epsilon^{2\pi nx} \text{ becomes } \left(\frac{r}{r_g}\right)^{2\pi n r_g} \tag{49}$$

$$2\pi ny \text{ becomes } 2\pi n r_g \phi \tag{50}$$

where r_g is the grid cylinder radius; r is the radius and ϕ the angle, describing any point T in the true cylindrical geometry. The hyperbolic sine and cosine terms are unchanged. The charge per centimeter length of true cathode is $2\pi n r_g(-\tau_c)$ of true plate $2\pi n r_g(+\tau_p)$, for a triode, as used in (17) and (18).

Transient Response of Single-Sideband Systems*

HEINZ E. KALLMANN†, ASSOCIATE, I.R.E., AND ROLF E. SPENCER‡, NONMEMBER, I.R.E.

Summary—*The deformations, which typical television transients suffer in single-sideband systems are shown. The distortion is due to a spurious component added in quadrature to the genuine signal. Distortion increases with the depth of modulation and with the ratio of the signal band width to the width occupied by the cutting slope of the sideband-suppressing filter. It also depends on the shape of the cutting slope.*

THE accommodation of wide frequency bands is one of the fundamental problems of television. Thus, any scheme which would appear to reduce the frequency range required for a given picture definition calls for careful examination. Prominent among such schemes is the use of only a single sideband whenever the television signals appear as the modulation of a high-frequency carrier. The exact symmetry of the sidebands indicates that the same intelligence is transmitted on both sidebands, and thus one sideband alone should communicate it as unambiguously and completely as both. The saving by means of this scheme of approximately half the band width is self-evident, but it carries with it certain inherent drawbacks which are hidden from a superficial examination.

The deformation of various representative transients, as brought about by various single-sideband systems, has been calculated and the results are given in this paper. The analysis involved some difficult mathematical problems. We have C. P. Singer to thank for their solution, as presented in very condensed form in the paper immediately following this.

Definition of a Single-Sideband System

A single-sideband system is idealized in the following way:

(1) The original transient $E_0(T)$ is fed to a distortionless modulator. The carrier wave may be of infinitely high frequency, so that its phase has no effect on the shape of the resulting envelope. The depth of the modulation with a transient may be described by the ratio N, indicating the ratio of the rise in carrier amplitude to the carrier amplitude before the application of the transient. Doubling the carrier amplitude then corresponds to $N=1$, whereas any rise from zero carrier amplitude corresponds to $N=\infty$.

(2) The modulated carrier is fed to a high- or low-pass filter which may have no attenuation in its pass band and infinite attenuation outside its pass band. An infinitely sharp cutoff is not physically realizable; neither does the calculation lead to finite integrals. Thus, cutoff within a finite cut range is assumed, along either a straight or a curved amplitude response, extending over the frequency range from $-\omega_c$ to $+\omega_c$, where the cut frequency ω_c is the highest modulation frequency affected by the cut. It is stipulated, however, that this cutoff response is always symmetrical around the value $A=0.5$ and that this value is reached exactly at the carrier frequency. Thus, whatever the shape of the cutoff response from $-\omega_c$ to $+\omega_c$, the sum of the amplitudes of each two corresponding sideband frequencies, below as well as above ω_c, is always unity.

(3) It is further assumed that there is no delay, or phase, distortion in the filter for any transmitted frequency.

(4) The output of the filter is fed to an ideal rectifier, which, under the assumed ideal load conditions, will produce an output voltage exactly proportional to the envelope of the carrier. The comparison of this output $E(T)$ with the original transient $E_0(T)$ will then show the effects of an ideal single-sideband system.

General Description of Resulting Transients

The mathematical analysis shows that whenever a transient $E_0(T)$ is applied to a single-sideband system, the resulting transient has the mathematical form

$$E(T) = \sqrt{E_0(T)^2 + F(T)^2};$$

that is, the original transient plus a fault amplitude $F(T)$ added in quadrature. $E_0(T)$ is the envelope of the original carrier. $F(T)$ is the envelope of an additional carrier of the same frequency which is displaced 90 degrees in phase towards the bulk of the one remaining sideband, and whose amplitude is the greater the higher the modulation frequencies included and the stronger the sideband amplitude are. In a double-sideband system this is counteracted by an identical additional carrier displaced 90 degrees in the opposite direction and thus completely canceled. A single-sideband system has lost this balance and thus exhibits a carrier phase moving with and towards its modulation frequencies. Moreover, for a symmetrical transient, like the unit step or the error integral *erf*, $F(T)$ is symmetrical around a peak at $T=0$ and decays gradually to zero for $T=\pm\infty$. Further, as $F(T)^2$ is always positive, the resulting transient, if plotted on a common time scale, appears to be always above the original transient.

The fault $F(T)$ depends of course on the shape of $E_0(T)$ of the original transient. Furthermore its com-

* Decimal classification: R583×R410 Original manuscript received by the Institute, March 28, 1940; revised manuscript received, September 16, 1940. Presented, Fourteenth Annual Convention, New York, N. Y., September 23, 1939. A Mathematical Appendix to this paper by Charles P. Singer immediately follows.

† Formerly Electrical and Musical Industries, Ltd., Hayes, Middlesex, England; now, Scophony Television, Ltd., New York, N. Y.

‡ Electrical and Musical Industries, Ltd., Hayes, Middlesex, England.

parative amplitude rises with N, the depth of modulation. And finally, its shape and duration depends very much on the steepness of cutoff of the single-sideband filter.

Unit-Step-Transient: Straight Cut of Amplitude Response

Within the limits given above, the unit-step transient is mathematically the simplest and the most general to apply to the single-sideband system. Also the simplest assumption of a finite cutoff is a straight line from $-\omega_c$ to $+\omega_c$ being 0.5 at the carrier frequency, as shown in the inset to Fig. 1. If the filter has no time

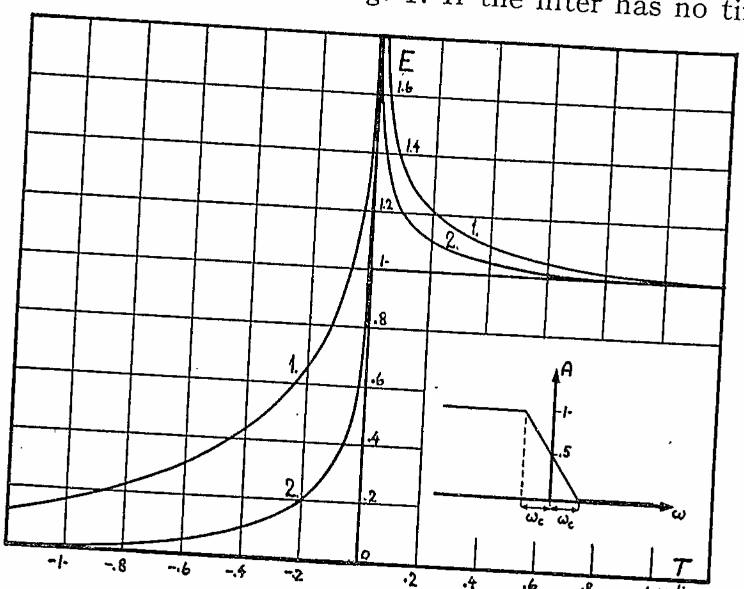


Fig. 1

delay, the unit step cannot give rise to any action before $T=0$, the time of arrival of the transient. Any practical filter would however give delay, which would allow faults to appear before as well as after the arrival of the undistorted signal, and to obtain complete results the unit step has technically to be replaced, e.g. by the limiting case of the *erf* transition of infinite steepness; in this case, action before $T=0$ is possible because the function is continuous.

The general equation for this case is given in the following paper as (9). Curve (1) in Fig. 1, equations (9a), (9b), show the transient-response curve for $N=\infty$, which corresponds to a start from zero carrier level (100 per cent modulation). Curve (2) corresponds to $N=1$, that is doubling the carrier amplitude (as from 33 to 66 per cent peak level or "33 per cent modulation," equations (9c) and (9d)). The distortion is of similar type, but somewhat less pronounced. The abscissa of Fig. 1 represents the time axis, its units being radians of the cut frequency ω_c. The ordinate of this and the following figures are plotted in units of the applied transient, always in the same scale regardless of the depth of modulation N. Thus for $N=1$ the value $E=0$ may correspond to 50 per cent peak level and the value $E=1$ to twice that value, viz., 100 per cent peak level.

The main feature of Fig. 1 is the fault near $T=0$, which lifts the transients to a sharp-peaked overswing, which indeed for an ideally steep transient would be infinitely high. Although the fault is the same for positive and negative values of T, its lifting effect is larger at the lower level due to the addition in quadrature, since $F>\sqrt{E^2+F^2}-E$.

That the output transient of the single-sideband system seems to have attained substantial values even before the time $T=0$ when the input transient was applied is explained by the simplifying assumption that the single-sideband system has no time delay. In "reality", an infinitely long filter is required, with accordingly infinite time delay, to produce exactly the straight cut as indicated in the inset figure in Fig. 1.

Unit-Step Transient: Parabolic Cut of Amplitude Response

It seemed important to establish to what extent such transition curves are influenced by the shape of the filter cutoff response. Instead of the sharp-cornered straight cut another was substituted with very pronounced rounded corners. The inset in Fig. 2 shows a cut composed of two parabolas, with their origins joined at the value $A=0.5$ as a point of inflexion. The general equation for this shape of cutoff is given in (11). Fig. 2, equations (11a), (11b), (11c), (11d), shows

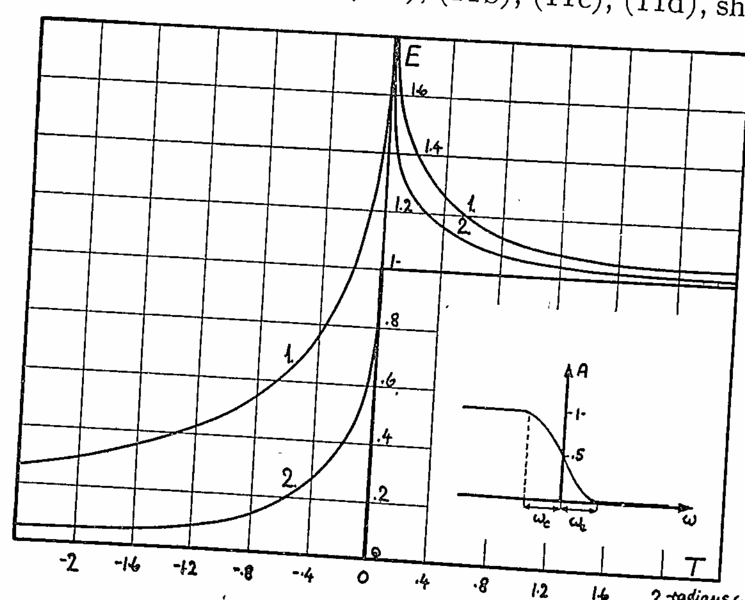


Fig. 2

again the cases of $N=\infty$ (100 per cent modulation) and of $N=1$ (33 per cent modulation). The scale of the abscissa is the same, and the type of transient is very similar to that of Fig. 1. The ordinate is again in time units of radians ω_c, but is compressed by the factor 2 compared with Fig. 1. Our conclusion was that the parabolic cut produces a similar type of distortion to that of a straight cut, much worse for comparable cut width, but not so substantially different as to warrant further investigation of different shapes of cutoff.[1] For further calculation a straight cut was again taken as a standard.

[1] Yet H. A. Wheeler and J. C. Wilson, in a paper presented after this paper was completed at the Annual Convention of the I.R.E. in Boston, June 29, 1940, have shown that reducing the filter slope near the carrier frequency to zero results in a more desirable shape.

erf Transition: Straight Cut of Amplitude Response

No lucid interpretation can be based on transient-response curves with infinite overswing, which is clearly caused by the idealized assumption of an infinitely steep transient, the unit step. This then had to be abandoned in favor of a transient of finite steepness.

Many simple shapes offer themselves, but to avoid complications all unsymmetrical transients were excluded, as well as transients with overswing and oscillation. Furthermore, such other transients as the straight rise and the sinusoidal transition, which have an oscillatory frequency response, requiring negative values of gain A, do not seem representative of practical cases. Thus the smooth *erf* transition was selected. It is so called because it embodies the error function integral[2] *erf*. Its shape is that of the curves K=0 in Figs. 3 and 4 (equation (16)). It has a nonoscillatory amplitude response $A=e^{-\omega^2/\omega_0^2}$, whose band width is described by the value ω_0, where the gain A has dropped to $1/e=0.37$. The highest useful frequency is, however, as explained elsewhere,[3] the much lower frequency $0.5\omega_0$, which is attenuated to 0.78.

The resulting transition was calculated on the basis of (16) and (17). Its shape depends largely on the ratio of "useful band width" $0.5\omega_0$ of the *erf* transition to "cut width" ω_c. This ratio $0.5\omega_0/\omega_c$ may be called K.

If now transient responses for different values of K are to be compared, we are less interested in a constant cut width ω_c with different band width $K\omega_c$ than in a

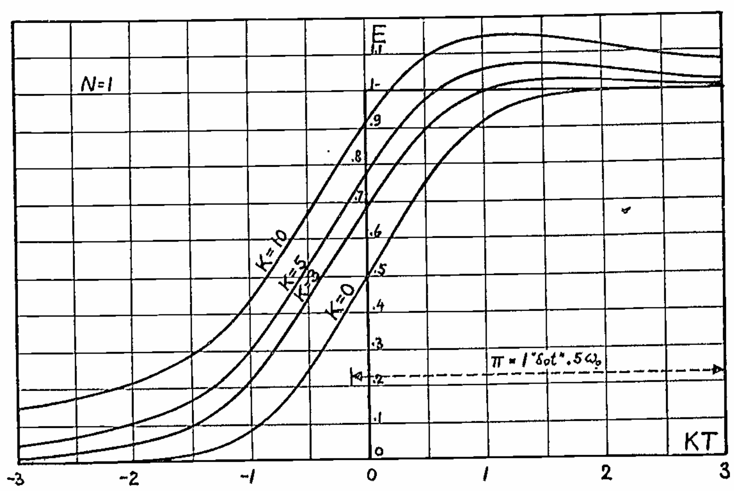


Fig. 3

constant *erf* transient with various cut widths ω_c. Thus the time axis of Figs. 3 and 4 was made KT instead of T, so that the same *erf* transient is always applied and only the cut width is varied. $K=0$ represents no cut and thus gives the original *erf* transient. In both Figs. 3 and 4 the duration of a "picture dot" is indicated assuming it to be one half cycle of $0.5\omega_0$ (i.e., π radians of $0.5\omega_0$).

The fault amplitude S of the *erf* transition, given in (16), is very tiresome to evaluate. This has however been done for values up to $KT=\pm 3$. For values of $K>5$ a very good approximation is given by (17), which however is no less tiresome for values of $KT>\pm 3$. Some values of the infinite series enclosed in brackets [. . .] in equation (17) are given in Table I of the paper which follows.

The results of the calculation for $N=1$ (33 per cent modulation) are given in Fig. 3, equation (16a). They show a general lift of the curve extending over two "picture dots," no sharp-peaked overswing and, apart from the long tail, no loss of steepness. Transients with $K=3$ and even $K=5$ show tolerably little distortion.

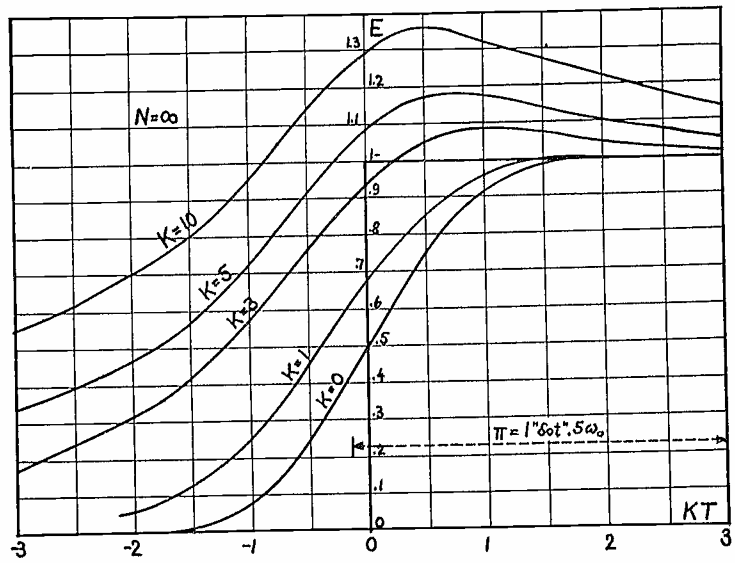


Fig. 4

But $K=10$ (10 per cent cut width) seems to be already useless. The same family of curves for $N=\infty$ (100 per cent modulation) is shown in Fig. 4, equation (16b). The distortion has become much more noticeable; even $K=3$ seems to be hardly useful. From the equations it appears to be possible, but not very probable, that the transients are slightly oscillatory for values $KT>\pm 3$. The case $K=1$ represents an amplifier with zero gain on one end of the total band and straight rise of gain to unity just on the other end of the total band. This gives tolerably little distortion indicating that a very uneven amplification of the two sidebands can be allowed in any carrier amplifier as long as there is no phase distortion introduced.

No separate calculations are needed for the case of a sudden drop in carrier amplitude, since in the cases of symmetrical transients the fault F is identical for $\pm T$. Thus such "negative transients" are represented by simply reading from right to left the curves in Figs. 1 to 4. $N=1$ then corresponds to halving the carrier amplitude, $N=\infty$ to switching off the carrier.

Conclusions

The cases discussed above, especially that of the *erf* transition, appear to be sufficiently representative of any practical application to allow some general conclusions. They indicate that single-sideband systems are suitable for general adoption only if special measures

[2] E. Jahnke and F. Emde, "Tables of Functions," B. G. Teubner, Leipzig, 1933, pp. 97–104.

[3] H. E. Kallmann, R. E. Spencer, and C. P. Singer, "Transient response in television," Proc. I.R.E. vol. 27, p. 613; September, 1939 (summary only).

are taken to reduce or avoid the distortion demonstrated above.

The first measure would be to keep the factor K always as small as possible, i.e., to cut with the least possible steepness. Evidently there is always an optimum compromise between allowable distortion and desirable saving of band width. Fig. 5 illustrates this by showing the resulting band width for various values of K. Full band width is required for $K<1$; the ideal saving of 50 per cent is attained only with $K=\infty$. But a value

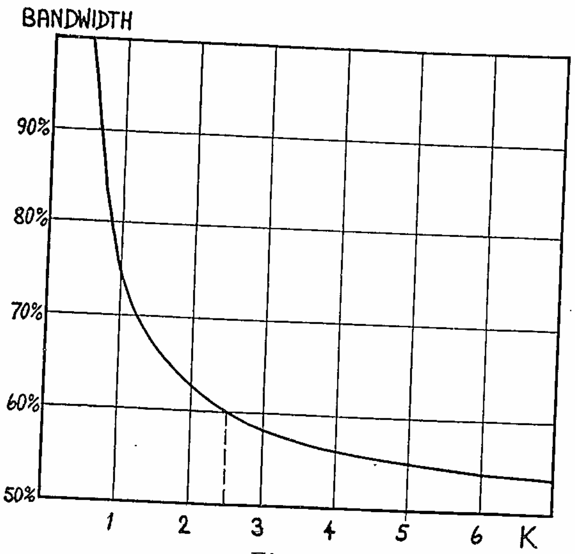


Fig. 5

of $K=2.5$ would reduce the band width to 60 per cent of the total without causing quite intolerable distortion.

The second measure would be to keep N, the depth of modulation, down to the lowest reasonable level. If for example a television system with positive modulation and negative synchronizing pulses is assumed, then a lift of the "black" level from 30 per cent white to 50 per cent white results in a value of $N_{max}=1$, Fig. 3, but requires only doubling (49/25) of the transmitter power for identical signal-to-noise ratio at the receiver. And this measure together with a value of $K=2.5$ would give 40 per cent saving of band width with tolerably little distortion.

To devise means of correcting the distortions due to single-sideband transmission appears to be rather hopeless; for example, subsequent phase correction might improve the shape of an "up" transition, but then the same transition "down" would be correspondingly worse, because a phase corrector shifts the balance between the beginning and the tailing of a transition, whereas the fault amplitude F due to single sidebands is always positive, whether the transition is "up" or "down." The fault is not simply a function of the amplitude or the frequency, but completely intertwined with both the shape of the transient and its height. For the same reason it must be concluded that phase distortion near the cut of the filter, which is to suppress the one sideband, will not offer any improvement. The main feature, that the fault amplitude is always positive, obviously remains and it must be assumed that phase distortion will result in deformation of the fault amplitude.

It is theoretically possible, however, to suppress the distortion completely in the rectification. It has been explained that the arriving signal consists of the original component $E_0(T)$ and the fault $F(T)$, added in quadrature. Any detector which is sensitive exclusively to the phase of the former, will be insensitive to the phase of the latter. Such a device is the multiplicative homodyne system, as exemplified by a hexode, to the one grid of which are fed the incoming signals, to the other grid of which are fed oscillations of exactly the genuine carrier frequency and exactly in phase with the $E_0(T)$ component of the incoming signal. This oscillation would have to be of extremely stable frequency and phase. It cannot be locked to the incoming signal as the $F(T)$ component would tend to detract it from its proper phase.

The fact that the fault amplitude due to single sidebands is always positive has some bearing on the choice of "positive" or "negative" modulation in television. Two opposing arguments may be brought forward, according to the degree of perfection aimed at. To minimize the fault, positive modulation with a black level not less than 50 per cent would be indicated, neglecting any possible deterioration of the synchronizing pulse near the zero level. However, if a slight fault due to single sidebands is to be tolerated, then it would perhaps be less conspicuous in systems with "negative" modulation; because the fault would tend to darken the edges of darker fields adjacent to brighter areas and this would somewhat counteract the effect of spreading of stray light, or stray electrons, from the brighter parts into the adjacent darker areas.

Note

Since this investigation was completed, a series of papers have appeared, all dealing with substantially the same problem. In all these papers, by S. Goldman,[4] R. Urtel,[5] H. Nyquist and K. W. Pfleger,[6] and R. D. Kell and G. L. Fredenhall,[7] the importance of the quadrature component is recognized, also generally the increase of distortion with increased depth of modulation and in some of them the importance of avoiding single-sideband filters with a steep cutoff near the carrier frequency. Thus, so far there is complete agreement between all investigators and in some of these papers the valuable observation is discussed that two unit dots separated by a unit gap (representative of two separate vertical lines in a picture) profit more

[4] S. Goldman, "Television detail and selective-sideband transmission," Proc. I.R.E. vol. 27, pp. 725–732; November, 1939.

[5] R. Urtel, "Observations regarding single-sideband transmission in television," *Telefunken Hausmitteilungen*, vol. 20, pp. 80–83; July, 1939.

[6] H. Nyquist and K. W. Pfleger, "Effect of the quadrature component in single sideband transmission," *Bell Sys. Tech. Jour.*, vol. 19, pp. 63–73; January, 1940.

[7] R. D. Kell and G. L. Fredenhall, "Selective side-band transmission in television," *RCA Rev.*, vol. 4, pp. 425–440; April, 1940.

from the introduction of a single-sideband system than the study of a single unit step suggests.

Yet confronted with an infinitely high transient response, when a unit step is fed to an ideal single-sideband system, as shown in Fig. 1, all of these investigators have chosen to limit the pass band of the system by means of what amounts to an ideal low-pass filter, with infinitely sharp cutoff. As a result, all the transients plotted show a slowly decaying oscillatory overswing which is due to the shape of cutoff of this assumed low-pass filter and not due to the single-sideband system proper. The authors of this paper, however, believe it to be important to avoid all but a very small percentage of overswing since its evidence is annoying in all pictures with satisfactory gradation. Thus, in spite of the mathematical difficulties, an *erf* transient was chosen, rather than the Si function with 9 per cent overswing followed by 4 per cent underswing, since the *erf* function has no overswing and thus is a very suitable representative of a well-shaped television signal. Consequently in the curves so obtained, Figs. 3 and 4, all overswing is solely due to the single-sideband distortion and it is hoped that these curves may be helpful in a decision, just how much depth of modulation and steepness of single-sideband-filter cutoff can be permitted without causing intolerable overswing.

A Mathematical Appendix to Transient Response of Single-Sideband Systems*

CHARLES P. SINGER†, NONMEMBER, I.R.E.

LET a sinusoidal carrier voltage $v(t)=\cos \nu t$ be modulated with the Heaviside unit function, defined by the requirements

$$H(t) = 0 \quad \text{for} \quad t < 0$$
$$H(t) = 1 \quad \text{for} \quad t > 0.$$

Then $H(t)$ may be represented by the Fourier integral

$$H(t) = \tfrac{1}{2} + 1/\pi \int_0^\infty \frac{\sin \omega t}{\omega} d\omega. \tag{1}$$

Also let the depth of modulation be $N \cdot 100$ per cent; then the resulting voltage as a function of time is

$$E(t) = \cos \nu t[1 + N \cdot H(t)] \tag{2}$$

$$= \left(1+\frac{N}{2}\right)\cos \nu t + \frac{N}{2\pi}\left[\int_0^\infty \frac{\sin (\omega+\nu)t}{\omega} + \int_0^\infty \frac{\sin(\omega-\nu)t}{\omega}\right]d\omega \tag{3}$$

or

$$E(t) = \cos \nu t \left\{1 + \frac{N}{2} + \frac{N}{\pi}\int_0^\infty \frac{\sin \omega t}{\omega}\cdot d\omega.\right\} \tag{4}$$

Thus, since

$$\int_0^\infty \frac{\sin \omega t}{\omega}\cdot d\omega = \begin{cases} 0 & \text{when} \quad t = 0 \\ \pi/2 & \text{when} \quad t > 0 \end{cases}$$

it follows that, when retaining both sidebands, the resulting response is undistorted.

Now let the device be such that one sideband is eliminated. From (1) it is first clear that the modulation frequencies increase to infinity, but on the other hand the band width is limited to the carrier frequency and we must therefore make the latter large compared with the cutoff frequency; for the same reason it is unnecessary to include a phase constant in the carrier voltage.

Eliminating the upper sideband, we have from (3)

$$E(t) = \left(1+\frac{N}{2}\right)\cos \nu t + \frac{N}{2\pi}\int_0^\infty \frac{\sin(\omega-\nu)t}{\omega}\cdot d\omega$$

or

$$E(t) = \cos \nu t\left\{1+\frac{N}{2}+\frac{N}{2\pi}\int_0^\infty \frac{\sin \omega t}{\omega}d\omega\right\} - \sin \nu t\left\{\frac{N}{2\pi}\int_0^\infty \frac{\cos \omega t}{\omega}\cdot d\omega\right\} \tag{5}$$

where

$$\int_0^\infty \frac{\cos \omega t}{\omega} d\omega \rightarrow \infty.$$

The trouble around the carrier frequency (i.e., for $\omega=0$) is due to the assumed frequency characteristic of the single-sideband filter, that is one cutting abruptly at the carrier, thus causing $E(t)$ to rise to infinity. We must therefore stipulate a gradual filter cutoff.

Unit Step: Straight Cut

The amplitude response $A(\omega)$ of the first single-sideband filter investigated is shown by the inset figure in Fig. 1. The transition curve is symmetric about the point $A=0.5$, $\omega=0$. Equation (3) then becomes

* Decimal classification: R410. Original manuscript received by the Institute, March 28, 1940; revised manuscript received, September 16, 1940.

† Electrical and Musical Industries, Ltd., Hayes, Middlesex, England.

$$E(t)=\tfrac{1}{2}\left(1+\frac{N}{2}\right)\cos\nu t+\frac{N}{2\pi}\left\{\int_0^{\omega_c}A(\omega)\frac{\sin(\omega+\nu)t}{\omega}\cdot d\omega\right.$$
$$+\int_0^{\omega_c}A(-\omega)\frac{\sin(\omega-\nu)t}{\omega}d\omega \qquad (6)$$
$$\left.+\int_{\omega_c}^{\infty}\frac{\sin(\omega-\nu)t}{\omega}d\omega\right\}$$

and in the case that the transition curve is a straight line

$$A(\omega)=\tfrac{1}{2}(1-\omega/\omega_c).$$

Equation (6) leads to

$$2E(t)=\cos\nu t\left\{1+\frac{N}{2}+\frac{N}{\pi}\int_0^{\infty}\frac{\sin\omega t}{\omega}\cdot d\omega\right\}$$
$$-\sin\nu t\left\{\frac{N}{\pi}\left[\frac{1}{\omega_c}\int_0^{\omega_c}\cos\omega t+\int_{\omega_c}^{\infty}\frac{\cos\omega t}{\omega}\right]d\omega.\right\} \qquad (7)$$

This expression eventually reduces to

$$E(t)=\left(\frac{1+N}{2}\right)\cos\nu t-\sin\nu t\left[\frac{N}{2\pi}\left\{\frac{\sin T}{T}-Ci(T)\right\}\right] \qquad (8)$$

where Ci is the cosine integral.[1]

The envelope of this signal is then found to be of the shape

$$2\,|\,E(T)\,|=\sqrt{(1+N)^2+\frac{N^2}{\pi^2}\left[\frac{\sin T}{T}-Ci(T)\right]^2}. \qquad (9)$$

The curves of Fig. 1‡ were plotted from this equation, the left half of the curve 1 ($N=\infty$) from (9a), its right half from (9b); also the left half of curve 2 ($N=1$) from (9c) and its right half from (9d).

$$E(T)=\sqrt{1/\pi^2\left[\frac{\sin T}{T}-Ci(T)\right]^2} \qquad (9a)$$
$$E(T)=\sqrt{1+1/\pi^2\left[\frac{\sin T}{T}-Ci(T)\right]^2} \qquad (9b)$$
$$E(T)=\sqrt{1+1/\pi^2\left[\frac{\sin T}{T}-Ci(T)\right]^2}-1 \qquad (9c)$$
$$E(T)=\sqrt{4+1/\pi^2\left[\frac{\sin T}{T}-Ci(T)\right]^2}-1 \qquad (9d)$$

For $T=0$, $|E(T)|\to\infty$ and for $T\to\infty$, $|E(T)|=1+N/2$. The infinite amplitude for $T=0$ is due to the infinitely steep slope of $H(t)$ for $t=0$, and similar remarks will apply to any other time function having an infinite slope at the point $t=0$.

If, instead of eliminating the upper sideband, we eliminate the lower one, then in (8) $\sin\nu t$ will have a positive sign instead of a negative, while the envelope remains the same in both cases.

[1] E. Jahnke and F. Emde, "Tables of Function" B. G. Teubner, Leipzig, 1933, pp. 79–86.

‡ Figures will be found in the preceding paper to which this is an Appendix.

Unit Function: Parabolic Cut

Next we shall consider a parabolic cut $A(\omega)$, which, as shown in the inset figure in Fig. 2, has a steeper slope at the carrier frequency. We have

$$A(\omega)=\tfrac{1}{2}(1-\sqrt{\omega/\omega_c}) \qquad (10)$$

and inserting this in (6) and evaluating by means of Fresnel functions, defined as follows:[1]

$$C(z)-jS(z)=\int_0^z\frac{e^{-ju}}{\sqrt{2\pi u}}du$$

we find the envelope

$$2\,|\,E(T)\,|=\sqrt{(1+N)^2+N^2/\pi^2\left[\sqrt{2\pi/T}\,C(T)-Ci(T)\right]^2}. \qquad (11)$$

The curves of Fig. 2 were plotted from this equation, using for the left half of curve 1 ($N=\infty$) equation (11a), for its right half, (11b); also for the left half of curve 2 ($N=1$) equation (11c) and for its right half, (11d).

$$E(T)=\sqrt{1/\pi^2\left[\sqrt{2\pi/T}\cdot C(T)-Ci(T)\right]^2} \qquad (11a)$$
$$E(T)=\sqrt{1+1/\pi^2\left[\sqrt{2\pi/T}\cdot C(T)-Ci(T)\right]^2} \qquad (11b)$$
$$E(T)=\sqrt{1+1/\pi^2\left[\sqrt{2\pi/T}\cdot C(T)-Ci(T)\right]^2}-1 \qquad (11c)$$
$$E(T)=\sqrt{4+1/\pi^2\left[\sqrt{2\pi/T}\cdot C(T)-Ci(T)\right]^2}-1 \qquad (11d)$$

Error-Integral Function: Straight-Line Cut

Lastly, we shall replace $H(t)$ by a continuous function

$$a(T)=\tfrac{1}{2}+1/\sqrt{\pi}\int_0^{KT}e^{-x^2}dx=\tfrac{1}{2}(1+erf\,KT) \qquad (12)$$

where

$$K=\tfrac{1}{2}(\omega_0/\omega_c)$$

and

$$T=\omega_c t.$$

Now let $y=\omega/\omega_c$, then, as shown elsewhere,[2] we may write (12) in the form

$$a(T)=\tfrac{1}{2}+1/\pi\int_0^{\infty}e^{-(y/2K)^2}\cdot\frac{\sin yT}{y}dy \qquad (13)$$

which agrees with (1), except for the decay factor under the integral.

But the response function corresponding to (1) with a straight-line cut is given by (7), and so the response function corresponding to (16) with a straight-line cut is given by

[1] Loc. cit., p. 108.

[2] H. E. Kallmann, R. E. Spencer, and C. P. Singer, "Transient response in television," Proc. I.R.E., vol. 27, p. 613; September, 1939 (summary only).

$$2E(t) = \cos \nu t\left\{1 + \frac{N}{2} + \frac{N}{2}\, erf KT\right\} - \sin \nu t \cdot \frac{N}{\pi}\left\{\frac{dI_1}{dT} + I_2\right\} \tag{14}$$

where

$$\left.\begin{aligned} I_1 &= \int_0^1 e^{-(y/2K)^2}\, \frac{\sin yT}{y}\, dy \\ I_2 &= \int_1^\infty e^{-(y/2K)^2}\, \frac{\cos yT}{y}\, dy \end{aligned}\right\}. \tag{15}$$

After evaluation of these integrals the envelope of (14) is found to be of the shape[1]

$$2\,|\,E(T)\,| = \sqrt{\{1 + N/2(1 + erf KT)\}^2 + \left(\frac{NS}{\pi}\right)^2}, \tag{16}$$

where

$$S = K\sqrt{\pi}\cdot e^{-(KT)^2}\cdot erf\,\frac{1}{2K} - \frac{1}{2}\, Ei\left(-\frac{1}{4K^2}\right)$$
$$- e^{-1/4K^2}\{A_4(KT)^4 - A_6(KT)^6 + A_8(KT)^8 - \cdots\}$$

and

$$A_4 = \frac{1}{3}\cdot\frac{1}{2!}(3-2) = \frac{1}{3!}$$

$$A_6 = \frac{1}{5\cdot 3}\cdot\frac{1}{3!}\left[\frac{1}{2K^2} + 5\cdot 3 - 4\cdot 2\right]$$

$$A_8 = \frac{1}{7\cdot 5\cdot 3}\cdot\frac{1}{4!}\left[\frac{1}{(2K^2)^2} + \frac{7\cdot 5 - 6\cdot 4}{2K^2} + 7\cdot 5\cdot 3 - 6\cdot 4\cdot 2\right]$$

$$A_{10} = \frac{1}{9\cdot 7\cdot 5\cdot 3}\cdot\frac{1}{5!}\left[\frac{1}{(2K^2)^3} + \frac{9\cdot 7 - 8\cdot 6}{(2K^2)^2} + \frac{9\cdot 7\cdot 5 - 8\cdot 6\cdot 4}{2K^2} + 9\cdot 7\cdot 5\cdot 3 - 8\cdot 6\cdot 4\cdot 2\right]$$

[1] Loc. cit. For the integral Ei, pp. 79–80.

The curves in Fig. 3 ($N=1$) were plotted from (16a) and those in Fig. 4 ($N=\infty$) from (16b).

$$|\,E(T)\,| = \sqrt{\{1 + \tfrac{1}{2}(1 + erf KT)\}^2 + S^2/\pi^2} \tag{16a}$$

$$|\,E(T)\,| = \sqrt{\{\tfrac{1}{2}(1 + erf KT)\}^2 + S^2/\pi^2}. \tag{16b}$$

For large K's ($K > 5$) only the last terms in each coefficient A_n are significant figures and the very good approximation equation (17) may be used.

$$S \approx \left\{K\sqrt{\pi}\cdot erf\,\frac{1}{2K}\cdot e^{-(KT)^2} - \frac{1}{2}\, Ei\left(-\frac{1}{4K^2}\right)\right.$$
$$- e^{-1/4K^2}\cdot\left[\frac{1}{3}\,\frac{(KT)^4}{2!} - \frac{(KT)^6}{3!}\left(1 - \frac{2\cdot 4}{3\cdot 5}\right)\right.$$
$$\left. + \frac{(KT)^8}{4!}\left(1 - \frac{2\cdot 4\cdot 6}{3\cdot 5\cdot 7}\right) - \cdots\right]. \tag{17}$$

When $K \to \infty$, the decay factor $e^{-(y/2K)^2}$ becomes unity, and the envelope is thus identical with (9) corresponding to the unit function, except that it is now also defined for negative values of the argument, since the *erf* function is a continuous function.

TABLE I

SOME VALUES OF THE INFINITE SERIES ENCLOSED IN [ ] OF EQUATION (17)

$\pm KT$	[ ]
0	0
0.25	0
0.5	+.009
0.75	+.041
1−	+.106
1.5	+.356
2−	+.58
2.5	≈+.8
3−	≈+1−

Linear Plate Modulation of Triode Radio-Frequency Amplifiers*

CHAO-YING MENG†, NONMEMBER, I.R.E.

Summary—*An analysis of the conditions to produce linear complete plate modulation of triode radio-frequency amplifiers shows that the grid-excitation voltage and bias potential of triode amplifiers have to be modulated together with the plate-supply potential. Besides giving linear modulation characteristics, the reflected impedance of the radio-frequency amplifier to the modulator is a constant resistance in the present scheme and excessive grid dissipation at the trough of modulation is eliminated. Calculations using the point-to-point method on different triodes and experimental measurements made on the triode 801 agree very well with the theory.*

INTRODUCTION

IT IS well known that oscillators have very good plate-modulation characteristics though it is not possible to obtain complete modulation with it. Due to the accompanying frequency modulation and limited power output, oscillators are now almost universally replaced by class C amplifiers. Better modulation characteristics can be obtained when a grid-leak bias is used than with a fixed bias. But there is the danger of damaging the tube on losing excitation, so a combination of fixed and grid-leak bias is generally used. The grid excitation is usually constant. When properly adjusted it is possible to obtain modulation characteristics with distortions of only a few per cent. There are, however, the defects of excessive grid dissipation at the trough of modulation, decrease of reflected impedance to the modulator and decrease of plate-circuit efficiency at the peak of modulation, localized plate heating due to "focusing effect," and

* Decimal classification: R135×R355.8. Original manuscript received by the Institute, October 29, 1940.

†Radio Research Institute, National Tsinghua University, Kunming, China.

others. To obviate these difficulties the amplifier is usually designed so that the full capacity of the tube is not utilized and the percentage of modulation is kept low in high-quality stations. It is the purpose of this paper to obtain analytically the conditions to produce distortionless modulation without these defects and verify the same by point-to-point calculations and experimental measurements.

Part I. Theoretical Analysis

When the load of a radio-frequency amplifier is a constant impedance, a distortionless plate-modulation characteristic means that the radio-frequency potential difference across, or the radio-frequency current through, the load is a linear function of the plate-supply potential, and for complete modulation, directly proportional to it. The latter case may be expressed as follows:

$$\frac{E_p}{E_{p_0}} = \frac{I_p}{I_{p_0}} = \frac{E_b}{E_{b_0}}, \tag{1}$$

where E_p and I_p are the radio-frequency potential difference across, and current through, the load impedance, respectively, and E_b, the plate-supply potential. The subscript 0 denotes that the corresponding quantities are values under carrier conditions. We shall call this the radio-frequency condition.

As far as the modulated amplifier is concerned, fulfillment of the radio-frequency condition is sufficient for distortionless modulation. However, the changing direct plate current I_b of the modulated amplifier reflects a load on the modulator equal to $\Delta E_b/\Delta I_b$, where ΔI_b is the change of the direct plate current when the plate-supply potential is changed by ΔE_b. Unless this reflected load is a constant resistance, wave-form distortion will result in the modulator, because vacuum-tube modulators are not constant-voltage generators without internal resistance. For complete modulation, I_b necessarily reduces to zero when E_b is reduced to zero; therefore, we further require

$$\frac{I_b}{I_{b_0}} = \frac{E_b}{E_{b_0}}. \tag{2}$$

We shall call this the direct-current condition. Heising[1] pointed out the above conditions clearly as early as 1921, but it seems that no attention was paid to the direct-current condition later and perfect modulation characteristics are assumed when the radio-frequency condition alone is satisfied.

Oscillators fulfill the above conditions closely, but it is not possible to fulfill both conditions simultaneously in class C amplifiers, whether fixed or grid-leak bias is used. Plate-modulated amplifiers using fixed bias give less output at the peak of modulation than if the modulation were linear and distortions of 10 or 20 per cent are not uncommon. Grid-leak-biased amplifiers give better characteristics. The grid bias is decreased at the peak of modulation, especially in high-power tubes, in which secondary emission from the grid takes place; so the output there can be made equal to or even more than that of distortionless modulation. The combined effect of the increased plate-supply potential and decreased bias at the peak of modulation makes the plate-circuit efficiency low. The direct plate current and plate dissipation, therefore, are greatly increased. Appreciable plate current flows when the grid potential is negative, and the localized heating of the plate by the "focused" electron beams decreases the effective plate dissipating power of the tube. Also, the grid dissipation at the trough of modulation can be excessively large. Hence it is customary to design the amplifier so that only a fraction of the tube capacity is used and the percentage of modulation is made low.

Mouromtseff and Kozanowski[2] have shown that by modulating the excitation with the plate-supply potential, it is possible to "compensate" the modulation so that the radio-frequency condition is satisfied at the peak of modulation. The degree of modulation of the exciter they used was "generally much less than 100 per cent." No mention of the direct-current condition was made, but it is evident from their oscillograms that it is not satisfied. The varying resistance will produce wave-form distortion in the modulator even though the radio-frequency condition is satisfied.

We shall now proceed to analyze the plate-modulated amplifier and find the necessary operating conditions for distortionless complete modulation of triodes.

A. *Plate Current of a Triode as a Function of Grid and Plate Potentials*

After the usual assumptions have been made, the space current follows the well-known three-halves-power law

$$i_s = G\left(e_g + \frac{e_p}{\mu}\right)^{3/2}, \tag{3a}$$

where i_s is the space current; G, a constant of the triode; e_g and e_p, the grid and plate potentials, respectively; and μ, the amplification factor of the tube. i_s is different from zero only when the expression in the parenthesis is positive.

We are, however, interested in the plate and grid currents separately. The way the space current divides itself between the electrodes depends upon the geometrical configuration of the electrodes and the relative magnitudes of the plate and grid potentials. For any particular set of conditions in a tube, the plate current i_p may be expressed by the relation

$$i_p = G\left(e_g + \frac{e_p}{\mu}\right)^x, \tag{3b}$$

[1] R. A. Heising, "Modulation in radio telephony," Proc. I.R.E., vol. 9, pp. 305–352; August, 1921.

[2] I. E. Mouromtseff and H. N. Kozanowski, "Analysis of the operation of vacuum tubes as class C amplifiers," Proc. I.R.E., vol. 23, pp. 752–778; July, 1935.

An examination of the characteristic curves, for instance, the constant-plate-current curves, of typical triodes reveals that unless the plate potential is equal to or smaller than the grid potential, they are very nearly parallel straight lines. This shows that for considerable portions of the whole family of characteristic curves, G and x in the above equation are constants with respect to variations of e_g and e_p. But the values of G and x vary with the value of i_p. As far as a mathematical expression for the family of curves is concerned, we may consider G to be a constant and x varying with i_p. Calculations show that the value of x is substantially constant for all values of i_p except when the latter is extremely small. There, the contribution to whatever quantity may be involved is also very small, so we may consider x to be a constant also for the straight portions of the whole family of curves.

The constant-current curves begin to bend upwards when e_g becomes comparable to e_p. We may still consider G to be a constant and make x vary in order to make (3b) fit the actual courses of the curves. The value of x decreases as the ratio e_g to e_p increases. But calculations from the curves of typical triodes also show that unless the ratio becomes much larger than 1, the value of x varies only slightly.

If we make the usual assumptions that the grid and plate circuits of a class C amplifier consist of tuned circuits, that the load impedance in the plate circuit is a constant resistance, and that neutralization is perfect and transit time of the electrons may be neglected, we can express the plate current as a function of time t as follows:

$$i_p = G\left[\left(E_c + \frac{E_b}{\mu}\right) + \left(E_g - \frac{E_p}{\mu}\right)\cos\omega t\right]^x, \qquad (3c)$$

where E_c is the bias potential; E_b, the plate-supply potential; E_g, the amplitude of the grid-excitation potential; E_p, the amplitude of the radio-frequency potential across the load resistance; and ω, the angular velocity of the amplified frequency. The expression in the parenthesis may be considered as the effective potential, and plate current begins to flow when E_c+E_b/μ is equal to $(E_g-(E_p/\mu))\cos\omega t$, but opposite in sign. Designating this particular value of ωt as θ, we have

$$\cos\theta = -\frac{E_c + \dfrac{E_b}{\mu}}{E_g - \dfrac{E_p}{\mu}}. \qquad (4)$$

Substituting into (3c)

$$i_p = G\left(E_g - \frac{E_p}{\mu}\right)^x(\cos\omega t - \cos\theta)^x. \qquad (3d)$$

B. The Average Plate Current and the Amplitude of the Fundamental Component of the Radio-Frequency Current Through the Load Resistance

Integrating the plate current i_p over a radio-frequency cycle, we obtain the average plate current

$$I_b = \frac{1}{2\pi}\int_{-\theta}^{\theta} G\left(E_g - \frac{E_p}{\mu}\right)^x(\cos\omega t - \cos\theta)^x d(\omega t). \qquad (5a)$$

G is a constant and E_g and E_p also vary so little in a radio-frequency cycle that they may be considered constant in the integration. x is nearly constant for all values of plate current except when the latter is exceedingly small and when the minimum plate potential is smaller than the maximum grid potential. So for normal operating conditions x may be considered as constant in the integration without appreciable error. In any event it is possible to find a constant effective x_b which when used in the integration would give the same average plate current as when the varying x is used. Then

$$I_b = \frac{G}{2\pi}\left(E_g - \frac{E_p}{\mu}\right)^{x_b}\int_{-\theta}^{\theta}(\cos\omega t - \cos\theta)^{x_b}d(\omega t).$$

The integral is now a function of θ and x_b only. Let us designate it as $F_b(\theta)$; then

$$I_b = \frac{G}{2\pi}\left(E_g - \frac{E_p}{\mu}\right)^{x_b}F_b(\theta). \qquad (5b)$$

The amplitude of the fundamental component of the radio-frequency current through the load resistance is

$$I_p = \frac{G}{\pi}\int_{-\theta}^{\theta}\left(E_g - \frac{E_p}{\mu}\right)^x(\cos\omega t - \cos\theta)^x\cos\omega t\, d(\omega t). \qquad (6a)$$

Similarly we can substitute for the varying x a constant effective x_p, which when used in the integration would give the same I_p. Then (6a) becomes

$$I_p = \frac{G}{\pi}\left(E_g - \frac{E_p}{\mu}\right)^{x_p}\int_{-\theta}^{\theta}(\cos\omega t - \cos\theta)^{x_p}\cos\omega t\, d(\omega t).$$

Writing the integral which is a function of θ and x_p only as $F_p(\theta)$,

$$I_p = \frac{G}{\pi}\left(E_g - \frac{E_p}{\mu}\right)^{x_p}F_p(\theta). \qquad (6b)$$

C. Modulation To Be Applied to E_g and E_c in Order to Satisfy the Radio-Frequency and Direct-Current Conditions Simultaneously

The radio-frequency condition requires that

$$\frac{I_p}{I_{p_0}} = \frac{\left(E_g - \dfrac{E_p}{\mu}\right)^{x_p}F_p(\theta)}{\left(E_{g_0} - \dfrac{E_{p_0}}{\mu}\right)^{x_{p_0}}F_{p_0}(\theta)} = \frac{E_b}{E_{b_0}}, \qquad (7)$$

and the direct-current conditions requires that

$$\frac{I_b}{I_{b_0}} = \frac{\left(E_g - \dfrac{E_p}{\mu}\right)^{x_b}F_b(\theta)}{\left(E_{g_0} - \dfrac{E_{p_0}}{\mu}\right)^{x_{b_0}}F_{b_0}(\theta)} = \frac{E_b}{E_{b_0}}. \qquad (8)$$

The simultaneous fulfillment of the radio-frequency and direct-current conditions requires that

$$\frac{E_b}{E_{b_0}} = \frac{I_p}{I_{p_0}} = \frac{I_b}{I_{b_0}},$$

or

$$\frac{\left(E_g - \frac{E_p}{\mu}\right)^{x_p} F_p(\theta)}{\left(E_{g_0} - \frac{E_{p0}}{\mu}\right)^{x_{p0}} F_{p0}(\theta)} = \frac{\left(E_g - \frac{E_p}{\mu}\right)^{x_b} F_b(\theta)}{\left(E_{g_0} - \frac{E_{p0}}{\mu}\right)^{x_{b_0}} F_{b_0}(\theta)}. \quad (9)$$

When an amplifier is operated so that its dynamic characteristic curves never cut the static constant-current characteristic curves where they bend upwards, x_b, x_p, x_{b_0}, and x_{p_0} are equal to one another. Equation (9) then becomes

$$\frac{F_p(\theta)}{F_{p_0}(\theta)} = \frac{F_b(\theta)}{F_{b_0}(\theta)}. \quad (10)$$

$F_b(\theta)$ and $F_p(\theta)$ are quite different and are general functions of θ. The only way they can vary according to (10) when E_b is varied is when both functions remain unchanged, i.e., θ has to remain constant, or

$$\frac{E_c + \frac{E_b}{\mu}}{E_g - \frac{E_p}{\mu}} = \frac{E_{c_0} + \frac{E_{b_0}}{\mu}}{E_{g_0} - \frac{E_{p0}}{\mu}}. \quad (11)$$

Equation (7) then becomes

$$\frac{E_b}{E_{b_0}} = \left[\frac{E_g - \frac{E_p}{\mu}}{E_{g_0} - \frac{E_{p0}}{\mu}}\right]^x.$$

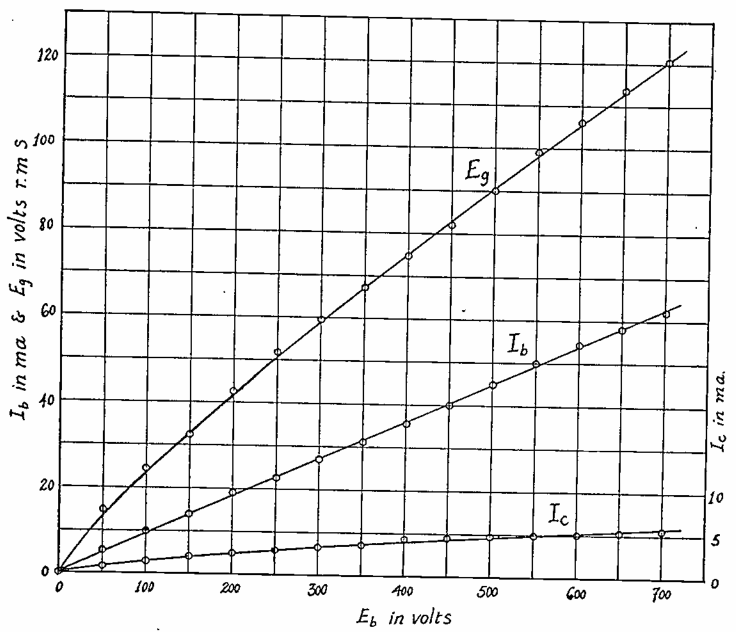


Fig. 1—E_g curve. Calculated values (solid curve) and measured values (small circles on the curve) of E_g to give distortionless modulation of triode 801. $E_c = -43.75$ volts at carrier and 100 per cent modulated with E_b. $E_p = 210$ volts root-mean-square at carrier.

Solving for E_g

$$E_g = \frac{E_p}{\mu} + \left(E_{g_0} - \frac{E_{p0}}{\mu}\right)\left(\frac{E_b}{E_{b_0}}\right)^{1/x}.$$

By (2)

$$E_g = E_{g_0} m^{1/x} + \frac{E_{p0}}{\mu}(m - m^{1/x}). \quad (12)$$

where $m \equiv E_b/E_{b_0}$. Substituting this value of E_g into (11) and solving for E_c, we obtain

$$-E_c = -E_{c_0} m^{1/x} + \frac{E_{b_0}}{\mu}(m - m^{1/x}). \quad (13)$$

Equations (12) and (13) give the necessary modulations to be applied to E_g and E_c, respectively, in order to satisfy the radio-frequency and direct-current conditions simultaneously. It is to be noted that they are exactly of the same character, for E_c and E_{c_0} are intrinsically negative quantities.

D. *Discussions of Equations (12) and (13)*

The form of the modulations to be applied to E_g and E_c is very similar to the modulation characteristic of conventional plate-modulated amplifiers. (Compare Fig. 1 with the E_p curves of Fig. 2 below.) The magnitude of the modulation depends on the value of x and the carrier operating parameters. The former depends on the construction of the tube and has a value ranging from 1.5 to 1.25. When the tube is operated as an ordinary class C amplifier under carrier conditions, the peak values of E_g and E_c required, when E_b is doubled, is about 85 to 90 per cent. Thus all the electrode potentials are nearly proportionally changed.

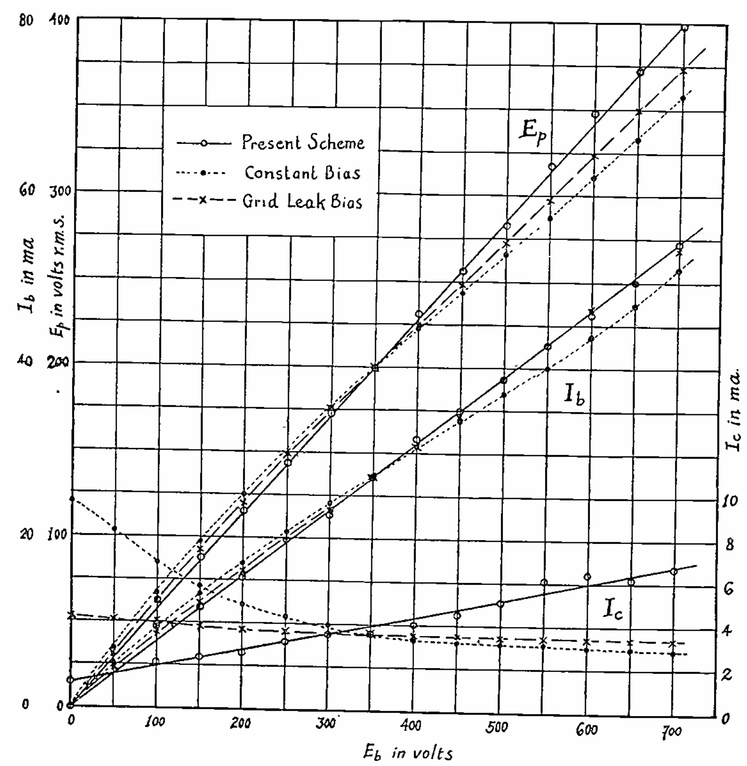


Fig. 2—Modulation characteristics of triode 801 under normal operating conditions. Carrier conditions: $E_{b_0} = 350$ volts; $E_{c_0} = -90$ volts; $E_{g_0} = 105$ volts root-mean-square. Solid curves: E_c is 100 per cent modulated and E_g is 90 per cent modulated with E_b.

The electric-field distribution between the electrodes at any particular phase angle in any two radio-frequency cycles is, therefore, also nearly the same. This means that $x_p = x_{p_0}$ and $x_b = x_{b_0}$ even when the amplifier is so loaded or excited that the dynamic characteristic curves do cut the static constant-current curves where they depart from their straight courses or when depressions in the plate-current pulse occur. It may seem that $x_p \neq x_b$ under such conditions due to the multiplying factor cos ωt in the integral of (6a). But close examination shows that where depression in plate-current pulse occurs, the multiplying factor cos ωt is approximately constant and very nearly equal to unity. Therefore (12) and (13) give the necessary modulations in order to satisfy the radio-frequency and direct-current conditions simultaneously even when the curved portion of the characteristic curves is traversed by the dynamic curves. It is thus shown that there is no restriction whatsoever on any of the operating parameters E_{c_0}, E_{g_0}, E_{b_0}, and E_{p_0}, or in other words, the tube may be operated even as a class B amplifier if not for considerations of power output, efficiency, etc., in contrast with the conventional plate-modulated amplifier whose grid bias has to be two or more times the cutoff bias and excitation enough to give voltage saturation of plate current.

If we set $E_{c_0} = -E_{b_0}/\mu$, (13) becomes

$$E_c = -\frac{E_b}{\mu}. \tag{14}$$

Thus the required modulation on E_c is linear and to the same degree as E_b. The bias is always maintained at

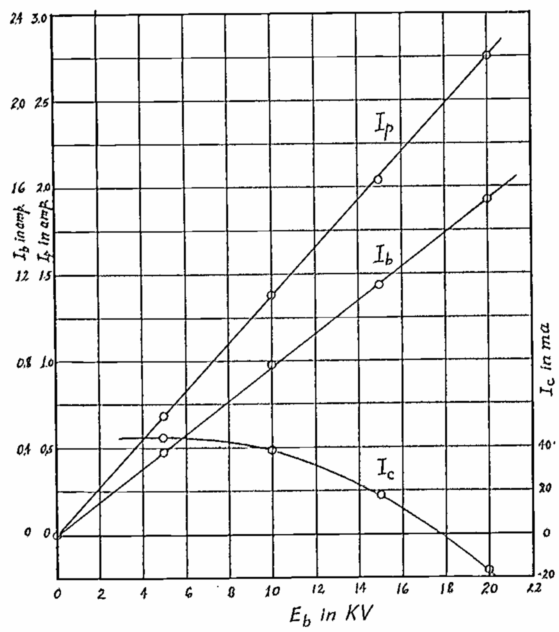


Fig. 3—Modulation characteristics of triode UV-207. Carrier conditions: $E_{b_0} = 10$ kilovolts; $E_{c_0} = -800$ volts; $E_{p_0} = 8.7$ kilovolts; $E_{g_0} = 1330$ volts. E_c is 100 per cent modulated and E_g is 85 per cent modulated at the peak of modulation.

the cutoff value and plate-circuit efficiency will be about the same as that of ordinary class B amplifiers, i.e., about 60 per cent. Operation under these conditions will not be very practical although it is one interesting possibility. The usefulness of this mode of operation lies in the fact that it offers a very convenient means for calculating x. In the general case the degrees of modulation on E_g and E_c are different. They have to be adjusted at the same time until both the radio-frequency and direct-current conditions are simultaneously satisfied. This will be a very long and difficult process. Making use of this special case, it is only necessary to try different values of E_g to give distortionless modulation. The value of x may then be calculated from (12).

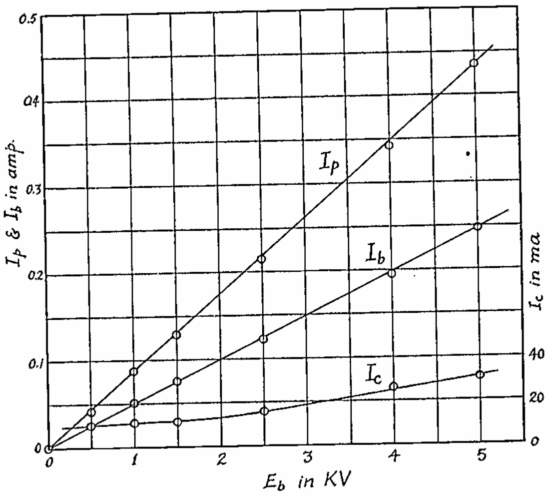


Fig. 4—Modulation characteristics of triode 100TH. Carrier conditions: $E_{b_0} = 2.5$ kilovolts; $E_{c_0} = -125$ volts; $E_{p_0} = 2.2$ kilovolts; $E_{g_0} = 234$ volts. E_c is 100 per cent modulated and E_g is 80 per cent modulated at the peak of modulation.

Part II. Point-to-Point Calculations

To check the above theory, point-to-point calculations were made on tubes of different power ratings and amplification factors. The carrier conditions were arbitrarily fixed, and use was made of the special case mentioned above. From the value of E_g required to give the correct values of I_p and I_b at m equal to 2, the value of x was calculated from (12). When this value of x was used, E_g for other values of m was calculated and compared with those obtained by the point-to-point calculations. Characteristic curves of several water-cooled tubes were given in Mouromtseff and Kozanowski's paper[2] and those of the triode 100TH were supplied by the manufacturers. Typical results are given in Table I.

TABLE I

Triode UV-207 Carrier Conditions: $E_{b_0} = 10$ kilovolts; $E_{c_0} = -600$ volts; $E_{p_0} = 9$ kilovolts; $E_{g_0} = 1130$ volts.			Triode 100TH Carrier Conditions: $E_{b_0} = 3000$ volts; $E_{c_0} = -75$ volts; $E_{p_0} = 2600$ volts; $E_{g_0} = 190$ volts.		
m	E_g calculated from (12)	E_g calculated by point-to-point method	m	E_g calculated from (12)	E_g calculated by point-to-point method
0.5	613	660	0.5	108	120
0.8	927	925	0.75	150	157
1.0	1130	1130	1.0	190	190
1.2	1329	1325	1.25	229	231
1.5	1620	1630	1.50	266	268
1.8	1907	1890	1.75	302	307
2.0	2090	2090	2.0	338	338

An actual plot of (12) or (13) will show that the required modulations of E_g and E_c resemble very closely the modulation characteristics of ordinary plate-modulated amplifiers, so the grid excitation can be easily arranged to satisfy (12). But in the following calculations we shall make the grid bias to be modulated proportional to E_b, and E_g linearly modulated so that the radio-frequency condition is satisfied at the peak of modulation. Modulation characteristics of the tubes 207 and 100TH are given in Figs. 3 and 4.

The plate current, output, etc., at the peak of modulation using the present method and those using constant bias, grid-leak bias, and a combination of the two are given in Table II.

TABLE II
Triode 100TH: Modulation Characteristics at Peak of Modulation
Carrier Conditons: $E_{b_0}=2500$ volts; $E_{c_0}=-125$ volts; $E_{p_0}=2200$ volts; $E_{g_0}=235$ volts.

Type of operation	I_p in milliamperes	I_b in milliamperes	Power input in watts	Power output in watts
Constant bias	370	228	1140	703
Grid-leak bias	390	245	1225	780
Combination bias*	392	246	1230	774
Present scheme	436	247	1235	968
Distortionless case	432	246	1206	952

* Fixed part of bias at carrier = −50 volts; bias produced by grid leak at carrier = −75 volts.

Part III. Experimental Measurements

A. *Experimental Check of Equations (12) and (13)*

The bridge method of Noyes[3] was used to measure the modulation characteristics of the triode 801. The circuit arrangement is essentially the same as that used by Noyes. Using a bias equal to the cutoff value, the grid-excitation voltage E_g required to give an output voltage E_p proportional to the plate-supply potential E_b was determined. The amplifier was loaded so that the minimum plate potential and maximum grid potential were approximately equal under carrier conditions. From the values of E_g at the carrier and at the peak of modulation, the value of x was calculated. Using this value of x, E_g for other values of E_b was calculated. The solid curve of E_g in Fig. 1 is a plot of the calculated values. The small circles are the measured values. All of them fall on the curve within the range of precision of the measurements. The curve joining the measured values of the average plate current is a straight line passing through the origin. Thus it is seen that when the grid-excitation potential and the grid bias to the amplifier are modulated according to (12) and (13) respectively, the radio-frequency and direct-current conditions are simultaneously satisfied.

B. *Comparison of the Modulation Characteristics Using the Present Scheme and Those Obtained Using Constant Grid Bias and Using Grid-Leak Bias*

1. Under Normal Operating conditions: In ordinary operation of class C amplifiers, the grid bias is usually made equal to about twice the cutoff value and the grid excitation is sufficient to make the minimum plate potential equal to or slightly larger than the maximum grid potential. Measurements show that these are also the optimum operating conditions considering efficiency, output, driving power, etc. So the above are here taken as normal operating conditions.

As in the calculations by the point-to-point method, the grid bias was made 100 per cent linearly modulated and the grid excitation so modulated as to satisfy the radio-frequency condition at the peak of modulation. The output voltage across the load resistance, the average plate current and grid current measured, together with those obtained using constant bias and grid-leak bias are plotted against the plate-supply potential in Fig. 2.

It is seen that the present scheme gives almost perfect modulation characteristics and the reflected resistance on the modulator is constant. Using fixed bias and constant excitation, the second-harmonic distortion at the peak of modulation is about 6 per cent and the reflected resistance varies from about 14,500 to 10,000 ohms. Grid current increases rapidly when E_b is decreased and grid dissipation at the trough of modulation is ten times that under carrier conditions. The use of grid-leak bias gives better modulation characteristics. The second-harmonic distortion at the peak of modulation is about 3 per cent and the grid current is approximately constant.

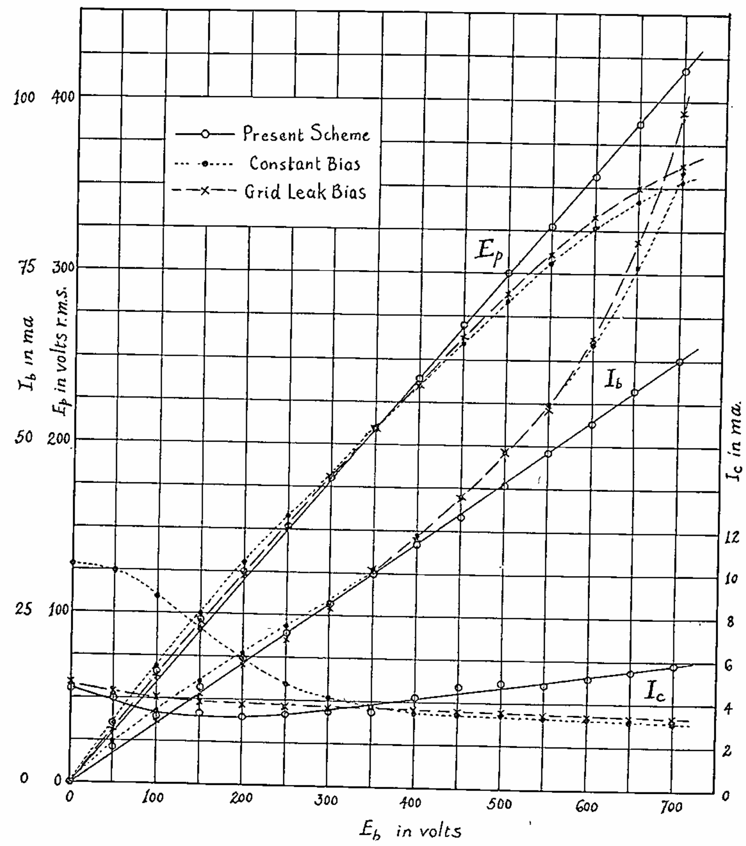


Fig. 5—Modulation characteristics of triode 801 under conditions of insufficient bias. Carrier conditions: $E_{b_0}=350$ volts; $E_{c_0}=-43.75$ volts; $E_{g_0}=67$ volts root-mean-square. Solid curves: E_c is 100 per cent modulated and E_g is 77.5 per cent modulated with E_b.

2. Under Conditions of Insufficient Bias: The present scheme should give linear modulation characteristics under practically all conditions of operation. Using a grid-bias voltage equal to the cutoff value, the modulation characteristics obtained using the present scheme and those using constant bias and grid-leak bias are shown in Fig. 5. The grid excitation used in

[3] Atherton Noyes, Jr., "A sixty-cycle bridge for the study of radio-frequency power amplifiers," Proc. I.R.E., vol. 23, pp. 785–806; July, 1935.

obtaining the solid curves in Fig. 5 is 78 per cent linearly modulated instead of using the theoretically required values. This explains the slight deviations of the E_p and I_b curves obtained using the present scheme from the straight-line courses. It is evident that when constant bias or grid-leak bias is used, the output wave form will be badly distorted even though a modulator with zero internal resistance is used. The second-order harmonic distortion is about 10 per cent in both cases. The reflected resistance varies from 11,000 to about 2800 ohms in the case of constant bias and to about 2000 ohms when grid-leak bias is used. So the resultant distortion when an actual modulator is used will be still higher. Efficiency will be low and the average plate dissipation, when 100 per cent modulated by a sine wave, will be higher than 1.5 times that under carrier conditions.

It is true that modulated amplifiers are almost never operated under these conditions. It was shown[4] that the wave form of some modulating signals is very unsymmetrical. When the proper polarity of the microphone is used, it is possible to modulate to 150 per cent or more on the positive half cycle without reducing the plate voltage to zero on the negative half cycle of modulation. So when sufficient bias is used at carrier, it will be insufficient at the positive peaks.

3. Distortion Produced in Conventional Methods of Modulation as a Function of Excitation Potential: The present scheme of modulation gives linear modulation characteristics under practically all conditions of operation. Using a bias equal to twice the cutoff value at carrier, the output voltage and average plate current are both straight lines passing through the origin when the ratio of the maximum grid potential to the minimum plate potential r is 0.21 and 1.9. But the distortion produced using conventional methods is a function of this ratio. The difference in per cent of the actual and the ideal output voltage across the load resistance at the peak of modulation, as a function of the grid excitation potential, is shown in Fig. 6 for the cases of constant and grid-leak bias, respectively. In both cases the distortion is a maximum when the ratio r is about equal to unity, or just where operating conditions are optimum. This may be explained as follows. The efficiency at the carrier is always higher than that at the peak of modulation. In the case of constant bias, the input power at the peak of modulation is larger than four times the input at carrier when the excitation is small. This larger input counteracts the smaller efficiency and the resultant distortion is therefore small. As the excitation is increased, the efficiency at carrier increases more rapidly than that at the peak and the input power at the peak gradually becomes equal to and eventually smaller than four times the input at the carrier. Thus the distortion also increases. After the ratio r becomes equal to unity, the efficiency at carrier increases less rapidly while that at the peak of modulation increases at the initial rate. Therefore the distortion becomes smaller. In the case of grid-leak bias, the power input at the peak of modulation is always larger than four times the input at carrier and the difference becomes larger with the excitation potential. This tends to counterbalance the smaller efficiency at the peak. Therefore, the maximum

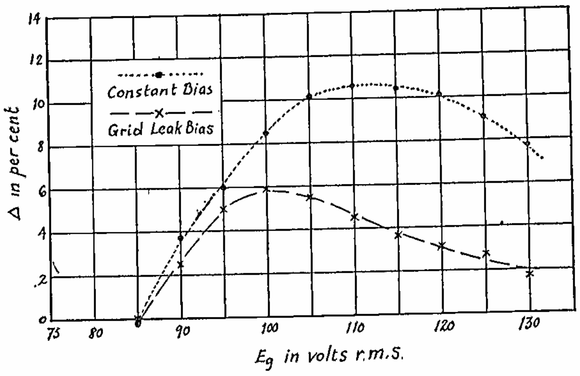


Fig. 6—Difference between the ideal and measured output voltage across the load resistance of triode 801, Δ as a function of E_g. Load resistance = 6000 ohms.

of distortion has a smaller value and occurs before the ratio r has reached the value unity.

Discussion

The above measurements were made on the small radiation-cooled triode 801. The use of grid-leak bias usually produces smaller distortion than when constant bias is used. But the variation of the reflected resistance on the modulator is more severe. A combination of constant and grid-leak bias is usually used, because though the use of grid-leak bias produces smaller distortion, there is the disadvantage of damaging the tube on losing excitation. The distortion is consequently intermediate between that of the two. In high-power triodes, there is usually secondary emission from the grid, the latter affects its operation on the positive half cycle of modulation when grid-leak bias is used. The bias and efficiency will be decreased but the input power increased. The deficiency of output on the positive half cycle can be, or more than, made up at the expense of increased input and plate dissipation. Usually localized plate heating or "focusing" takes place due to the decreased bias; and in severe cases the tube may "run away" or be permanently damaged.

The present scheme, besides giving linear modulation characteristics, avoids all these defects. Efficiency is constant throughout the cycle of modulation. A constant bias of low internal impedance modulated by the plate-modulating signal prevents the detrimental effects of secondary emission. Practical circuit arrangements suggest themselves.

Acknowledgment

The author wishes to acknowledge his thanks to his colleagues of the Institute, especially Professor C. K. Jen, for their discussion and interest in the course of the work.

[4] J. L. Hathaway, "Microphone polarity and over-modulation," *Electronics*, vol. 12, pp. 28, 29, and 51; October, 1939.

The Ionosphere and Radio Transmission, November 1940, with Predictions for February 1941*

NATIONAL BUREAU OF STANDARDS, WASHINGTON, D.C.

AVERAGE critical frequencies and virtual heights of the ionospheric layers as observed at Washington, D. C., during November are given in Fig. 1. Critical frequencies for each day of the month are given in Fig. 2. Fig. 3 gives the November average values of maximum usable frequencies, for undisturbed days, for radio transmission by way of the regular layers. The maximum usable frequencies were determined by the F layer at night and by the F_2 layer during the day. Fig. 4 gives the expected values of the maximum usable frequencies for radio transmission by way of the regular layers, average for undisturbed days, for February, 1941. All of the fore-

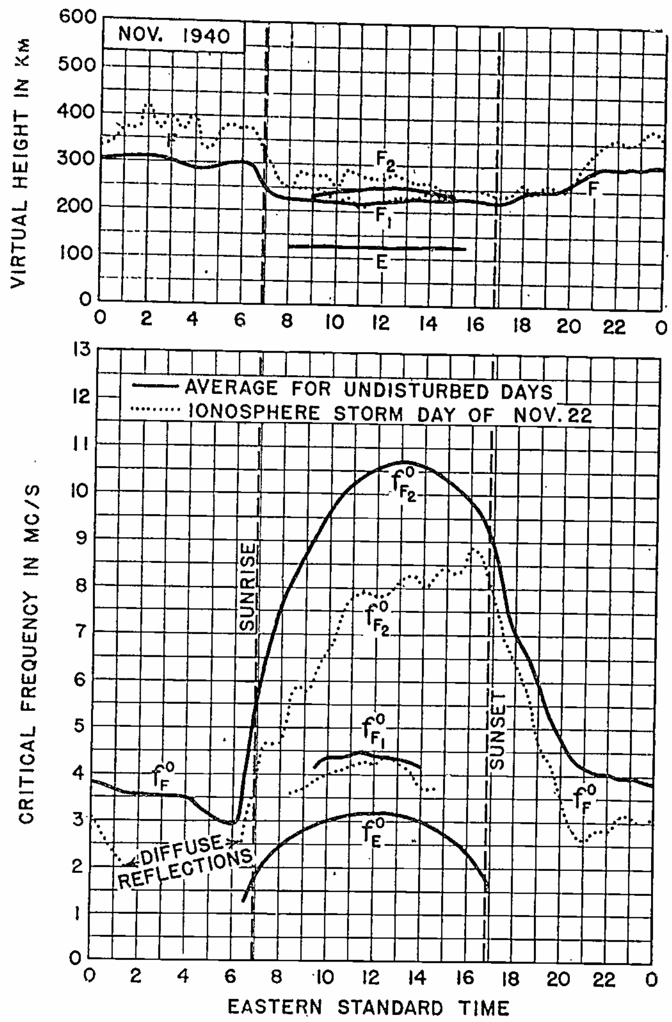


Fig. 1—Virtual heights and critical frequencies of the ionospheric layers, observed at Washington, D. C., November, 1940.

TABLE I

IONOSPHERIC STORMS
(approximately in order of severity)

Day and hour E.S.T.	h_F before sunrise (km)	Minimum $f_F{}^0$ before sunrise (kc)	Noon $f_F{}^0{}_2$ (kc)	Magnetic character[1] 00–12 G.M.T.	Magnetic character[1] 12–24 G.M.T.	Ionospheric character[2]
November						
21 (from 2100)	—	—	—	—	—	0.5
22	377	diffuse	7800	1.0	0.7	1.2
23	387	diffuse	7500	1.0	0.5	1.3
24 (through 0700)	310	2100	—	0.1	0.1	0.1
12 (from 1500)	—	—	—	0.5	1.1	1.0
13	373	diffuse	11700	1.4	0.8	1.2
14	342	3000	9600	0.6	0.6	0.5
15 (through 0300)	302	3200	—	0.4	0.5	0.1
25 (from 0400)	304	<1600	9500	0.5	1.1	0.8
26	347	<1600	9500	0.9	0.5	0.9
27 (through 0600)	318	2300	—	0.1	0.3	0.2
9	330	2400	8100	0.6	0.4	0.6
10 (through 0300)	307	—	—	0.0	0.0	0.1
4 (from 1100)	—	—	12000	0.3	0.8	0.5
5 (through 0600)	289	3100	—	0.8	0.4	0.2
29 (from 2200)	—	—	—	—	—	0.3
30 (through 0700)	325	<1600	—	0.9	0.5	0.3
21 (through 0800)	329	diffuse	—	0.9	0.8	0.3
28 (from 2300)	—	—	—	—	—	0.2
29 (through 0700)	357	2000	—	0.9	0.9	0.2
1 (0200 through 0700)	334	2100	—	0.6	0.2	0.1
27 (from 2300)	—	—	—	—	—	0.1
28 (through 0700)	327	2100	—	0.0	0.3	0.1
For comparison: average for undisturbed days	299	2960	10510	0.2	0.2	0.0

[1] American magnetic-character figure, based on observations of seven observatories.

[2] An estimate of the intensity of the ionospheric storm at Washington, on an arbitrary scale of 0 to 2, the character 2 representing the most severe disturbance.

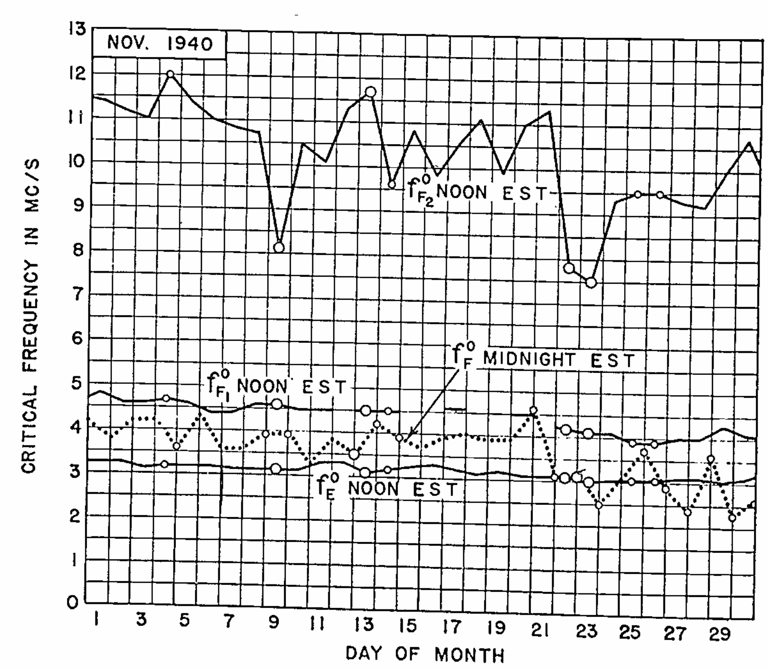


Fig. 2—Midnight $f_F{}^0$ and noon $f_E{}^0$, $f_{F_1}{}^0$, and $f_{F_2}{}^0$, for each day of November. Open circles indicate critical frequencies observed during ionospheric storms; sizes of circles represent approximately the severity of the storms.

* Decimal classification: R113.61. Original manuscript received by the Institute, December 12, 1940. These reports have appeared monthly in the PROCEEDINGS starting in vol. 25, September, 1937. See also vol. 25, pp. 823–840; July, 1937. Report prepared by T. R. Gilliland, N. Smith, and F. R. Gracely.

going are based on the Washington ionospheric observations, checked by quantitative observations of long-distance reception.

Ionospheric storms are listed in Table I. No sudden ionospheric disturbances were observed during November. The details of one ionospheric storm day are shown

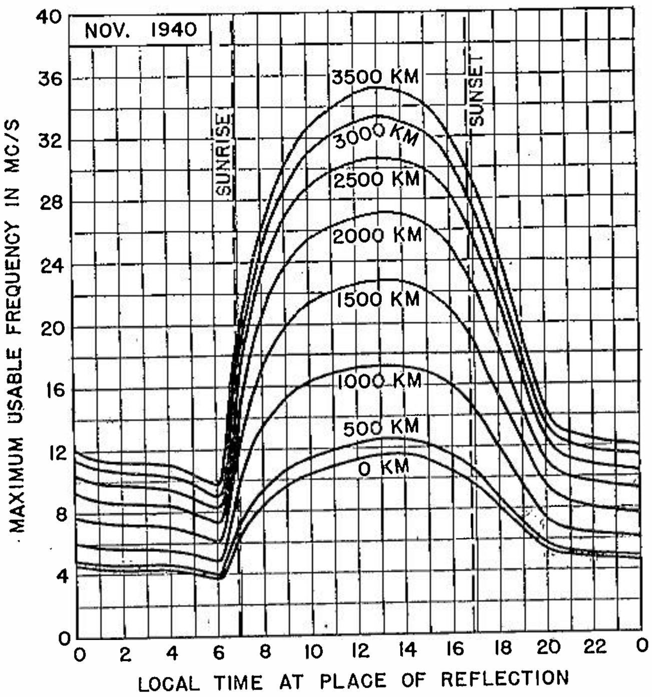


Fig. 3—Maximum usable frequencies for dependable radio transmission via the regular layers, average for undisturbed days for November, 1940. These curves and those of Fig. 4 also give skip distances, since the maximum usable frequency for a given distance is the frequency for which that distance is the skip distance.

in Fig. 1. The open circles in Fig. 2 indicate the noon and midnight critical frequencies observed during the ionospheric storms listed in Table I. The sizes of the circles roughly represent the severity of the storms. Table II gives the approximate upper limit of frequency of strong sporadic-E reflections at vertical incidence.

TABLE II

Approximate Upper Limit of Frequency in Megacycles of the Stronger Sporadic-E Reflections at Vertical Incidence

Day	Hour, E.S.T.																							
	00	01	02	03	04	05	06	07	08	09	10	11	12	13	14	15	16	17	18	19	20	21	22	23
2				4						4	4					4	4							
6																			4	3				3
7	5	4	4	3															4					
8		4					4											4	8	8	4			
9		4																	4	3	5	3		
10				4	7	5	5	4	4	7	4	4	5				4	5	7	9		4		
13													5									3	3	
14																					3	3		
15								4	5	7														
16																		3	5	3				
17							5	5																
18																		4						
19					8	7	6	7									4							5
20	7					6																		
21					4																			
24						4		3																3
25	3						4																	
26													4											
27							4	4																
28		4						4														3	3	
29													5	7	7	7	4							
30				4		3	4																	

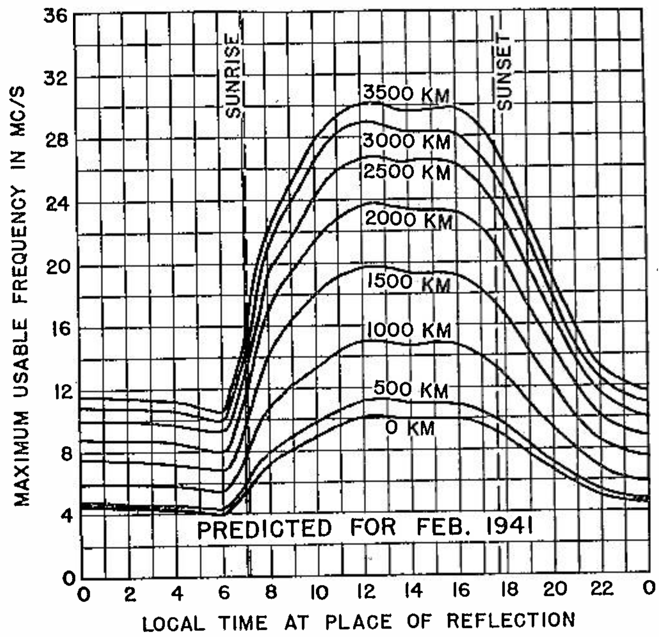


Fig. 4—Predicted maximum usable frequencies for dependable radio transmission via the regular layers, average for undisturbed days, for February, 1941. For information on use in practical radio transmission problems, see Letter Circulars 614 and 615 obtainable from the National Bureau of Standards, Washington, D. C., on request.

The Ionosphere and Radio Transmission, December, 1940, with Predictions for March, 1941*

NATIONAL BUREAU OF STANDARDS, WASHINGTON, D.C.

Average critical frequencies and virtual heights of the ionospheric layers as observed at Washington, D. C., during December are given in Fig. 1. Critical frequencies for each day of the month are given in Fig. 2. Fig. 3 gives the December average values of maximum usable frequencies, for undisturbed days, for radio transmission by way of the regular layers. The maximum usable frequencies were determined by the F layer at night and by the F_2 layer during the day. Fig. 4 gives the expected values of the maximum usable frequencies for radio transmission by way of the regular layers, average for undisturbed days, for March, 1941. All of the foregoing are based on the Washington ionospheric observations, checked

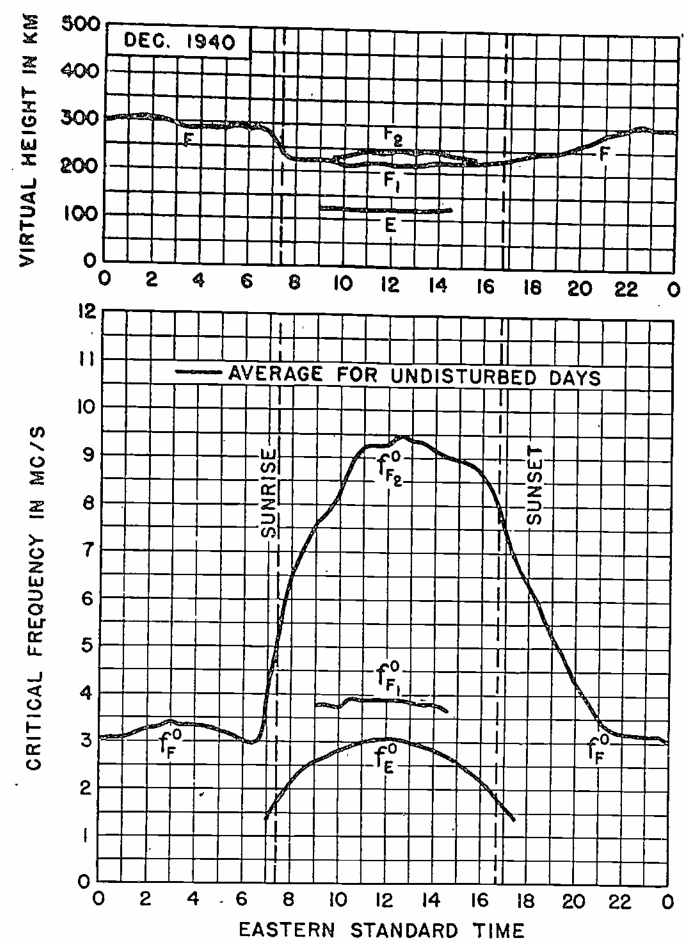


Fig. 1—Virtual heights and critical frequencies of the ionospheric layers, observed at Washington, D. C., December, 1940.

TABLE I
IONOSPHERIC STORMS (APPROXIMATELY IN ORDER OF SEVERITY)

Day and hour E.S.T.	h_F before sunrise (km)	Minimum $f_F{}^0$ before sunrise (kc)	Noon $f_{F_2}{}^0$ (kc)	Magnetic character[1] 00–12 G.M.T.	Magnetic character[1] 12–24 G.M.T.	Ionospheric character[2]
December						
20 (from 2100)	—	—	—	0.7	1.2	0.3
21 (until 0800)	340	1800	—	1.0	0.9	0.3
30 (0300 through 0700)	354	<1600	—	1.0	1.0	0.3
For comparison: average for undisturbed days	293	3030	9270	0.3	0.4	0.0

[1] American magnetic-character figure, based on observations of seven observatories.
[2] An estimate of the intensity of the ionospheric storm at Washington, on an arbitrary scale of 0 to 2, the character 2 representing the most severe disturbance.

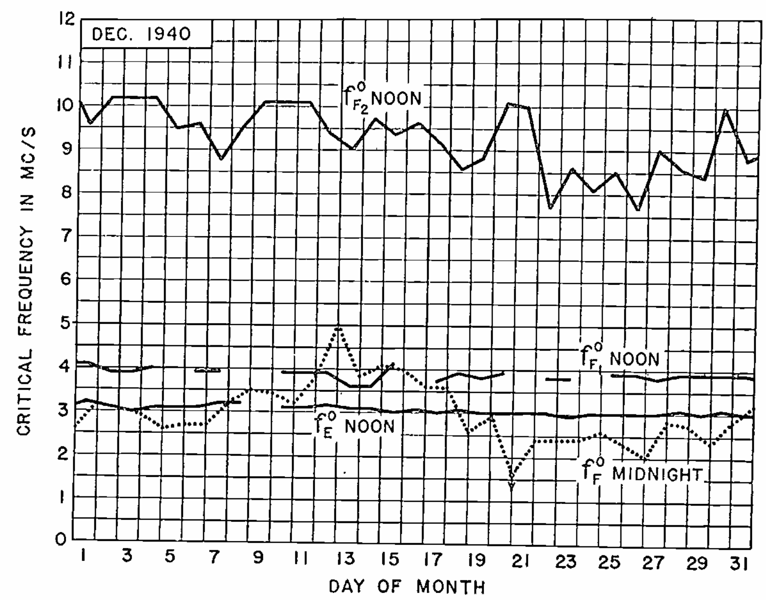


Fig. 2—Midnight $f_F{}^0$ and noon $f_E{}^0$, $f_{F_1}^0$, and $f_{F_2}{}^0$, for each day of December. The open circle and arrow indicate midnight critical frequency of less than 1.6 megacycles observed during ionospheric storm of Dec. 20–21.

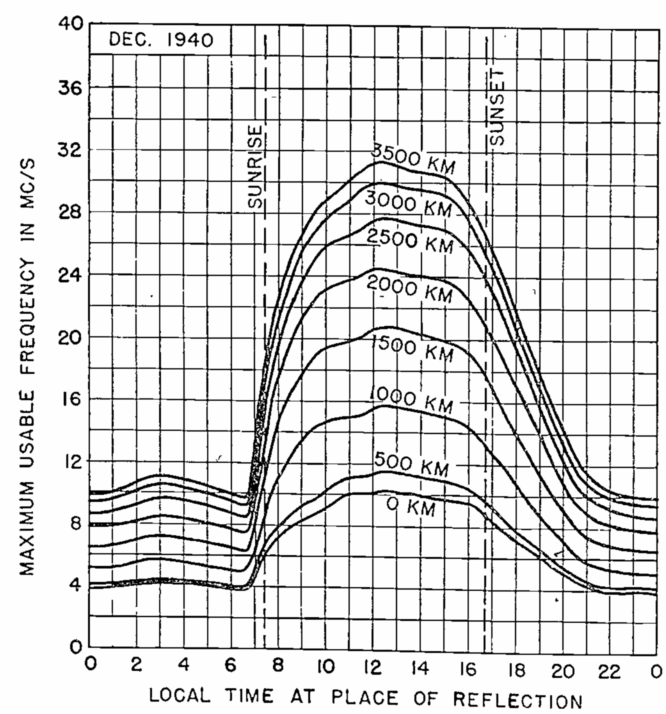


Fig. 3—Maximum usable frequencies for dependable radio transmission via the regular layers, average for undisturbed days for December, 1940. These curves and those of Fig. 4 also give skip distances, since the maximum usable frequency for a given distance is the frequency for which that distance is the skip distance.

* Decimal classification: R113.61. Original manuscript received by the Institute January 13, 1941. These reports have appeared monthly in the PROCEEDINGS starting in vol. 25, September, 1937. See also vol. 25, pp. 823–840; July, 1937. Report prepared by N. Smith, T. R. Gilliland, A. S. Taylor, and F. R. Gracely of the National Bureau of Standards.

by quantitative observations of long-distance reception.

Ionospheric storms are listed in Table I. No sudden ionospheric disturbances were observed during December; indeed none have been observed since October 18.

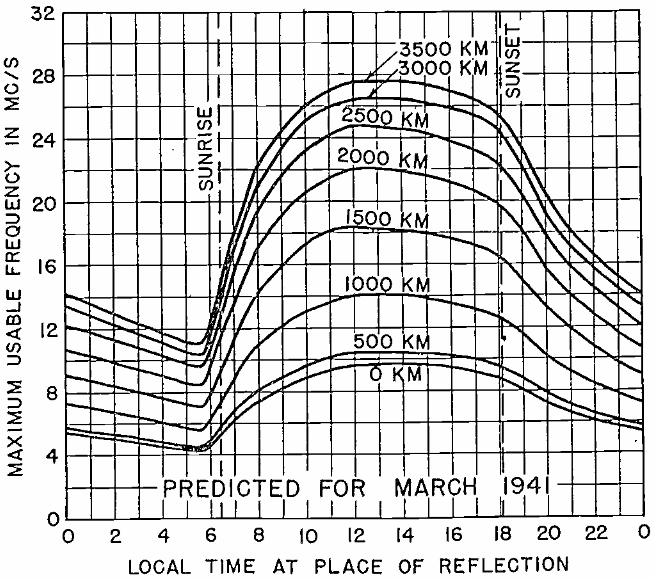


Fig. 4—Predicted maximum usable frequencies for dependable radio transmission via the regular layers, average for undisturbed days, for March, 1941. For information on use in practical radio transmission problems, see Letter Circulars 614 and 615 obtainable from the National Bureau of Standards, Washington, D. C., on request.

Ionsopheric storms also were almost nonexistent. The point indicated by an open circle in Fig. 2 indicates the night ionospheric storm of December 20–21. The arrow indicates that the critical frequency at this time was less than 1.6 megacycles (the lower frequency limit of

TABLE II

APPROXIMATE UPPER LIMIT OF FREQUENCY IN MEGACYCLES OF THE STRONGER SPORADIC-E REFLECTIONS AT VERTICAL INCIDENCE

Day	Hour, E.S.T.																						
	00	01	02	03	04	05	06	07	08	09	10	11	12	13	14	15	16	17	18	19	20	21	22	23
Dec.																								
1									5	4	5	4	5	3										
2						5		4																
3										4							8	9	10	8	4			
4																					9	8		
5												5	5											
6	4																							
7		4																						
10																	4			6				
11	3						3	10	9															
14																			7	8				
15														5										
16																				4	4	3		
17													5	8	8	6	4	3	3	3				
18																				4	4	3	3	3
19	3	6	4	3		3	5	4	3											3	6	7	7	3
20	3									4		8												
22				3	3	3	3											4	3	3			3	3
23			3	3	4	4	5	4																
24												7	7	9	3		6	4						
25					7					3						9	9	5						
26															4	4	7	5	4					
27						3	4	4	4					7	5	8	10	8			3	3	3	
28								3	3									3						
29			3	3																4				
30				7	7	4	4	4			8		4											
31																					3	3		

the recorder). No point is shown in Fig. 2 for the storm of Dec. 30 because it started after midnight and stopped before noon. Table II gives the approximate upper limit of frequency of strong sporadic-E reflections at vertical incidence.

Institute News and Radio Notes

Board of Directors

The last regular meeting of the 1940 Board of Directors was held on Wednesday, December 4, and those present were: L. C. F. Horle, president; Melville Eastham, treasurer; Austin Bailey, W. R. G. Baker, F. W. Cunningham, H. T. Friis (guest), Alfred N. Goldsmith, Virgil M. Graham, R. A. Heising, C. M. Jansky, Jr., F. R. Lack, F. B. Llewellyn, Haraden Pratt, B. J. Thompson, H. M. Turner, A. F. Van Dyck, H. A. Wheeler, L. P. Wheeler, and H. P. Westman, secretary.

Thirty-one applications for Associate and twenty-five for Student grade were approved.

The request of the Chicago Section to permit its affiliation with the Illinois Engineering Council was approved.

J. H. Dellinger was appointed as our representative on the American Documentation Institute.

The invitation from the Detroit Section for the Institute's summer convention to be held in Detroit on June 23, 24, and 25, was accepted.

The Annual Meeting of the Board of Directors was held on Wednesday, January 8, 1941, and those present were: F. E. Terman, president; H. T. Friis, Virgil M. Graham, O. B. Hanson, R. A. Heising, L. C. F. Horle, F. B. Llewellyn, B. J. Thompson, H. M. Turner, H. A. Wheeler, L. P. Wheeler, and H. P. Westman, secretary.

H. P. Westman was appointed to continue as Secretary.

Haraden Pratt was appointed Treasurer.

Alfred N. Goldsmith was appointed Chairman of the Board of Editors.

A. B. Chamberlain, I. S. Coggeshall, Melville Eastham, C. M. Jansky, Jr., and A. F. Van Dyck, were appointed Directors to serve during 1941.

Applications for the Associate grade numbering thirty-nine, two for Junior, and seventy-two for Student were approved.

Committees to serve during 1941 were appointed.

An Executive Committee consisting of the President, Treasurer, Secretary, and Melville Eastham, R. A. Heising, and B. J. Thompson was appointed to assist the Board in an advisory capacity.

A budget covering operations during 1941 was adopted.

A contribution of $200.00 to the operation of the Engineering Societies Library was made.

A Special Papers Committee to solicit manuscripts on material or surveys known to a relatively limited group of specialized workers but of interest to the general membership, consisting of B. E. Shackelford, chairman, Alfred N. Goldsmith, and H. A. Wheeler, was appointed.

In regard to the readmission of former members of the Institute, the following resolution was adopted:

> "RESOLVED, that the Board of Directors approves for readmission to the grade of membership previously held (or in the Associate grade if formerly a Junior and now over 21 years of age) those former members (a) whose memberships terminated in 1940 or earlier and who pay either current dues or all dues in arrears or (b) whose memberships terminate on March 31, 1941, and who pay dues for 1941 at a later date during that year. In all cases, the payment of a new entrance fee where such would normally be required is waived, the payment of a transfer fee is required where a transfer in grade is in order, and in the case of Associates who formerly had the right to vote it is understood that on their readmission voting privileges will no longer be theirs which is in accordance with the Constitution as now amended."

A petition for the establishment of an Institute Section in the Dallas-Fort Worth region of Texas was granted.

Electronics Conference

On October 11 and 12, an Electronics Conference was held at Stevens Institute of Technology in Hoboken, N. J.

These conferences are arranged for the benefit of the advanced workers in the electronics field. The contributions are not in the nature of formal papers but are of the discussion type.

Sessions were devoted to dense electron beams, the production of high-velocity particles, the electron microscope, high-vacuum pumping, and ultra-high-frequency measurements.

The registration for both days totaled 252.

Fourth Pacific Coast Convention

The Fourth Pacific Coast Convention of the Institute was held in Los Angeles, California, on August 28, 29, and 30, 1940. The program was published on pages 280–286 of the June PROCEEDINGS and all of the papers except these numbered 1 and 21 were given. Two seminars on high-fidelity audio-frequency systems and frequency modulation were held. A number of inspections trips were made.

The attendance totaled 356 of whom 24 were women.

Rochester Fall Meeting

At the Rochester Fall Meeting which was held on November 11, 12, and 13, twenty technical presentations were made in accordance with the program which was published on page 423 of the September PROCEEDINGS. The total attendance was 517.

New York Meetings

"Color Television" was the subject of a paper by P. C. Goldmark in collaboration with J. N. Dyer, E. R. Piore, and J. M. Hollywood of the Columbia Broadcasting System. The system described was designed to operate in the allotted 6-megacycle bands. An interlaced scanning of 343 lines was used.

A color disk carrying three filters, red, green, and blue, rotates in front of the television pickup tube and a similar disk is operated in front of the picture tube at the receiver. The filter colors are balanced to give the effect of pure white when the picture is white.

The picture is completely scanned 60 times a second which means that the odd lines are scanned in 1/120th of a second and the even lines in a similar length of time. The picture is first scanned in red for the odd lines. Green is then used for scanning the even lines. During the green scanning, the red odd-line image has faded and these lines are then scanned in blue. This has taken 3/120th of a second and has resulted in the picture being completely scanned one and one half times and the three color scannings to have been completed once. Thus we have a picture sequence of 60 times per second and a color sequence of 40 times per second. By proper shaping of the filters in relation to the decay factor of the fluorescent material, carry-over of an image from one color to the next is avoided.

While the primary colors are red, yellow, and blue, it has been found that green which includes yellow is more effective in providing a satisfactory picture.

The color disks at the transmitter and receiver must be operated synchronously and this can be arranged either through the use of synchronous motors operated on a common power system or by the transmission of a sychronizing signal.

Pictures transmitted in color may be received as black-and-white images on a receiver which is not equipped with a color disk. Thus the system permits the addition of color to receivers not originally designed for it.

A demonstration showing both color and black-and-white reception was given utilizing a transmitter located in a near-by building. Comparison between color and black and white was restricted to the use of 343 lines. The pickup at the transmitter was from color motion pictures which had been taken at 64 frames per second and were run at 60 frames per second.

October 3, 1940, L. C. F. Horle, president, presiding.

"Recent Developments in Marine Communication Equipment" was presented by J. F. McDonald of the engineering department of the Radiomarine Corporation of America.

It was pointed out that the development of new types of marine radio communication equipment is keeping pace with the current large shipbuilding program which has been undertaken by the United States Maritime Commission. A description was given of new types of radio transmitters and direction finders which have been developed for use on cargo and tanker vessels. The features of a new type of radiotelephone installation for small harbor craft were also described. The equipment specially designed and installed in the new passenger liner *S. S. America* was shown in detail.

November 6, 1940, L. C. F. Horle, president, presiding.

D. E. Noble, director of research of the Galvin Manufacturing Company presented a paper on "The Connecticut State Police Frequency-Modulated-Wave Radiotelephone System" which covered his work while a consultant to the State of Connecticut.

This frequency-modulation system includes ten fixed transmitters operating on a frequency of 39,500 kilocycles and 225 25-watt mobile stations operating on 39,180 kilocycles. Receivers are provided at the fixed stations for both frequencies.

In each of the ten patrol areas of the state, a high quiet location was selected as the site for the remotely controlled 250-watt transmitter. Coaxial antennas supported at the top of 180-foot guyed steel poles were used.

The antennas for the mobile units are mounted in the center sections of the steel roofs of the cars. These antennas provide a substantially circular field pattern in contrast to the highly directional pattern produced by the usual back-of-car mounted antenna.

The use of ten fixed stations has proved to be an economical arrangement. The failure of any fixed station in the system does not seriously handicap the service since the area of any station may be covered by the stations in the adjacent regions. The entire state system can be serviced by three men.

State-wide two-way coverage is provided with a large factor of safety. The use of frequency modulation makes it possible for areas not adjacent to each other to operate simultaneously without interference.

Two-way communications to mobile units over distances of 50 to 75 miles has been possible. Every fixed station in the state can hear and communicate with every other fixed station in the state.

December 4, 1940, L. C. F. Horle, president presiding.

Sections

Atlanta

A paper on the "Fifty-Kilowatt International Broadcast Station WLWO" was presented by J. W. Herbstreit of the Federal Communications Commission, Atlanta office.

In introducing the subject, emphasis was placed on the importance of international broadcasting as a means of promoting friendly relations with countries in this hemisphere and for the purpose of national defense. Over ten years ago the Crosley Corporation completed its first international transmitter, and its experimental call letters, W8XL, were recently changed to WLWO.

A general description of the transmitter was given. Particular attention was paid to the methods used for changing frequency rapidly. This is necessary in order to provide an adequate signal over a long period of time. An antenna system which gives a directional power gain larger than 10 on a number of different frequencies was described. In using a rhombic antenna, the power which would normally be dissipated in heat in the terminal resistor is returned to the antenna thus increasing the over-all efficiency and avoiding the inclusion of a resistor capable of dissipating large amounts of power.

October 18, 1940, P. C. Bangs, chairman, presiding.

Three students of the Georgia School of Technology presented a group of papers. The first on "Television Camera and Associated Circuits," was by J. R. Haeger. By means of a block diagram the various components of a television transmitter and their relationships were indicated. Scanning was considered in detail and the auxiliary equipment required for it was described. To obtain a high-quality picture, the beam must move at constant velocity along each line of the picture, the lines must be perfectly straight, and they must be of equal length. The type of wave form required to produce such scanning and methods of generating this wave form were considered.

The second paper, by F. L. Denton, described a method of "Applying Negative Feedback in a Radio Transmitter." Reverse feedback is advantageous in that it reduces hum and frequency and waveform distortion. Limitations of the amount of feedback usable and difficulties with "singing" were described. The use of a cathode-ray tube to determine the frequency of the "singing" and methods of eliminating it by the insertion of a capacitance-resistance network across the first amplifier were discussed.

The third paper was on "Recent Advances in Broadcast Equipment" and was presented by W. F. Wrye. It covered the field broadly, including studios, transmitters, and antennas. Developments in microphones, phonograph pickups, and program amplifiers were outlined. Consideration was then given to the construction, insulation, illumination, and ventilation of studios. At the transmitter, new types of tubes, increased efficiency in the operation of amplifiers, and improvements in buildings were described. The paper was concluded with some suggestions of improvements that may be made in the near future.

It was the opinion of the judges that the paper by Mr. Denton deserved the section award which is a year's dues in the Institute as an Associate.

November 22, 1940, P. C. Bangs, chairman, presiding.

Baltimore

G. H. Brown of the RCA Manufacturing Company (Camden), presented a paper on "Ultra-High-Frequency Antennas."

Dr. Brown explained first the methods of matching the transmission line to the antenna and included the use of quarter- and half-wave coaxial lines as matching media.

A combined 10- and 20-meter horizontal rotary antenna was described for the benefit of the amateurs. Performance curves showing the directional characteristics and the effects of proper impedance matching were given.

Multielement turnstile antennas for frequency modulation and television were then considered. A television antenna consisting of 6 elements, fed by concentric lines was described. Quarter- and half-wave matching is used throughout to give a 10-megacycle band having only small impedance changes and providing minimum reaction between the picture and the sound transmitters.

November 15, 1940, Ferdinand Hamburger, chairman, presiding.

"A Modern System for Upper-Air Investigation by Radio" was described by L. E. Wood and Ralph Chappel, development engineer and chief engineer, respectively, for Julien P. Friez and Sons.

Mr. Chappel presented the historical background of the field of transmitting information utilizing time cycles in a radio telemeter. After describing the troubles encountered with contacts, modified clocks, mechanical difficulties in cams, and injury during shipment, he introduced the system developed by Diamond-Hinman-Dunmore of the National Bureau of Standards.

Mr. Wood described the radio-sonde instrument produced by his organization. The radio transmitter complete with batteries weighs 2 pounds. It uses a double triode, 1C6G, one part of which operates as an audio-frequency oscillator controlled by the measuring device and the other structure produces a carrier at about 70 megacycles.

Air temperature is measured by an electrolytic resistor, the humidity affects the length of a hair which adjusts a resistor through a cam mechanism, and the barometric pressure operates on a diaphragm coupled to a switching device. Each of these functions varies the audio-frequency modulation which ranges between 10 and 200 cycles.

The signals are received on a superregenerative receiver and recorded on a strip chart. When the balloon bursts, the sonde descends by means of a parachute and about one half of them are returned for the offered reward. About 30 stations in this country utilize the radio sonde.

December 20, 1940, Ferdinand Hamburger, chairman, presiding.

Buenos Aires

The "Useful Sensitivity of Radio Receivers" was the subject of a paper by M. J. Kobilsky of the Douglas Laboratories of Buenos Aires.

After a general survey of all the noises that appear at the receiver output, those of external origin were excluded and the discussion was limited to noise generated in the receiver. Methods of diminishing several types of noises were described and it was shown that the residual noise results from the Johnson-Nyquist effect and space-charge limiting. Design formulas for the calculations of these two noises were given. A check of the available formulas for computing the space-charge noise in radio-frequency pentodes was made with the help of data provided by the manufacturers of the tubes and from the literature.

Methods used in Europe and America to produce noiseless types of tubes were reviewed. Data on the specific noise per kilocycle of pass band and of the noise resistance were given for tubes for both continents and included radio-frequency amplifiers and converters.

A simplified theory of noise currents in the presence of a carrier in the different stages of a receiver and methods of calculating the permissible signal-to-noise ratio were given.

The paper closed with a demonstration of the program quality which may be expected in a commercial receiver with different values of carrier amplitude. A standard-signal generator modulated by a broadcast program was used in the demonstration.

October 25, 1940, P. J. Noizeux, vice chairman, presiding.

A "History and Description of the Pacheco Station" was presented by A. L. Rodriguez, chief of commercial services of the Division of Radio Communications of the Argentine. The paper was presented as part of an inspection tour of the General Post Office transmitting station located at Pacheco in the Province of Buenos Aires.

Various transmitters for both long and short waves were inspected. Several had been designed and constructed in the Post Office shops in Buenos Aires. These transmitters are used for various services such as in the point-to-point-telegraph system and in supplying radiotelephone services to ships at sea and to distant points in Argentina.

Of special interest was a demonstration of the newly established public radiotelephone service to small ships, tugs, yachts, and airplanes. A truck in which direction-finding equipment had been installed to assist in locating illegal transmitters was on display. The operation of the equipment and methods of carrying on this work were described.

A. T. Cosentino, chairman of the section and Director of Radio Communications for Argentina, presided.

November 9, 1940.

"Basic Problems of Telephone Transmission" were discussed by Alejandro Nadosy, assistant to the general manager of the Cia. Internacional de Telefonas.

The subject was introduced with a description of methods basic to modern carrier telephony. Symmetrical modulators, filters, two-stage modulation, and group modulation were discussed.

An analysis was then presented of the principal characteristics and use of various conductors such as open wire lines, symmetrical and coaxial cables, and tubular ducts. Based on electromagnetic theory a series of simple final formulas were developed. In discussing tubular ducts, stress was placed on the enormous mechanical difficulties which must be met in the practical application of these structures for telephony in addition to the economic and physical requirements.

In the discussion, information was brought out concerning the transmission of television signals over wire lines, the use of coaxial and symmetrical cables for telephony, and the comparative effectiveness of various dielectrics at ultra-high frequencies.

November 29, 1940, P. J. Noizeux, vice chairman, presiding.

Buffalo-Niagara

A "Discussion of Fluorescent Materials" was given by B. S. Ellefson, research chemist for the Hygrade Sylvania Corporation. The paper considered fluorescent materials used in cathode-ray tubes and in commercial lamps.

Pure chemicals are not available directly from manufacturers and methods of obtaining materials of sufficient purity for lamp and cathode-ray tubes were described. The application of the fluorescent materials to the glass envelopes was outlined.

In commercial lamps, the conversion of energy to light is as high as 80 per cent, while in the cathode-ray tube it is between 10 and 15 per cent.

The color of the illumination given off by fluorescent coatings may be varied by mixing chemicals in proper proportion. The color resulting from the mechanical mixing of the materials may be estimated from a curve. The effects of these mixtures were shown by operating a number of small tubes using different proportions of chemicals. Commercial-type lamps of several different colors were demonstrated.

A number of illustrations used in presenting the paper were prepared with fluorescent material and became visible when irradiated by ultraviolet light.

December 11, 1940, B. Atwood, chairman, presiding.

F. W. Stillwagen, projection engineer of the Fairchild Aviation Corporation, presented "Some Notes on a Direct Lateral Crystal Recording Cutter Head."

The subject was introduced with an historical outline beginning with Edison's recording on foil and wax cylinders to the present methods. Lateral recording was then discussed in detail using a crystal cutting head, cutting either directly on aluminum or on a specially coated aluminum or cardboard disk. A combination of constant velocity and modified amplitude recording was recommended for maximum signal-to-noise ratio and greatest recording time. A crossover frequency of 800 cycles was recommended. The upper frequency limit of the recording head described was 8000 cycles.

The amplifier to be used with this recording head should have an output of plus 35 decibels (0.006 watt, reference level) with not more than 1 per cent distortion. Both low- and high-frequency gain controls should be provided. The average recording level used was plus 24 decibels. Rochelle-salt crystals are used in the cutting head. The resonant frequency of the moving system occurs at 4200 cycles and a selective damping pad is used to compensate for it. Other tuned circuits located between the amplifier and the recording head provide the desired response characteristic. A method of checking the response of the cutting head was described.

January 8, 1941, B. E. Atwood, chairman, presiding.

Chicago

The preliminary meeting was devoted to a discussion of "Modern Methods of Testing Receivers" by J. K. Johnson, head of the Hazeltine Laboratories of Chicago.

The regular meeting was addressed by H. L. Rusch, executive vice president of the A. C. Nielsen Company, on the subject of "Measuring Accurately the Listening Habits of the American Radio Audience."

The method described involved coupling to the radio receiver a graphic recording device known as the Audimeter. It indicates the time during which the receiver is in operation and the time at which each change in tuning is made. When these data are analyzed, a statistical picture of the listening habits of the radio audience may be developed.

October 18, 1940, E. H. Kohler, chairman, presiding.

J. F. Inman, president of Associated Research, Inc., discussed at the preliminary meeting "The Lie Detector, Its Principle and Uses." A stock model of one of these devices was demonstrated. In tests made on a member of the audience, Mr. Inman not only announced the number chosen by the subject but also two other numbers which the subject held in mind in an effort to defeat the detector.

At the regular meeting, Andrew Alford, of the International Telephone Development Company, presented a paper on "High- and Ultra-High-Frequency Measurements." Starting with a simple circuit consisting of inductance, resistance, and capacitance in series as might be used for circuits operating below 10 megacycles, it was shown that as the frequency was progressively increased above that value circuit changes must also be made to maintain the accuracy of the measurements.

November 15, 1940, R. A. Kay, chairman of the Membership Committee, presiding.

At the preliminary session, "A Proposed Bill for the Licensing of Engineers in the State of Illinois" was discussed by Mr. Le Clair of the Commonwealth Edison Company, also secretary of the Illinois Engineering Council.

It was pointed out that similar laws are now in effect in 42 states and cover all branches of professional engineering. The proposed bill is based on a "Model Law" prepared by a group of national engineering societies some years ago. The general provisions of the pending bill were outlined and discussed.

At the regular meeting, R. P. Glover of the Jensen Radio Manufacturing Company, spoke on "Recent Developments in Radio Speaker Design."

A consideration of the effectiveness of a loud speaker must be based on its environment in addition to any measured response-frequency characteristic. Tests in and out of doors reveal entirely different behaviors. The "piston" of the speaker may differ substantially from the size of the cone and may be controlled by an air column. The variation in response of the ear presents a problem in this field. The design of trumpet-type units was discussed. Efficiency and the effects of environment were considered at length.

This was the annual meeting of the section and in the balloting, G. I. Martin of the RCA Institutes was elected chairman; Carl Hassel of the Zenith Radio Corporation was named vice chairman; and R. A. Kay, of the RCA Institutes was designated secretary-treasurer.

December 13, 1940, E. H. Kohler, chairman, presiding.

Cincinnati

W. R. Jones, director of commercial engineering of the Hygrade Sylvania Corporation, explained "Why Engineers Pull Their Hair." He pointed out that there are three contending groups in a commercial enterprise; the customer, the manufacturing department, and the sales department. Difficulties arising among these groups may often be clarified through a fourth party, the commercial engineering department or the sales engineer.

An analysis of the problems presented by simple automatic-volume-control systems was given with special emphasis on the tube design. The two most serious problems facing the tube manufacturer involve contact potential and grid emission. Both affect the sensitivity of the receiving set. Another factor is the choice of a proper operating time constant.

The problems involved in the development of a special tube for a certain application were also described.

It was observed that problems seem to recur in 5-year cycles which may indicate a limit of memory on the part of the majority of those connected with the design of radio tubes and circuits.

November 19, 1940, C. H. Topmiller, chairman, presiding.

This meeting was held jointly with the local section and with the University of Cincinnati student branch of the American Institute of Electrical Engineers. A dinner meeting preceded the regular meeting and was addressed by W. H. Wolfram, a practicing physician who spoke on "What You Get."

W. L. Everitt, professor of electrical engineering at Ohio State University, presented a paper on "Frequency Modulation." The fundamental differences between amplitude and frequency modulation were outlined. The propagation characteristics of radio waves were then discussed. The problem of interference may be solved in part by using high-powered transmitters and selective receivers. Directive antenna systems and high percentages of modulation also help. The operation of both amplitude- and frequency-modulation systems was then presented graphically. A mechanical model assisted in clarifying the points of interest.

A demonstration was presented of the two systems by means of a receiver and two transmitters capable of operating on both types of modulation. The effect of noise from an electric razor on both systems was demonstrated. With the two transmitter carriers separated by about 6 kilocycles and with different programs on each, little hetrodyne interference appeared with frequency modulation as the signal strength of one transmitter was varied. With amplitude modulation the interference was very noticeable.

The frequency band widths used in frequency modulation and the elements of the receivers and the transmitters were then described. Two methods of obtaining frequency modulation in transmitters were discussed.

This was the annual meeting of the section and J. M. McDonald of Stations WLW-WSAI-WLWO, was elected chairman; G. L. Haller, Aircraft Radio Laboratory (Dayton), was named vice chairman; and W. L. Schwesinger, was reelected secretary-treasurer.

Connecticut Valley

A paper on a "Photo Timer" was presented by Kirby Austin, an amateur photographer, who is also an engineer for the General Electric Company (Bridgeport). The basic circuit of the device was described and modifications of it to increase the accuracy of timing were discussed. Leakage from a photoelectric tube proved a major problem. By removing the base of the photoelectric tube and using suitable insulating materials in the critical circuits, this requirement was met.

W. S. Bachman, also of the General Electric Company (Bridgeport), spoke on "Alnico and the Permanent-Magnet Speaker." The relation of Alnico to other magnetic materials was shown. As its name implies, Alnico is a mixture of aluminum, nickel, iron, and cobalt. The materials are mixed in powdered form, pressed to shape, and given final machining before sintering. The finished material is harder than most wheels so that it is difficult to grind it.

The design of loud-speaker magnetic fields was then outlined and a sample computation presented. By using this material, the field coils are eliminated and the size of the structure reduced. It has excellent age characteristics and reasonable cost for quantity production.

November 14, 1940, K. A. McLeod, chairman, presiding.

A paper on "Recent Improvements in Frequency-Modulation-Receiver Design" was given by J. A. Worcester of the General Electric Company (Bridgeport). A summary of this paper appears in the report of the December 5, 1940, meeting of the Philadelphia Section.

After the paper, Professor Cady welcomed the group to Scott Laboratory of Wesleyan University and invited those present to inspect the work which was going on in the various laboratories. He then described some of the work which he has been doing on both quartz and Rochelle-salt crystals.

Professor Van Dyke then described some of the research on crystals which is being carried on under his supervision.

December 12, 1940, K. A. McLeod, chairman, presiding.

Detroit

"A Unique Selective-Calling System" was the subject of a paper by F. C. Collings, an engineer of the RCA Manufacturing Company (Camden). This equipment was developed for calling individual stations or groups of stations in the emergency communication services such as police, public utility, ship, and aircraft systems.

1000-cycle pulses are produced by a commutator driven by a synchronous motor. Commutators bars are connected to push-button switches which are operated in accordance with a calling code. The pulses are impressed on the carrier of the transmitting station.

The decoding unit at the receiver also employs a commutator operating in step with the coding commutator. Synchronizing pulses precede the code pulses and control the operation of a vibrator which drives the decoding commutator. The bars of the decoding commutator are connected to relays which are arranged to permit only the proper code sequence to operate a master relay which actuates the calling device.

By employing a limiter and a band-pass filter, false operation resulting from noise is avoided. A steady 1000-cycle noise results in a canceled call. The calling equipment can be arranged to repeat a call four times.

September 20, 1940, J. D. Kraus, chairman, presiding.

G. H. Brown of the RCA Manufacturing Company (Camden) presented the paper "Ultra-High-Frequency Antennas" which has been described in the report of the November 15, 1940 meeting of the Baltimore section.

October 18, 1940, R. J. Schaefer, vice chairman, presiding.

S. J. Begun, research engineer for the Brush Development Company, presented a paper on "The Development of Crystal and Magnetic Cutters for Disk Recordings."

The general problems of disk recording were first considered. The design of both crystal and magnetic cutters was then outlined. A description was given of a crystal cutter capable of recording frequencies between 40 and 9000 cycles, having a response within 2 decibels of the value at 1000 cycles, giving distortion of less than 2 per cent, and requiring very little driving power. A description was given also of a temperature-controlled crystal cutter.

Magnetic-type recording was then considered. Its application in the broadcast field was described and the ease with which a previous recording may be obliterated and the tape used for an indefinite number of times was stressed. The use of tape recording to produce artificial reverberation was described.

The paper was concluded with a demonstration of a tape recorder. In addition, several commercial pressings were compared with instantaneous disk recordings.

November 15, 1940, R. J. Schaefer, vice chairman, presiding.

A. H. Waynick of Wayne University, presented the paper on "Experiments on the Propagation of Ultra-Short Radio Waves" which was published in the October issue of the PROCEEDINGS.

This was the annual meeting of the section and M. Cottrell, was elected chairman; R. J. Schaefer, Briggs Manufacturing Company, was named vice chairman; and Paul Frincke, WJBK, was re-elected secretary-treasurer.

December 20, 1940, R. J. Schaefer, vice chairman, presiding.

Emporium

"Chemistry as Applied to Radio Tubes" was presented by R. E. Palmateer, chief chemist of the Hygrade Sylvania Corporation, at a joint meeting with the local section of the American Chemical Society.

The paper was restricted to the chemical problems presented in the production of radio tubes. The metals used in tubes were listed and their allowable impurities considered. The process of preparing and assembling elements into a tube was then described. The chemistry involved in the function of each part and the preparatory treatment given to anodes, grids, cathodes, emitting materials, insulation, and getters was described.

November 19, 1940, B. S. Ellefson, chairman of the American Chemical Society section, presiding.

"Observations of an Engineer on the Continent and in the Near East on the Threshold of War" were presented by H. B. Richmond, treasurer of the General Radio Company. The observations were based on a trip made by the speaker just before the outbreak of the war. Particular attention was given to the attitude, morale, and general viewpoint of the peoples visited. It was pointed out that all these factors will influence either directly or indirectly the termination of the present war.

This was the annual meeting of the section and R. K. Gessford, of the Hygrade Sylvania Corporation, was elected chairman; L. D. Andrews, of the Stackpole Carbon Company, was named vice chairman; and C. W. Reash, of the Hygrade Sylvania Corporation, will be the new secretary-treasurer.

December 17, 1940, C. R. Smith, chairman, presiding.

J. J. Rogan, patent attorney, discussed "The Defects in Our Patent System." The present patent system is based on a provision of the original Constitution which has never been modified. One of its defects is the absence of protection given to those responsible for scientific discoveries. In research work the procedure necessary to attain a given result cannot be patented.

Patent pools and cross licensing result in the concentration of power in the hands of a few people holding the basic patents. Useful patents can be withheld from public application by refusal on the part of the owner either to manufacture or license others to produce the invention.

The interference procedure in obtaining a patent often delays its issuance for several years. The high cost of obtaining a patent discriminates against the poor inventor. In addition, there are so many courts which rule on a patent that long delays are very common.

Need for changes in our patent system was recognized as early as 1813. It is considered likely that some revisions may be made in the near future which may affect substantially our patent law.

Statistics were presented to show that there is a correlation between industrial growth and the rate of issuance of patents. The greatest number were issued during the reconstruction period following the Civil War when half a million patents were granted between 1860 and 1890.

January 7, 1941, R. K. Gessford, chairman, presiding.

Los Angeles

"Supersonic Oscillations, Their Production and Uses" was the subject of a paper by Chester Weaver, an engineer of the Los Angeles Bureau of Power and Light. A method of generating supersonic oscillations of high amplitude was described. Powers of a kilowatt were employed and the output oscillations are effectively transmitted to the load by means of a coupling system of wide application.

L. B. Beckwith, professor of biology at the University of California in Los Angeles, then discussed "The Effect of Supersonic Oscillations on Microorganisms." Dr. Beckwith has been associated with Mr. Weaver in this work. He outlined the effects of high-amplitude oscillations in pasteurizing substances and in destroying certain types of organisms which are difficult to kill by other means.

April 18, 1940, A. C. Packard, chairman, presiding.

G. C. Conner, commercial engineer of the Hygrade Sylvania Corporation, presented a paper on "New Vacuum Tubes and Their Uses." A major part of the paper was devoted to the problems met in the manufacture of tubes suitable for use in frequency-modulation receivers, the problems involved in the production of these receivers, and the advantages resulting from the use of frequency modulation. A demonstration contrasting amplitude modulation and frequency modulation was presented through the use of records.

May 14, 1940, A. C. Packard, chairman, presiding.

S. S. Mackeown, professor of physics at the California Institute of Technology, presented a paper on "The Theory of Frequency Modulation and Noise Reduction." The history of the development of frequency modulation was presented. By means of vector and mathematical formulas, the chief features of frequency modulation, such as noise reduction and sideband power were described.

June 25, 1940, A. C. Packard, chairman, presiding.

"The Facilities of the Columbia Recording Corporation in Hollywood" were described by Chester Boggs, engineer-in-charge of that organization. The problems met in the production of phonograph records for general sale were considered. This field was compared with that of recording for broadcasting and sound pictures. Various types of materials and recording equipment were discussed. A description was given of the recently installed facilities of the Columbia Recording Corporation.

M. T. Smith of the General Radio Company presented a paper on "The Twin-T Network," a null-indicating device for measuring impedances up to 30 megacycles.

October 15, 1940, A. C. Packard, chairman, presiding.

"Radio and Communication Facilities of the Civil Aeronautics Administration" was the subject of a paper by E. G. Crippen and Mr. Thomas of the Civil Aeronautic Administration. It included various types of radio beacons, localizers, and fan markers. The theory of air navigation was also discussed. The first author presented the theoretical aspects and the second author described the actual operation of the equipment and the experiences encountered with it.

This was the annual meeting and W. W. Lindsay, Jr., consultant and assistant director of sound for the General Service Studios, was named chairman; J. N. A. Hawkins, of the Walt Disney Studios, was elected vice chairman; and C. F. Wolcott of Gilfillan Brothers, Inc., became secretary-treasurer.

November 19, 1940, A. C. Packard, chairman, presiding.

R. S. Richardson of the Mount Wilson Observatory, presented a paper on "The Influence of Solar Activity on Radio Propagation." It was pointed out that many investigations of sunspots have been made in an attempt to correlate their effects on the earth. No definite results were established until 1936.

The equipment used at Mount Wilson for this research was described. The formation, movement, and rate of development of sunspots were outlined. Particular attention was drawn to the fact that most sunspots originate approximately 30 degrees from the sun's equator and progress towards but do not cross the equator. Particularly bright eruptions result in excess ultraviolet radiation and are correlated with ultra-high-frequency fade-outs. The diurnal variation in the earth's magnetic field is enhanced at these times but no correlation with magnetic storms has been established.

Magnetic storms follow a 27-day cycle and present predictions are correct about 68 per cent of the time where large sunspots are involved and about 40 per cent of the time for small spots. Data were given on the magnetic storm of March 24, 1940 which was the most violent ever recorded.

In all cases where correlation has been established, it has been proved that the energy involved is radiant rather than corpuscular.

December 17, 1940, W. W. Lindsay, Jr. chairman, presiding.

Philadelphia

D. G. C. Luck of the RCA Manufacturing Company (Camden), presented a paper on "Aerial Navigation by Omnidirectional Radio Range." The purpose of direction finders located on aircraft and on the ground and of directional beacons was described. The beacon offers an excellent system for use in the ultra-high frequencies. The direction finders have provided most service in the past. There is a place for an omnidirectional beacon which provides straight radial courses in all directions about itself and automatic indication on the receiving craft of the true bearing to the beacon.

An explanation of the operation of such a system was presented. Five fixed antennas in the form of a square with a center unit provide two crossed figure-eight directional patterns and a circular pattern. Two-phase modulation of the directive patterns and impulsive keying of all patterns produce a result which is almost cardioidal and which rotates uniformly. It is interrupted momentarily once during each revolution.

An ordinary receiver provides a sinusoidal output which corresponds to the rotation of the pattern and an impulse corresponding to the keying, the relative phase of these outputs depending on the direction of the receiver from the transmitter. The sinusoidal output is split in phase and moves the spot of a cathode-ray indicator steadily in a circular path. The impulse from keying moves the spot momentarily outward from the circle providing a radial mark, the position of which indicates on a suitable scale the bearing of the receiver in relation to the transmitter. A differential tube circuit provides an indication on a zero-center meter of whether the impulse occurs before or after the mean value of the sine wave. By using a phase splitter to set the sine wave, the deviations of the receiver from any desired bearing will be indicated with proper sense.

A description was given of an experimental 200-watt, 125-megacycle omnidirectional range beacon. Local ground calibration indicated over-all instrumental errors averaging less than 1 degree and nowhere exceeding 3 degrees. Field tests showed average errors of $3\frac{1}{2}$ degrees and not exceeding $10\frac{1}{2}$ degrees. Vagaries of wave propagation were considered responsible for the larger errors in flight.

The operation of the system was demonstrated by using large-scale indicators and a model airplane over a map. Cross-bearing navigation between two points not provided with radio facilities was demonstrated.

November 7, 1940, C. M. Burrill, chairman, presiding.

J. A. Worcester, Jr., of the General Electric Company (Bridgeport) presented a paper on "Recent Improvements in Frequency-Modulation Receivers."

Adequate sensitivity requires limiting on all inputs down to the ultimate limit imposed by tube noise and it was shown that this requires a total gain of about 4,000,000 ahead of the limiter. The limitations in the gain that can be realized at the intermediate frequency were considered and were followed by an analysis of the difficulties involved in obtaining the required total gain by additional amplification at the signal frequency. This amount of gain is not attainable with the conventional superheterodyne circuit and the double superheterodyne was considered briefly. Attention was then given to the oscillator and the cascade converter circuits used in the frequency-modulation translator being described. The performance characteristics of the limiting system, oscillator, the intermediate-frequency amplifier, and the discriminator circuit were discussed.

December 5, 1940, C. M. Burrill, chairman, presiding.

P. C. Goldmark, chief television engineer of the Columbia Broadcasting System, presented a paper on "Television in Color." A summary of this paper appears in the report of the November 13, 1940, meeting of the Washington Section.

January 2, 1941, C. M. Burrill, chairman, presiding.

Pittsburgh

D. L. Chestnut of the Atlantic Division of the General Electric Company, presented a paper on "Frequency Modulation." The general theory of amplitude and frequency modulation was discussed. It was pointed out that although coverage with high-frequency transmission is limited, it is expected that such transmission of frequency-modulated waves will provide quality programs over an area equal to the primary service area of an equivalent amplitude-modulated transmitter.

Two frequency-modulated and two amplitude-modulated transmitters were used to demonstrate the improvements which frequency modulation provides.

This meeting was held jointly with the Pittsburgh section of the American Institute of Electrical Engineers and G. R. Patterson of that organization presided.

November 12, 1940.

The "Design and Manufacture of Oil and Paper Condensers" was the subject of a paper by W. S. Franklin, chief engineer of John E. Fast and Company. In designing a condenser the information required includes the operating voltage, capacitance and capacitance tolerances, ambient temperature, intermittent or continuous service, operation on alternating or direct voltages, and the function of the condenser in service.

Aluminum has largely displaced copper and tin as electrode material. Kraft paper is generally employed as dielectric and several layers are used to prevent the few conducting particles in the paper from short-circuiting the condenser. The relative merits of natural and synthetic oils and waxes when used as impregnating agents were discussed. Consideration was also given to the protective container and associated apparatus.

The manufacturing processes were described and included winding, drying, impregnating, final assembly, and testing. Data were given on the ultimate breakdown voltage, power factor, insulation resistance, capacitance, and life of completed condensers.

November 18, 1940, R. E. Stark, chairman, presiding.

The "Application of Electronic Devices to Power Systems" was discussed by J. G. McKinley, electronics engineer for the West Penn Power Company. The application of electronic methods to measuring instruments, supervisory and control equipment, communications, and power-converting equipment was discussed. It was pointed out that reliability and costs are the primary considerations in choosing between established methods and the relatively newer electronic means.

The remote-metering arrangement now used by the West Penn Power Company was described. The problem of controlling the frequency of generators with sufficient accuracy to permit the use of synchronous clocks was discussed.

December 9, 1940, R. E. Stark, chairman, presiding.

San Francisco

A seminar meeting on antennas was devoted to a discussion of three papers. W. G. Wagener of Heintz and Kaufman, Ltd., led the review of the papers "Television Transmitting Antennas for Empire State Building" by N. E. Lindenblad and "Simple Television Antennas" by P. S. Carter which appeared in the *RCA Review*. H. A. Smith, also of Heintz and Kaufman, Ltd., led the discussion on "A Multiple-Unit Steerable Antenna for Short-Wave Reception" by H. T. Friis and C. B. Feldman which appeared in the PROCEEDINGS.

November 6, 1940, L. J. Black, vice chairman, presiding.

"Comments on Recent Technical Developments and Current Problems" were presented by Frederick Ireland of the General Radio Company. Various types of measuring devices were described with emphasis on measurements in the high-frequency region.

Measurements can be made at frequencies up to 300 megacycles with good accuracy and up to 600 megacycles with fair accuracy. The determination of an electromotive force can be done satisfactorily at frequencies in the order of 1200 megacycles.

A description was given of an oscillator having a range from 150 to 600 megacycles giving a maximum power output of 3.5 watts and at 600 megacycles an output of 1 watt. The accuracy of frequency setting is ± 3 per cent. The output is 75 ohms.

The susceptance-variation method of measuring dielectrics, impedances, and other characteristics was discussed. By determining the capacitance change required to detune a resonant circuit to reduce its current to 70.7 per cent of the resonant value, fairly accurate measurements can be made at high frequencies. At the conclusion of the paper some motion pictures of the General Radio Company plant and its operations were shown.

November 13, 1940, Carl Penther, chairman, presiding.

The Annual Meeting was devoted to a banquet tendered to President-Elect Terman. L. F. Fuller, head of the department of electrical engineering of the University of California, was toastmaster.

In the election of officers, L. J. Black of the department of electrical engineering of the University of California, was elected chairman; H. E. Held, manufacturers' representative, was named vice chairman; and Karl Spangenberg, of the department of electrical engineering of Stanford University, was designated secretary-treasurer.

Seattle

W. R. Hewlett of the Hewlett-Packard Company, spoke on "The Construction of Square-Wave Generators and Their Use in Testing Audio-Frequency Equipment."

Various methods of generating square waves were described and their advantages and disadvantages pointed out. The instrument manufactured by the Hewlett-Packard Company employs a modified multivibrator circuit feeding a "clipping" stage which operates with a small positive potential on the grid of the vacuum tube.

The application of square waves to the testing of audio-frequency amplifiers was described. In a number of cases which were illustrated, the measurements verified calculations based on the circuit constants. A great saving in time over that required for a rigorous mathematical approach was claimed. The greatest objection to square-wave analysis lies in the interpretation of the observed results.

An audio-frequency amplifier was tested at the conclusion of the paper. The effects of the treble and bass compensation controls were demonstrated.

November 25, 1940, R. M. Walker, chairman, presiding.

"New Police Radio Equipment and Technical Features of Police Radio Operation" were described by Floyd Hatfield, communication officer of the Seattle Police Department.

The history of police radio work in the United States was first presented. In 1932 the first high-frequency system and the first two-way mobile installations were placed in service. The development of police radio equipment was slow until about 1935 when the Association of Police Communication Officers was formed. At this time police personnel were unfamiliar with the advantages of two-way radio operation and were reluctant to accept anything so new and little understood by the average policeman.

A recent trend has been toward the use of frequency modulation. In Connecticut, a series of tests have shown that the state can be covered adequately with three 250-watt transmitters. While the conversion of many systems to frequency modulation would be advantageous, the cost is a serious barrier.

Seattle has 32 police cars equipped for two-way operation and 13 additional units equipped with receivers only. A study has shown that two officers operating with a two-way equipped car are the equivalent of six officers patrolling on foot. This ratio is based on the number of arrests made by the two groups.

December 20, 1949, R. M. Walker, chairman, presiding.

Toronto

"General Considerations Regarding Disk Recording and Playback" was the subject of a paper by A. L. Williams, president of the Brush Development Company.

The field of recording was divided into three parts, pressings for public sales, electrical transcriptions for broadcasting, and instantaneous recordings. The problems of the first two types were treated briefly and the major part of the paper was devoted to a discussion of instantaneous recording. A new type of cutting head and crystal pick-up were described.

The paper was closed with a demonstration of instantaneous recording and playback using the new cutting head and reproducer.

April 22, 1940, G. J. Irwin, chairman, presiding.

H. W. Parker, chief engineer of Rogers Radio Tubes, Ltd., presented a paper on "The Time Factor in Radio Design."

An analysis of the value of any manufactured product may be made on a time basis. Time, such as man-hours, might well be used as a measure of the production cost. The true worth of a product was then shown to be the ratio of the "consumption" time to the "production" time, where the production time included such factors as material, labor, taxes, and distribution costs. The paper was illustrated with everyday examples of the principles which were expounded. These principles were then applied more specifically to radio-receiver and television design and production.

In the election of officers which was held, H. S. Dawson, Rogers Radio Broadcasting Company, was named chairman; R. H. Klingelhoeffer, International Resistance Company, Ltd., was designated vice chairman; and L. C. Simmonds of A. C. Simmonds, was elected secretary-treasurer.

May 20, 1940, G. J. Irwin, chairman, presiding.

"Frequency Modulation" was the subject of a paper by Lee McCanne, assistant general manager of the Stromberg-Carlson Telephone Manufacturing Company.

In comparing frequency and amplitude modulation it was pointed out that frequency modulation permits lowered costs and less power and men required at the broadcast transmitter but that the receiving sets would necessarily cost more. He then showed that frequency modulation provides more reliable service within the limitations of the range of the transmitter.

In discussing the receiver, it was pointed out that the limiter eliminates all noise having an amplitude less than the amplitude of the signal, and in the case of interference from another station the limiter selects the strongest signal.

Intermediate frequencies for use in frequency-modulation broadcast receivers have been standardized at 4.3 megacycles. The band allocated to frequency-modulation broadcasting is from 40 to 50 megacycles. The channel is 150 kilocycles in width and 50-kilocycle guard bands are provided between channels.

Some of the merchandising problems met in the distribution of frequency-modulation receivers were discussed. The possibilities of chain broadcasting using a large number of relatively low-power stations all operating on the same frequency was also considered. The experiences obtained in Rochester where two stations are in operation and some of the reactions of the public were discussed.

December 2, 1940, R. H. Klingelhoeffer, vice chairman, presiding.

Washington

"Color Television" was the subject of a paper by P. C. Goldmark, chief television engineer for the Columbia Broadcasting System.

The fundamental problems involved in the development of a color-television system were outlined, together with three possible methods of scanning which give promise of providing this service within the 6-megacycle channel allowed for television transmissions.

The system which has been used in this work uses 343 lines and 60 frames per second. Methods used to separate the colors into three bands for transmission and means of providing a proper color balance in order that the color could be reproduced at the receiver without distortion were described. The advantages of color television in providing an apparent increase in detail in the pictures being received even though the actual number of picture elements transmitted was reduced were illustrated by colored slides.

The mechanical problems of rotating the filters for both the pickup equipment and the receiver were described. Light sources and their effect on the transmission of color were also considered.

November 13, 1940, L. C. Young, chairman, presiding.

"Wide-Band Antenna and Transmission-Line Combinations" were described by N. E. Lindenblad, research and development engineer of RCA Communications. In it there was presented the fundamental concepts of wave propagation on transmission lines, reflection, and other phenomena. The application of these fundamentals to design and measurement problems was discussed with particular attention being given to wide-band circuits. Wide-band radiators and problems arising in connection with directive systems were also considered.

In the election of officers, M. H. Biser, vice president and director of studies of the Capitol Radio Engineering Institute, was made chairman; E. M. Webster, principal engineer for the Federal Communications Commission, was elected vice chairman; and C. M. Hunt, chief engineer of WJSV, was named secretary-treasurer.

December 9, 1940, L. C. Young, chairman, presiding.

Membership

The following indicated admissions and transfers of memberships have been approved by the Admissions Committee. Objections to any of these should reach the Institute office by not later than February, 28, 1941.

Transfer to Member

Cady, C. A., c/o General Radio Company, 30 State St., Cambridge, Mass.
Chatterjea, P. K., Grange House, 7 Highbury Grange, London N. 5, England
Florez, H., 1717 Ave. N, Brooklyn, N. Y.
Kiser, N. L., Box 141, Emporium, Pa.
Mathis, J. D., 802 Telephone Bldg., Dallas, Tex.
McClintock, R. K., 307 E. 4th St., Emporium, Pa.
Packard, L. E., 90 West St., Rm. 1504, New York, N. Y.
Roemisch, J. L., 487-5th Ave., Lyndhurst, N. J.
Smith, C. R., 324 W. 4th St., Emporium, Pa.
Tucker, D. J., 5611 Anita St., Dallas, Tex.
Tuska, C. D., 717 W. Mt. Airy Ave., Philadelphia, Pa.
Weir, I. R., 97 N. Country Club Dr., Schenectady, N. Y.

Admission to Member

Doak, R. S., 47 Bellevue Ter., Collingswood, N. J.
Howard, R. V., 225 Mallorca Way, San Francisco, Calif.
Langstroth, F. D., 314 W. 5th St., Emporium, Pa.
Wilson, I. G., Bell Telephone Laboratories, Inc., 180 Varick St., New York, N. Y.

Admission to Associate (A), Junior (J), and Student (S)

Abramowitz, S., (S) 579 E. 140th St., New York, N. Y.
Ahern, W. R., (S) 43 Woodland St., Worcester, Mass.
Arbon, R. D., (A) 76—44th Ave., Lachine, Que., Canada
Athey, S. W., (S) Box 1150, Stanford University, Calif.
Bailey, F. M., (A) General Engineering Laboratory, General Electric Co., Schenectady, N. Y.
Bailey, R. C., (A) Box 112, Richmond, Va.
Baldwin, E. V., (A) Diagonal Norte 520, Buenos Aires, Argentina
Bamford, D., (S) 2 Willingdon Ave., Toronto 13, Ont., Canada
Barbour, C. W., Jr., (S) 96 Hemenway St., Boston, Mass.
Barton, L. M., (S) Box 321, Clemson, S. C.
Bastian, C. W., (A) 810 Ashland Ave., Buffalo, N. Y.
Berger, F., (S) 708 Pennsylvania Ave., Matamoras, Pa.
Boekhoff, J. H., (S) 1920 Selby Ave., St. Paul, Minn.
Bonney, R. B., (A) 1936 Lexington Ave., Merchantville, N. J.
Bookman, R. E., (S) 2030 E. Speedway, Tucson, Ariz.
Borkowski, C. J., (A) 4304 Beta Ave., Cleveland, Ohio
Bowne, W. L., (A) Y.M.C.A., 5 W. 63rd St., New York, N. Y.
Brady, C. L., (A) Radio Station WJIM, Lansing, Mich.
Brandt, E., (S) 623—13th Ave., S.E., Minneapolis, Minn.
Bridgland, C. J., (A) 71 Grove Ave., Ottawa, Ont., Canada
Brown, S. K., Jr., (S) Rockledge, Ithaca, N. Y.
Brush, E., (A) 107 E. 37th St., New York, N. Y.
Buggy, R. V., (A) 5934 N. 7th St., Philadelphia, Pa.
Burroughs, H. A., (A) 2128 Key Blvd., Arlington, Va.
Bushman, S. F., (A) 5554 W. North Ave., Chicago, Ill.
Cairnes, W. E., (A) 7937 S. Maplewood Ave., Chicago, Ill.
Caldwell, E. E., (S) Rockefeller Hall, Ithaca, N. Y.
Cardella, Q., (S) 1656 Joliette St., Montreal, Que., Canada
Carpenter, P., (S) 125 E. Dickson St., Fayetteville, Ark.
Carson, C. J., (S) 107 S. Lincoln, Urbana, Ill.
Carter, D. F., (S) 811 E. 2nd St., Tucson, Ariz.
Carter, R. H., (A) 2229 Harvard North, Seattle, Wash.
Carver, F. R., (S) 224 Roslyn Ave., Glenside, Pa.
Cates, G. R., (S) 116 Kendal Ave., Toronto, Ont., Canada
Chipman, L. D., (S) 27 Wiley St., Bangor, Maine
Christenson, C. K., (A) 3737 S.W. Hillside Dr., Portland, Ore.
Clark, J. W., (A) Bell Telephone Laboratories, Inc., 463 West St., New York, N. Y.
Clark, O., (S) 3 E. Market St., Iowa City, Iowa
Claussen, L. W., (S) 405 Oak St., S.E., Minneapolis, Minn.
Cleeton, C. E., (A) 1601 Addison Chapel Rd., N.E., Washington, D. C.
Cole, D. P., (A) 12 Sanford Ave., Catonsville, Md.
Collette, J. M., (A) Bean Station, Tenn.
Collis, H. O., (A) Owens-Corning Fiberglas Corp., Newark, Ohio
Corl, J. L., (S) 459 Madison, Corvallis, Ore.
Cramer, S., (S) M.I.T. Dormitories, Cambridge, Mass.
Cutler, S., (S) 2907 Buckingham Rd., Los Angeles, Calif.
Dallas, H. G., Jr., (S) 703 Grantley St., Baltimore, Md.
Danziger, A., (A) c/o Danziger Radio Laboratories, 119 W. 57th St., New York, N. Y.
Davies, H. E., (S) Marshal's Residence, University of Wash., Seattle, Wash.
Davis, C. F., (A) 31 Channing St., Worcester, Mass.
Davison, J. M., (S) University of Western Ontario, London, Ont., Canada
Dean, W. N., (S) 525 W. 113th St., New York, N. Y.
De Lange, O. E., (A) Bell Telephone Laboratories, 463 West St., New York, N. Y.
Demaree, F. E., (S) 318 Welch Ave., Ames, Iowa
Dennison, B. H., Jr., (S) 2312—5th St., Lubbock, Tex.
Derey, C. H., (S) Box 481, Blacksburg, Va.
Deri, H., (A) 659 W. 162nd St., New York, N. Y.
Doyle, J., (A) Freeport Sanatorium, Kitchener, Ont., Canada
Dozier, E. C., (A) 40 Monroe St., New York, N. Y.
Duffy, D. P., (A) Radio Manufacturing Engineers, Inc., 111 Harrison St., Peoria, Ill.
Dulin, W. E., (A) 84 U St., N.W., Washington, D. C.
Durkee, C. E., (S) 4245 Elston Ave., Chicago, Ill.
Durkin, W. P., (A) 5341—66th St., Maspeth, N. Y.
Dyke, E. L., (S) 2540 College Ave., Berkeley, Calif.
Easton, A., (A), Siebenthaler Manufacturing Co., 410 W. 6th St., Kansas City, Mo.
Edstrom, C. A., (S) 412 E. 3rd St., White Bear Lake, Minn.
Edwards, J. M., (A) 226 W. Lanvale St., Baltimore, Md.
Eldred, D., (S) 706 Ontario St., Schenectady, N. Y.
Eichwald, B., (S) 2054 E. 21st St., Brooklyn, N. Y.
Elliott, J. O., (S) 100 Sylvia St., West Lafayette, Ind.
Ernst, J. P., (A) Route 2, Marine Dr., Bellingham, Wash.
Euler, G. M., (S) 64 Portland Pl., Cliffside Park, N. J.
Evans, W. E., (A) 829 Belle Ave., Fort Wayne, Ind.
Exwood, M., (A) "Aviemore," Burtree Lane, Darlington, England
Falk, A., (S) 1220 Prospect St., Ann Arbor, Mich.
Fallows, E. M., (S) 9 Dana St., Cambridge, Mass.
Fee, J. H., (A) 1441 Drummond St., Montreal, Que., Canada
Fenmen, S. M., (S) M.I.T. Dormitories, Cambridge, Mass.

Ferguson, K., (S) University of Western Ontario, St. Thomas, Ont., Canada
Field, J. S., (S) Box 734, Gainesville, Fla.
Fish, P., (A) 2995 Berkshire Rd., Cleveland Heights, Ohio
Fishman, S., (A) Newark College of Engineering, Newark, N. J.
Floyd, G. H., (S) 1925 E. Lee St., Tucson, Ariz.
Foote, D. A., (S) 4528—5th Ave., N.E., Seattle, Wash.
Ford, G. T., (A) 16 Midvale Rd., Mountain Lakes, N. J.
Frank, M., (S) 315 N. Sandusky Ave., Upper Sandusky, Ohio
Frankart, W. F., (A) 741 Michaels Ave., Fort Wayne, Ind.
Frawley, J. E., (S) 2736 University Ave., New York, N. Y.
Freiberg, J. L., (S) Box 202-B Shawnee Run Rd., Indian Hill, Cincinnati, Ohio
Fristoe, H. T., (S) 229 N. Block St., Fayetteville, Ark.
Froland, L., (S) 1800—4th St., S.E., Minneapolis, Minn.
Geisert, W. O., (S) 2512 E. Union St., Seattle, Wash.
Gellert, A. W., (S) 100 Bennett Ave., New York, N. Y.
Ghormley, R. K., (S) 1418 Fairchild St., Manhattan, Kan.
Gillespie, C. N., (J) c/o Post Office, Patchogue, N. Y.
Glasford, G., (S) 301 Stanton Ave., Ames, Iowa
Glass, L. M., (A) 404 E. 10th St., Kansas City, Mo.
Gordon, R. A., (A) 900 W. Madison Ave. Hyattsville, Md.
Grainger, R. E., (A) National Commercial Broadcasting Service, Wellington, New Zealand
Grantham, R. W., (A) 2107 Miller Ave., Richmond, Va.
Greer, J. W., (A) Navy Department, Bureau of Ships, Radio Section, Washington, D. C.
Guernsey, G., (A) 1357 Jordan Ave., Dayton, Ohio
Gumpel, J. E., (A) Philips Argentina, Calle Herrera 527-41, Buenos Aires, Argentina
Gunning, W. F., (S) 807 S. Bundy Dr., West Los Angeles, Calif.
Gurley, J. W., (A) 17 Inman St., Cambridge, Mass.
Hafstrom, J. R., (A) 422 Ash Ave., Ames, Iowa
Hargen, L. A., (S) WATL, Henry Grady Bldg., 26 Cain St., Atlanta, Ga.
Hargens, C. W., III, (S) 101 Nichols Hall, Massachusetts Institute of Technology, Cambridge, Mass.
Harris, H. F., (S) R.F.D. 3, Eaton Rapids, Mich.
Harris, L., (A) 383 Bainbridge St., Brooklyn, N. Y.
Harrison, A. E., (A) Sperry Gyroscope Co., Garden City, N. Y.
Hasenbeck, H. W., (A) 3445 Oakwood Ter., Washington, D. C.
Hastings, H. A., (S) 1444 Yale Station, New Haven, Conn.
Hatfield, F., Jr., (A) Radio Police Department, 417 Yestler Way, Seattle, Wash.
Haus, R., (A) 537 Hyde St., San Francisco, Calif.
Herring, R.A., Jr., (A) 2400—20th St., N.W., Washington, D. C.
Hersey, J. B., (S) Physics Department, Lehigh University, Bethlehem, Pa.
Hill, D. B., (A) 3 Parker Rd., Winchester, Mass.
Hinchey, R. L., (S) 25 Everett St., Waltham, Mass.
Hobart, R. D., (S) Box 582, Virginia Polytechnic Institute, Blacksburg, Va.
Hoeffner, C. H., (S) M.I.T. Graduate House, Cambridge, Mass.
Holcomb, L., (S) Box 400, Cary Hall East, West Lafayette, Ind.
Holme, W. M., (A) 360 Bellevue Ave., Yonkers, N. Y.
Horne, R. C., Jr., (A) 422 Parade, Lexington, Va.
Houck, G. B., Jr., (S) 85 Wesley Ave., Port Chester, N. Y.
Howell, P. N., (A) 1401 Pearl St., Joplin, Mo.
Huber, H. E., (A) 2830 Colerain Ave., Cincinnati, Ohio
Humphrey, A. J., (S) 1105 W. Clark St., Urbana, Ill.
Iden, F. W., (S) 353 Massachusetts Ave., Cambridge, Mass.
Ingels, E. D., (A) 214 Griffith St., Jackson, Miss.
Ishii, R., (S) 9-127 Tokiwacho, Urawa, Japan
Ittelson, W. H., (S) 915 West End Ave., New York, N. Y.
Jacobs, C. F., (A) 9 Prospect Park West, Brooklyn, N. Y.
Jaksha, F., (S) Box 7308, University of Minnesota, Minneapolis, Minn.
Jameson, J. W., (J) 541 E. 182nd St., New York, N. Y.
Janowsky, R. H., (A) 17 Hagen Ave., North Tonawanda, N. Y.
Johnston, D. L., (S) 31 Wilmington Way, Brighton 6, Sussex, England
Jones, J. P., (A) 1616 Whitman Ave., Butte, Mont.
Jordan, N. L., (S) 123 E. Dickson, Fayette, Ark.
Kahgan, J. J., (A) 2323—62nd St., Brooklyn, N. Y.
Kahnke, J., (S) 710—13th Ave., S.E., Minneapolis, Minn.
Karr, A. C., (S) 943 Malcolm Ave., West Los Angeles, Calif.
Kass, S., (A) 243 Eastbourne Ave., Long Branch, N. J.
Kechker, A. I., (A) Calle 42, No. 396, La Plata, Argentina
Keller, J. F., (A) Klamath Falls Union High School, Klamath Falls, Ore.
Kerby, G. A., (S) 14 Dimitri Pl., Larchmont, N. Y.
Kirkpatrick, G. M., (S) Roseville, Ill.
Klemm, G. H., (S) 189 Belleville Rd., New Bedford, Mass.
Kluegel, J., (S) 108 N. Saratoga Ave., St. Paul, Minn.
Kohler, F. L., (S) 1321 Holman Rd., Oakland, Calif.
Koontz, C. E., (A) 216 W. Pall Mall St., Winchester, Va.
Kramer, K., (A) Jensen Radio Manufacturing Co., 6601 S. Laramie Ave., Chicago, Ill.
Kransz, P. A., (A) 5888 Ridge Ave., Chicago, Ill.
Krause, A. J., (S) 17 Ellison Ave., Freeport, L.I., N. Y.
Kwong, L. Y., (S) 422 N. Salisbury, West Lafayette, Ind.
Lafaye, W. P., (S) 5758 Canal Blvd., New Orleans, La.
Lambert, J. B., (S) 322—13th Ave., S.E., Minneapolis, Minn.
Large, W., (S) Hotel Vendome, Minneapolis, Minn.
Latour, M. H., (S) Box 2336, University Station, Gainesville, Fla.
Leap, G. A., (A) 3420 N.W. 3rd St., Miami, Fla.
Leslie, F. E., (S) 209 E. Armory Ave., Champaign, Ill.
Lewis, M. T., (A) 802 Telephone Bldg., Dallas, Tex.
Ley, G. S., (A) 201 Park Dr., Catonsville, Md.
Lifschitz, S., (S) 1245 Eastern Pkwy., Brooklyn, N. Y.
Liimatainen, T. M., (S) 412 Camden Ct., Ann Arbor, Mich.
Locke, R. C., (S) 336 Roux St., Gainesville, Fla.
Love, G. R., (S) 293 Huron St., London, Ont., Canada
Loveberg, A. G., Jr., (S) 2218 N. New Hampshire Ave., Hollywood, Calif.
Mabry, F. S., (A) 38 Holmehurst Ave., Catonsville, Md.
Mackay, J. R., (A) 242 Hunter Bldg., Ottawa, Ont., Canada
Marks, W. K., (A) Northern Broadcasting & Publishing, Ltd., Thomson Bldg., Timmins, Ont., Canada
Markusen, D. L., (S) 152 Orlin Ave., S.E., Minneapolis, Minn.
Marston, R. S., (S) 4 Van Brunt Ave., Dedham, Mass.
Martin, W. R., (S) WRUF, Gainesville, Fla.
Mason, V. V., (S) 1355 Kingston Rd., Toronto, Ont., Canada
Mathewson, E. P., Jr., (A) R.F.D. 1, Scotch Plains, N. J.
May, L. W., Jr., (A) 9428 Hobart St., Dallas, Tex.
McCoy, R. E., (A) 523 N.E. Monroe St., Portland, Ore.
McCurdy, R. V., III, (A) Woods Hole Oceanographic Institute, Woods Hole, Mass.
McDonald, D., (S) 66 Bay State Rd., Boston, Mass.
McGlumphy, W., (A) Box 204, West Liberty, W. Va.
McKeown, M., (S) 13590 Artesian Ave., Detroit, Mich.
Meneley, C. A., (S) 1104 W. Springfield Ave., Urbana, Ill.
Mengel, J. T., (A) 947 Maple Ave., Schenectady, N. Y.
Merchant, J. H., (A) 4 Hamilton St., Binghamton, N. Y.
Merrill, G. E., (S) 675 Amherst St., Palo Alto, Calif.
Middlecoff, H. W., (S) 2321 Dwight Way, Berkeley, Calif.
Miller, H. M., (A) 67 Park Terrace East, New York, N. Y.

Miller, H. H., 3519-5th Ave., Pittsburgh, Pa.
Miller, J. M., Jr., (A) 1429 Madison Ave. N.W., Washington, D. C.
Millspaugh, F. W., (A) 2706 East Dr., Fort Wayne, Ind.
Mischler, W. D., (A) Bell Telephone Laboratories, Inc., 180 Varick St., New York, N. Y.
Molero, F. F., (A) Calle Ancon 5383, Buenos Aires, Argentina
Moore, K. A., (S) Route 1, Box 377, Kirkland, Wash.
Moreira, A. L., (A) Rua Morais E. Silva 78, (Engenho Velho), Rio de Janeiro, Brazil
Moriwaki, Y., (A) 24 Maruyama-Tyo, Sibuya-Ku, Tokyo, Japan
Morrison, D. O., (A) 171 S. Broadway, Mount Clemens, Mich.
Moses, A. J., (J) 550—79th St., Brooklyn, N. Y.
Murphy, G., (S) 454 Webster St., Bethlehem, Pa.
Murphy, H. B., (A) 513 S. Naches Ave., Yakima, Wash.
Murray, J. W., (S) 206 St. James St., London, Ont., Canada
Nand, P., (A) India Meteorological Department, Agra, U.P., India
Naylor, R. W., (S) 45 Rosemount Ave., Toronto, Ont., Canada
Neill, W. E., (A) 311 S. 13th St., Philadelphia, Pa.
Nelson, D. E., (S) 1007 W. Springfield St., Urbana, Ill.
Newman, G., (S) 4755 Brooklyn Ave., Seattle, Wash.
Nicholas, J. C., (S) 627 Northwestern Ave. Ames, Iowa
Nickell, D. H., (S) 768 Williams St., Atlanta, Ga.
Nims, A. A., Jr., (A) 200 S. Rolling Rd., Catonsville, Md.
Norimatsu, T., (S) c/o Waichi Yamamoto, 472 Kitasenzoku Omori, Tokyo, Japan
Nunner, R. J., (S) Box 189-D, Route 3, Port Orchard, Wash.
O'Brien, J. A., (S) 88 Park St., Orono, Maine
Okochi, M., (S) 17 Yanakashimizucho Shitaya, Tokyo, Japan
Osborne, W. D., (A) Radiophysics Laboratory, University Grounds, Sydney, N.S.W., Australia
Ota, M., (S) 546 Hatsudai, Yoyogi, Shibuya-ku, Tokyo, Japan
Ozono, K., (S) c/o Shibuta, 364 Sanchome, Mabashi, Suginami-ku, Tokyo, Japan
Packard, M. E., (S) 13 Edgewood Way, Corvallis, Ore.
Pagacz, J. R., (A) 509 Ralston St., Flint, Mich.
Paramasivayya, G. S., (A) Lingaraj College, Belgaum, India
Patterson, G. K., (S) Box 2645, Stanford University, Calif.
Pearson, R. J., (S) 420 E. High St., Portland, Ind.
Pelitsch, E. J., (S) 63-12 Forest Ave., Ridgewood, L.I., N. Y.
Petersen, E. L., (A) 2509 Cleveland St., Kansas City, Mo.
Peterson, R. A., (A) 98-76 Queens Blvd., Forest Hills, L.I., N. Y.
Petree, B., (S) Battenfeld Hall, Lawrence, Kan.
Petrucelly, V., Jr., (S) 2250 McDonald Ave., Brooklyn, N. Y.
Pickarski, S., (J) 60 N. Hamilton St., Poughkeepsie, N. Y.
Potter, J. T., (A) 9 N. Church St., Schenectady, N. Y.
Preininger, L., Jr., (S) 914 Ridge Ave., McKees Rocks, Pa.
Prickett, T., (S) 342 Knapp Hall, Texas Technological College, Lubbock, Tex.
Pride, E. W., (A) c/o Bendix Radio Corporation, 920 E. Fort Ave., Baltimore, Md.
Queen, W. H., (A) 645 Maple Ct., San Leandro, Calif.
Quentin, C. F., (A) 1949 Grande Ave., Cedar Rapids, Iowa
Radford, W. H., (A) Massachusetts Institute of Technology, Cambridge, Mass.
Rambo, S. I., (A) 2519 Wilkens Ave., Baltimore, Md.
Rao, N. S. S. N., (A) c/o S. Srinivasa Rao, 10 Ground Floor, "Sukh Nivas," Matunga, Bombay 19, India
Redfield, A. D., (A) 2034 Meryton Pl., Cincinnati, Ohio
Regen, R. R., (A) 55 W. 42nd St., New York, N. Y.
Reid, J. D., (A) Box 67, Mount Healthy, Ohio
Reimers, W., (A) 1149 Fteley Ave., New York, N. Y.
Reynolds, M. T., (S) 134 S. Grant St., West Lafayette, Ind.
Robinson, D., III, (S) 232 Witherspoon, Princeton, N. J.
Rolfs, J. C., (S) 433 Hector Ave., Metairie, La.
Rosenthal, A. L., (S) 62 Fairview Blvd., Toronto, Ont., Canada
Runyon, C. R., III, (A) Hastings House, Hastings-on-Hudson, N. Y.
Saad, T. S., (S) 15 Highview Ave., West Roxbury, Mass.
Salabes, M. K., (S) 1311 Lafayette Pkwy., Williamsport, Pa.
Salerno, A., (S) 150-55 Hoover Ave., Jamaica, L.I., N. Y.
Sander, H. F., (A) 416—7th Ave., Pittsburgh, Pa.
Schmidt, R. B., (S) 512 E. Springfield Ave. Champaign, Ill.
Schoonover, H. D., (S) 1392 Midway Pkwy., St. Paul, Minn.
Schreiber, R. E., (S) 500 Sunset Dr., Manhattan, Kan.
Schreiner, L. W., (S) 1405—5th St., S.E., Minneapolis, Minn.
Schrock, N. B., (S) 337 Kellogg Ave., Palo Alto, Calif.
Schultz, M. A., (A) 3802 Belle Ave., Baltimore, Md.
Schwaikert, W. H., (A) c/o American Telephone & Telegraph Co., 32—6th Ave., New York, N. Y.
Scott, C. L., (S) 4533-2nd Ave. S., Minneapolis, Minn.
Segall, H., (S) 112 W. 15th St., New York, N. Y.
Seshadri, B., (A) Department of Electrical Technology, Indian Institute of Science, Hebbal, Bangalore, India
Settle, N. C., Jr., (A) 3553 Asbury St., Dallas, Tex.
Shuler, M. H., (S) 822—7th St., S.E., Minneapolis, Minn.
Simson, L. H., (A) 1918 Kirkwood Rd., Arlington, Va.
Singel, J. B., (A) 114 Oak Dr., Catonsville, Md.
Skerry, C. E., (A) c/o Bahrein Petroleum Co., Ltd., Bahrein Islands, Persian Gulf
Smith, C. W., (A) 421 West Ave., Elyria, Ohio
Smith, N. L., Jr., (S) 537 Leverette St., Fayetteville, Ark.
Spehr, C. J., (S) 218 E. 5th St., Hoisington, Kan.
Stiles, R. G., (S) Department of Physics, Lehigh University, Bethlehem, Pa.
Stoecker, W. K., (A) 1156 Cedar Ave., Cincinnati, Ohio
Stover, E. B., Jr., (S) 2317 Sabine St., Austin, Tex.
Strempel, E. L., (S) 330 The Parkway, Ithaca, N. Y.
Sturtevant, C. E., (S) 625 University Ave., Ithaca, N. Y.
Szegho, C. S., (A) c/o Gaumont British Picture Corp., 1600 Broadway, New York, N. Y.
Tabeling, R. H., Jr., (S) Box 2222, Gainesville, Fla.
Takano, H., (S) c/o Jomogakusha 600, Kyodo-cho, Setagaya-ku, Tokyo, Japan
Taylor, A. H., (A) 2910 Pennsylvania Ave. S.E., Washington, D. C.
Te Hsien, P., (S) c/o Electrical Engineering Department, Stanford University, calif.
Theodore, C., (S) 109 N. Hayworth Ave., Los Angeles, Calif.
Thompson, A. G., (S) 606 S. 6th Ave., Bozeman, Mont.
Treanor, P. A., (A) 529 E. 22nd St., Brooklyn, N. Y.
Treufitt, E. V., (A) British Air Commission, 15 Broad St., New York, N. Y.
Turner, F. T., (A) 220—72nd St., Brooklyn, N. Y.
Ulman, L. J., (S) 420 Oak St., S.E., Minneapolis, Minn.
Valentine, J. M., (A) 150-03—85th Dr., Jamaica, L.I., N. Y.
Valentine, R. W., (A) 8 St. George Rd., Great Neck, L.I., N. Y.
Vorperian, H., (A) 1745 Bathgate Ave., New York, N. Y.
Ward, W. C., (S) 231 Manor Rd., E., Toronto, Ont., Canada
Warner, J., (A) 5 Wallace St., Charing Cross, Sydney, N.S.W., Australia
Washino, M., (S) 3-2467 Okumachi Arakawa, Tokyo, Japan
Wataki, S., (S) 893 Sanchome, Kamikitazawa, Setagaya-ku, Tokyo, Japan
Weaver, C. S., (S) Bluestone, W. Va.
Weber, D., (S) 2340 Seabury Ave., Minneapolis, Minn.
Weden, W. E., (S) 4650 Park Ave., S., Minneapolis, Minn.
West, C. F., (S) Box 152, Sunol, Calif.
West, G. P., (S) Box 152, Sunol, Calif.
Wheatland, D. P., (A) Physics Laboratories, Harvard University, Cambridge, Mass.

Whinnery, J. R., (A) 12 N. Ferry St., Schenectady, N. Y.
White, G. E., (S) 273 Beacon St., Boston, Mass.
White, M. I., (A) 763 Broadway, Marshfield, Ore.
Whitford, J. R., (S) 30 Salem St., Wakefield, Mass.
Whyte, J. R., (S) 74 Summerhill Ave., Toronto, Ont., Canada
Wickersham, W. H. W., (A) 45 River Rd., Passaic, N. J.
Wighton, J. J., (A) c/o CBR, Vancouver Hotel, Vancouver, B. C., Canada
Wilson, M. S., (A) 1725 Highwood Pl., S.E., Washington, D. C.
Wilson, W. R., (S) 1405—5th St., S.E., Minneapolis, Minn.
Wissel, F. A., (A) Box 432-B, Route 14, Westwood, Cincinnati, Ohio
Woodcock, A. J., (A) 641 S. Serrano St., Los Angeles, Calif.
Yamada, R., (S) c/o Kiyoshi Murobushi, 2-24 Hakozakicho Nihonbashi, Tokyo, Japan
Young, E. W., (S) 9566 Manor Ave., Detroit, Mich.
Zimmermann, E. R., (A) Juncal 2144, Buenos Aires, Argentina
Zweig, F., (S) 10 Hillhouse Ave., New Haven, Conn.

Incorrect Addresses

Listed below are the names and last-known addresses of members of the Institute whose correct addresses are unknown. It will be appreciated if anyone having information concerning the present addresses of the persons listed will communicate with the Secretary of the Institute.

Allen, Hugh Silas, Jr., 1720 Pacific Ave., San Francisco, Calif.
Akmentins, Aleksanders, Rigas Galvenais Telegrafs, Riga, Latvia.
Ames, Everett K., 116 Frederick St., Apt. 35, San Francisco, Calif.
Anderson, Robert La Verne, 616 Lincoln Ave., Apt. 308A, St. Paul, Minn.
Anderson, Wm. A., Pan American Airways, Treasure Island, San Francisco, Calif.
Arthur, Garfield M., Beta Theta Pi House, Orono, Me.
Bell, J. Steven, 8076 Wisconsin, Detroit, Mich.
Benner, Howard J., 240 Norfolk St., Springfield, Mass.
Bluenthal, Herbert, Jr., 902 Concord Ave., Wilmington, Del.
Bryan, William J., 2717 Orchard Ave., Los Angeles, Calif.
Burnett, E. M., Jr., Box 169, Georgia Institute of Technology, Atlanta, Ga.
Byer, Earle D., U. S. Naval Radio Station, Summit, Canal Zone.
Cameron, John A., 200 Franklin St., Bloomfield, N. J.
Clyne, William E., 3815 S.E. 28th Ave., Portland, Ore.
Conn, Jesse Thereon, 4226—11th Ave. N.E., Seattle, Wash.
De Silva, H. B. F., Wireless Station, Welikada, Colombo, Ceylon.
Dickinson, Wesley E., Box 55, Dilworth, Minn.
Dion, Alfred J., Battery Service Station, 680 St. Vallier St., Quebec, P. Q., Canada.
Dorrington, William H., 137 Colney Hatch Lane, London, N. 10, England.
Dorte, Philip N., 17A Fitzgeorge Ave., London, W.12, England.
Dyson, H. R., 111 East Park Ave., Oaklyn, Audubon, N. J.
Erdman, Donald C., 3514-D Riverside Dr., Burbank, Calif.
Erickson, Lennart Gottfrid, 1324 Mission St., San Francisco, Calif.
Evans, Donald J., 325 University Station, Grand Forks, N. D.
Feldmanis, Nicholas, Valsts Elektrotechniska Fabrika, Brivibas Gatve 19, V.E.F., Riga, Latvia.
Fernandez, Dan T., 6709 N. Glenwood Ave., Chicago, Ill.
Fiegna, Edmondo, Piazza Maresciallo Giardino, 1-Int. 25, Roma, Italy.
Franklin, Mortimer, 600 South Michigan Ave., Chicago, Ill.
Friend, Halton H., Navy Dept., Ordnance Bureau, Section ND, Washington, D. C.
Gates, Howard A., 7760 La Salle Blvd., Detroit, Mich.
Giovanini, Frank, Roslyn, Wash.
Granberry, Doyle, 6210 S. Kimbark, Chicago, Ill.
Gray, Richard I., 3732 Columbus Dr., Baltimore, Md.
Hardenbergh, George A., 1301 E. California St., Pasadena, Calif.
Heard, Burton E., 3008 Rosedale, Dallas, Tex.
Hess, Howard M., Wayne University, 4841 Cass Ave., Detroit, Mich.
Jackson, Hubert E. A., Carlton Bank, Sawood, Stainland, Near Halifax, Yorks, Great Britain.
Jayasekera, Don P., Department of Posts & Telegraphs, Ceylon Central Telegraph Office, Colombo, Ceylon.
Johnson, Earl M., 2350 Stratford, Cincinnati, Ohio.
Jukna, Robert, Valsts Elektrotechniska Fabrika, Brivibas Gatve Nr. 19, Riga, Latvia.
Kennedy, E. Donald, 2026 East 107th St., Cleveland, Ohio.
Lamarsh, Joseph E., 4432 Dover St., Chicago, Ill.
Landsberg, Klaus U., 325 W. 82nd St., New York, N. Y.
Lawsine, Leo, 826 S. Wabash Ave., Chicago, Ill.
Lebert, Andrew W., 43-25 169th St., Flushing, L. I., N. Y.
Levering, Charles A., 515 E. 15th St., Santa Ana, Calif.
Mavy, Elsmore, 1529 Webster St., Oakland, Calif.
Mayder, Wesley, 2718 Myrtle St., Oakland, Calif.
Mayer, Rollins H., 841 N. Madison, Pasadena, Calif.
McCurdy, Robert G., 6720 Locust St., Kansas City, Mo.
McIntosh, Frank H., Shell Rd., Mill Valley, Calif.
McKay, William M., c/o United Air Lines, Inc., Chicago, Ill.
McLeish, C. William, 179 York St., Ottawa, Ont., Canada.
McNeill, Louis P., Box 3313, University of Kentucky, Lexington, Ky.
Moon, Benjamin F., 1217 East Windsor Rd., Glendale, Calif.
Morris, Lloyd P., 3802 N. Seeley Ave., Chicago, Ill.
Nagai, Kenzo, 837 Ramona St., Palo Alto, Calif.
Nessell, Harry Hirsh, 312—16th Ave., S.E., Minneapolis, Minn.
Newberry, Denny A., Flag Radio, North Island, San Diego, Calif.
Nicholson, Taft, Chief Plant Eng., KOMA, 1450 N.W. 42nd St., Oklahoma City, Okla.
Nunan, Craig S., 2627 Ridge Rd., Berkeley, Calif.
Ordung, Philip Franklin, 1224—4th St., Brookings, S. D.
Price, Wayne E., 34-60—112 St., Corona, L. I., N. Y.
Rich, Edwin S., 2200 Prospect Ave., Cleveland, Ohio.
Rickard, Charles E., 5 Baldwyn Gardens, Acton, London, W.3, England.
Rockwell, Paul, c/o Pan-American Grace Airways, Lima, Peru.
Rowe, Victor G., 6747—5th Ave., Los Angeles, Calif.
Sabine, Leon, 4019 Queen Ave., W., Minneapolis, Minn.
Sedon, Paul H., 436 W. 5th St., Greenville, N. C.
Shermund, Ralph C., 2300—16th Ave., San Francisco, Calif.
Slee, John A., 92A Holland Rd., London W.14, England.
Sokoloff, Paul W., Airplane & Marine Direction Finder Corporation, Clearfield, Pa.
Storm, Hans Otto, 559 Cowper St., Palo Alto, Calif.
Stupp, Russell W., 1013 Emerald Ave., Collingswood, N. J.
Sugnet, Robert F., 112 Howard Hall, Notre Dame, Ind.
Swinton, Kurt R., 4, Leinster Gardens, London, W. 2, England.
Tatibana, Mituo, No. 1, 611 Koyama-Tyo, Itabasi-Ku, Tokyo, Japan.
Trebby, Francis J., 704 W. 6th St., Apt. 404, Spokane, Wash.
Turitzin, N., 217 North Ingalls, Ann Arbor, Mich.
Turner, Charles W., 6329½ Franklin Ci., Hollywood, Calif.
Veverka, R. E., 3071 S. 32nd St., Omaha, Neb.
Whitmore, Jack, Hotel Lincoln, 8th Ave., New York, N. Y.
Woodward, John D., c/o WBOE, Cleveland Board of Education, 1380 E. 6th St., Cleveland, Ohio.
Woodward, Oakley M., Jr., Seismograph Service Corp., Kennedy Bldg., Tulsa, Okla.
Ziemelis, Valdis, Radio Department, c/o Valsts Elektrotechniska Fabrika, Brivibas Gatve 19, Riga, Latvia.

Books

Elements of Acoustical Engineering, by H. F. Olson

Published by D. Van Nostrand Co., 250 Fourth Ave., New York, 1940. xv+335 pages+8-page index. 198 figures. $6\frac{1}{2}\times9\frac{1}{4}$ inches. Price $6.00.

"The book includes the current acoustic practices in radio, phonograph, sound motion pictures, public address, sound reenforcing and sound measurements." (From the author's preface.) Its contents, by chapters, are: I Sound waves; II Acoustical Radiating Systems; III Mechanical Vibrating Systems; IV Electrical, Mechanical and Acoustical Analogies; V Acoustical Elements; VI Driving Systems; VII Direct Radiator Loud Speakers; VIII Horn Loud Speakers; IX Microphones; X Miscellaneous Transducers; XI Measurements; XII Architectural Acoustics and the Collection and Dispersion of Sound; and XIII Speech, Music and Hearing.

The chapters most directly concerned with loud-speakers and microphones, II, VII, VIII, IX, and parts of XI, contain an excellent up-to-date treatment of the two most important groups of electroacoustic transducers. They will be as highly appreciated by specialists as by general engigeering readers. These chapters acount for some 175 of the 335 pages in the text. The numerous other topics are necessarily less completely treated in the remaining pages. The mathematical knowledge presumed on the part of the reader is quite modest, generally no more than that which goes with the elementary theory of alternating currents. Many of the basic equations are given without derivation, which is consistent with the compactness of the book and with the desire not to overtax the reader's mathematics. The method of acoustic impedances and the electrical circuit analogies are used throughout.

The physics of the situation is not always stressed, especially where the delicate matter of transmission from the regime of "lumped" constants to that of wave motion is concerned. For example, acoustic capacitance is defined as a cavity having linear dimensions small compared to the wavelength. The unaided reader might have a little trouble in assigning the proper order of magnitude to "small"; in a cylindrical volume the "smallness" requirements in the radial and axial directions are not the same. He might again have difficulty in deciding on the frequency range for which the formulas given for "small" tubes and slits is valid. In addition to the criteria given in the text this would involve the notion of viscous wavelength.

The standard of the illustrations and graphs is high and the scales of co-ordinates chosen make the most of the space available. The book is thoroughly commended to any one seriously interested in the design, performance, and testing of loud speakers and microphones.

L. J. Sivian
Bell Telephone Laboratories, Inc.
New York, N. Y.

Contributors

H. A. Wheeler

Harold Alden Wheeler (A'27, M'28, F'35) was born at St. Paul, Minnesota, on May 10, 1903. He received his B.S. degree in physics at George Washington University in 1925. From 1925 to 1928 he took postgraduate work in the physics department of Johns Hopkins University and was a lecturer there from 1926 to 1927. From 1921 to 1922 he was an assistant in the Radio Section of the National Bureau of Standards. In 1923 Mr. Wheeler became an assistant to Professor Hazeltine and in 1924 he enetered the Hazeltine Corporation and the Hazeltine Service Corporation as an engineer. He is a member of Sigma Xi.

❖

W. G. Dow

W. G. Dow (M'39) was born at Faribault, Minnesota, on September 30, 1895. He received the B.S. degree in electrical engineering in 1916, and the E.E. degree in 1917 from the University of Minnesota; also the M.S.E. degree in electrical engineering from the University of Michigan in 1929. Mr. Dow served as a commissioned officer in the United States Engineers during the World War, chiefly in sound ranging work. For six years following the close of the war he was engaged in commercial activities with the Westinghouse Electric and Manufacturing Co., and since 1926 has been a member of the faculty of the University of Michigan, where he is at present associate professor of electrical engineering, in charge of instruction in electronics. He is the author of a textbook "Fundamentals of Engineering Electronics," published in 1937 by John Wiley and Sons, Inc., and of several technical papers on electronic subjects. He is a member of Eta Kappa Nu, Tau Beta Pi, Sigma Xi, the American Institute of Electrical Engineers, Society for the Promotion of Engineering Education, and the American Welding Society, A.W.S.

❖

Heinz E. Kallman

Heinz E. Kallmann (A'38) was born on March 10, 1904, at Berlin, Germany. He received his Ph.D. degree from the University of Goettingen in 1929. From 1929 to 1934, Dr. Kallmann was a research engineer in the laboratories of the C. Lorenz A. G., and from 1934 to 1939 he was an engineer in the Research and Design Department of Electric and Musical Industries, Ltd. In 1940 Dr. Kallmann joined the New York laboratory staff of Scophony Television, Ltd.

❖

Horatio Wellington Lamson (F'33) was born on January 2, 1893, at Somerville, Massachusetts. He received the S.B. degree in physics in 1915 from Massachusetts Institute of Technology and the A.M. degree in physics in 1917 from Harvard University. Mr. Lamson was a member of the United States Naval Reserve Force during the World War and Assistant to Dr. G. W. Pierce in the development of antisubmarine devices. Subsequently he became a physicist in the United States Navy Department. Since 1921 he has been a research engineer with the General Radio Company. Mr. Lanson is a Fellow of the Acoustical Society of America and a Member of the American Institute of Electrical Engineers.

H. W. Lamson

Proceedings of the I·R·E

Published Monthly by
The Institute of Radio Engineers, Inc.

VOLUME 28—1940

The Institute of Radio Engineers, Inc.
330 West 42nd Street
New York, N. Y.

GENERAL INFORMATION

The Institute

The Institute of Radio Engineers serves those interested in radio and allied electrical-communication fields through the presentation and publication of technical material.

Membership has grown from a few dozen in 1912 to more than five thousand. Practically every country in the world is represented in our roster of membership, with approximately a quarter of it located outside of the United States. There are several grades of membership, depending on the qualifications of the applicant, with dues ranging from $3.00 per year for Students to $10.00 per year for Members.

PROCEEDINGS, Standards Reports, and any other published material are sent to members without further payment.

The PROCEEDINGS

The PROCEEDINGS has been published without interruption from 1913 when the first issue appeared. Almost 2000 technical contributions have been included in its pages and portray a currently written history of developments in both theory and practice. The contents of every paper published in the PROCEEDINGS are the responsibility of the author and are not binding on the Institute or its members. Material appearing in the PROCEEDINGS may be reprinted or abstracted in other publications on the express condition that specific reference shall be made to its original appearance in the PROCEEDINGS. Illustrations of any variety may not be reproduced, however, without specific permission from the Institute.

Subscriptions

Annual subscription rates for the United States of America, its possessions, and Canada, $10.00; to college and public libraries when ordering direct, $5.00. Other countries, $1.00 additional.

Back Copies

The Institute endeavors to keep on hand a supply of back copies of the PROCEEDINGS for sale for the convenience of those who do not have complete files. However, some issues are out of print and cannot be provided.

Standards

In addition to the material published in the PROCEEDINGS, Standards on Electroacoustics, Electronics, Radio Receivers, and Radio Transmitters and Antennas were published in 1938. These are available to members free of charge as long as the supply lasts; others may purchase them for fifty cents each.

CONTENTS OF VOLUME 28—1940*

VOLUME 28, NUMBER 1, JANUARY, 1940

VOLUME 28, NUMBER 2, FEBRUARY, 1940

VOLUME 28, NUMBER 3, MARCH, 1940

VOLUME 28, NUMBER 4, APRIL, 1940

VOLUME 28, NUMBER 5, MAY, 1940

VOLUME 28, NUMBER 6, JUNE, 1940

* A cumulative index of the same type as this but covering the PROCEEDINGS from its start to the end of 1936 is available at $1.00 per copy. The 1937, 1938, and 1939 Indexes will be found in the December, 1937, December, 1938, and December, 1939, issues of the PROCEEDINGS respectively.

Volume 28, Number 7, July, 1940

Volume 28, Number 8, August, 1940

Volume 28, Number 9, September, 1940

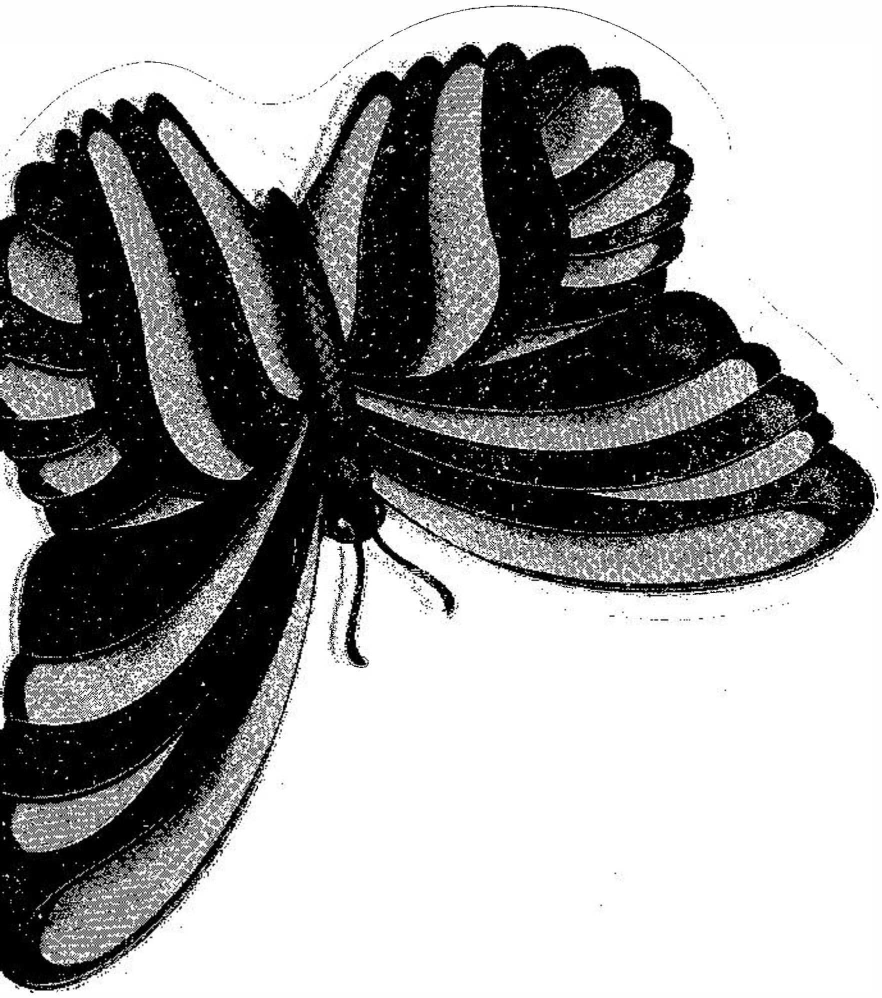
Volume 28, Number 10, October, 1940

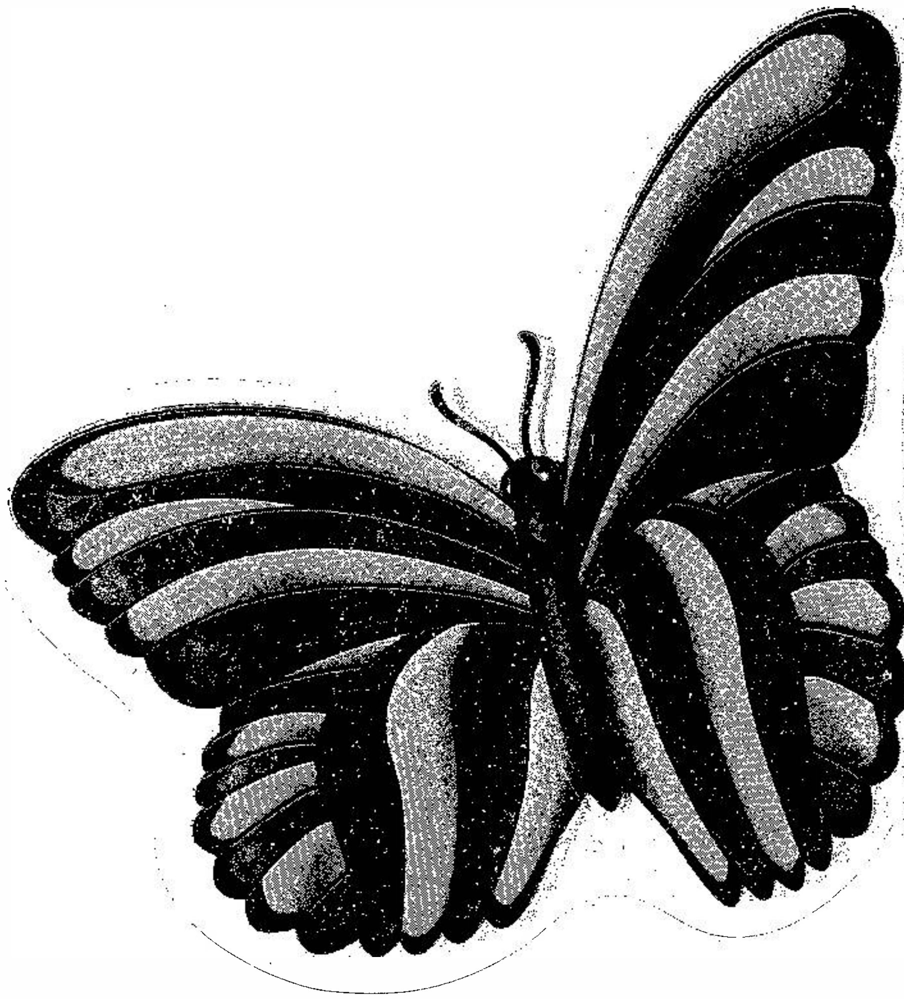
Volume 28, Number 11, November, 1940

VOLUME 28, NUMBER 12, DECEMBER, 1940

AUTHOR INDEX

Numbers refer to the chronological list. **Bold-face** type indicates papers and *italics* refer to books and book reviews.





AUTHOR INDEX

Numbers refer to the chronological list. **Bold-face** type indicates papers and *italics* refer to books and book reviews.

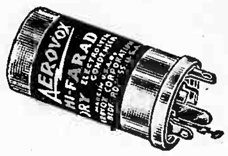
SUBJECT INDEX

NONTECHNICAL INDEX

A prime factor . . .

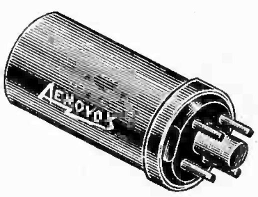
Choice

• Universal - mounting condensers and inverted-mounting types, with leads or soldering-lug terminals. Also strap-mounting and stud-mounting types. In full-sized and ultra-compact sizes.



• Prong - base midgets, with fibre mounting washers (for insulated can) and metal (grounded can) washers. Improved construction eliminates drawbacks heretofore associated with this type.

• Midget tubular electrolytics in hermetically-sealed metal cans and in cardboard cases, where space and price are limited.



• Latest plug-in type, developed and introduced by Aerovox, to facilitate testing and replacing of condensers in continuous-service equipment .

• Some two dozen different types of Aerovox electrolytics are available to you as *standard* items. And from this widest selection of cans, sizes, terminals, mountings, capacities, combinations, voltages, *special* types may be readily and economically evolved.

Such exceptional choice has far-reaching implications. It means that you can pick out types properly fitted to your purposes. Certainly you need not alter or improvise or curtail your plans in the absence of those limitations that would be imposed by a lesser choice of electrolytics.

It is this exceptional Aerovox choice of types, combined with rare application experience, that can and usually does spell substantial economies for your assemblies.

Therefore, assuming that everything else is equal—quality, reputation, delivery, price—don't overlook that prime consideration: CHOICE of types.

A.A.E.* Service . . .

Your application *as a whole*—that is the approach of Aerovox sales engineers to your condenser requirements. That is why A.A.E.* service is constantly effecting substantial economies for users of Aerovox condensers. Let us study your condenser needs. Our engineers will gladly submit recommendations quite as well as specifications, quotations, samples.

Latest engineering data on all types of condensers, available on request.

* *Aerovox Application Engineering*

AEROVOX CORPORATION

New Bedford, Mass.

Sales Offices in All Principal Cities

POSITIONS OPEN

●

The following positions of interest to I.R.E. members have been reported as open on February 5. Make your application in writing and address to the company mentioned or to

Box No. . . . . . .
PROCEEDINGS of the I.R.E.
330 West 42nd Street, New York, N.Y.
Please be sure that the envelope carries your name and address

RECEIVER DESIGN ENGINEERS

A large midwestern radio-receiver manufacturer has openings for experienced automotive and household radio-receiver-design engineers. Applicants should state education, experience, and give references. Our own employees know of these openings. Box 233.

OPERATOR—FOREIGN SERVICE

Radio operator with some experience in electrical installation work. Should have at least a second-class radio-telegraph and first-class radio-telephone licenses and be able to do his own service work. Two year contract. Location: foreign. Box 234.

RADIO OR ELECTRICAL ENGINEERS

Radio or electrical engineers wanted for permanent positions on the staff of the Washington Institute of Technology. Applications held confidential. Address the Washington Institute of Technology, McLachlen Building, Washington, D.C.

ENGINEERING DRAFTSMEN

Engineering draftsmen in various optional fields (including radio) are urgently needed by the United States Government. Salaries range from $1,620 to $2,600 per year. Applications will be accepted until December 31, 1941. For further information and application forms, consult the Secretary of the Board of U. S. Civil Service Examiners at any first- or second-class post office.

TRANSFORMER-DESIGN ENGINEER

An eastern manufacturer has an opening for an engineer who is experienced in transformer design and familiar with the properties of magnetic materials. Box 235.

HIGH-SPEED RADIO-EQUIPMENT OPERATORS

To meet the urgent need for high-speed radio-equipment operators for the National Defense Program, the United States Civil Service Commission has announced that it will accept applications to fill these positions until further notice. One year

(Continued on page vii)

Associate membership affiliates you with the Institute and brings you the PROCEEDINGS each month as well as notices of meetings held near you.

(Typewriting preferred in filling in this form) No..........

RECORD OF TRAINING AND PROFESSIONAL EXPERIENCE

Name ....
(Give full name, last name first)

Present Occupation ....
(Title and name of concern)

Business Address ....

Home Address ....

Place of Birth .... Date of Birth .... Age ......

Education ....

Degree ....
(College) (Date received)

TRAINING AND PROFESSIONAL EXPERIENCE
(Give dates and type of work, including details of present activities)

Record may be continued on other sheets this size if space is insufficient.

Receipt Acknowledged .... Elected .... Notified ....

Requirements—For Associate membership, an applicant must be at least twenty-one years of age, of good character, and be interested in or connected with the study or application of radio science or the radio arts.

Sponsors—Three sponsors who are familiar with the work of the applicant must be named. Preferably these should be Associates, Members, or Fellows of the Institute. In cases where the applicant is so located as not to be known to the required number of member sponsors, the names of responsible nonmember sponsors may be given.

Dues—Dues for Associate membership are six dollars per year. The entrance fee for this grade is three dollars and should accompany the application.

Other Grades—Those who are between the ages of eighteen and twenty-one may apply for Junior grade. Student membership is available to full-time students in engineering or science courses in colleges granting degrees as a result of four-year courses. Member grade is open to older engineers with several years of experience. Information on these grades may be obtained from the Institute.

POSITIONS OPEN

(Continued from page iv)

of experience as radio operator in commercial or Government radio communications work is required. Although training in radio operation at a service school may be substituted for this experience, all applicants must have had three months' experience in the operation of high-speed radio-communication equipment such as transcribing to typewriter syphon recorder tape and transmitting messages by hand or bug. For further information and application forms, consult the Secretary of the Board of U. S. Civil Service Examiners at any first- or second-class post office.

RADIO ENGINEER

There is an opening in an eastern laboratory for an engineer with a good general knowledge of radio-receiver-design and tube-application problems. Good education required. Will be expected to work on his own with only general supervision. Box 236.

TUBE ENGINEER

A manufacturer of power tubes is looking for a man who is experienced in electronic tubes and mercury-vapor rectifiers. He also must have a full knowledge of vacuum theory. Box 237.

RADIO INSPECTOR POSITIONS

Examinations have been announced to fill the positions of radio inspector in the Federal Communications Commission at $2,600 per year and of assistant radio inspector in various departments at $2,000 per year. Applications must be filed not later than March 6 and March 10. For further information, application forms, etc., consult the Secretary of the Board of Civil Service Examiners at any first- or second-class post office.

PHYSICISTS

The Civil Service Commission has modified the requirements for the positions of physicist (various grades) and extended the application deadline to December 31, 1941. These positions pay from $2,600 to $5,600 per year. For further information, consult the Secretary of the Board of Civil Service Examiners at any first- or second-class post office.

Attention Employers . . .

Announcements for "Positions Open" are accepted without charge from employers offering salaried employment of engineering grade to I.R.E. members. Please supply complete information and indicate which details should be treated as confidential. Address: "POSITIONS OPEN," Institute of Radio Engineers, 330 West 42nd Street, New York, N.Y.

The Institute reserves the right to refuse any announcement without giving a reason for the refusal

DISPLAY ADVERTISERS

Current Literature

New books of interest to engineers in radio and allied fields—from the publishers' announcements. A copy of each book marked with an asterisk (*) has been submitted to the Editors for possible review in a future issue of the Proceedings of the I.R.E.

* TELEVISION RECEIVING EQUIPMENT. BY W. T. COCKING. New York: Nordeman Publishing Company, Inc., 1940. viii+295+3 index pages, illustrated, $4\frac{3}{4}\times7\frac{1}{4}$ inches, cloth. $2.25.

* FESSENDEN BUILDER OF TOMORROWS. BY HELEN M. FESSENDEN. New York: Coward-McCann, Inc., October, 1940. vi+362 pages, 6×9 inches, cloth. $3.00.

* BROADCASTING IN INDIA. BY LIONEL FIELDEN, Controller of Broadcasting, All India Radio. Delhi: Manager of Publications, 1940. xiv+212+18 index pages, illustrated $6\frac{1}{2}\times10$ inches, paper board. Rupees 3 or 5 shillings.

ELECTRONIC PROCESSES IN IONIC CRYSTALS. BY N. F. MOTT and R. W. GURNEY. New York: Oxford University Press, July, 1940. 287 pages, illustrated, $6\times9\frac{1}{4}$ inches cloth. $5.50.

* TELEVISION TODAY AND TOMORROW. BY SYDNEY A. MOSELEY and H. J. BARTON-CHAPPLE. New York: Pitman Publishing Corporation, 1940. xix+175+3 index pages, illustrated, $5\frac{1}{2}\times8\frac{1}{2}$ inches cloth. $3.00.

* MODERNE MEHRGITTER-ELEKTRONENROHREN (Modern Multi-grid Electron Tubes). Second Edition. BY M. J. O. STRUTT. Berlin: Julius Springer, 1940. viii+278+5 index pages, illustrated, 6×9 inches, cloth. 25.80 rm.

* GEOPHYSICAL EXPLORATION. BY C. A. HEILAND, Professor of Geophysics, Colorado School of Mines. New York: Prentice-Hall, Inc., November, 1940. xiii+963 +48 index pages, illustrated, 6×9 inches, cloth. $10.00.

* SPECIFYING A POLICE RADIO COMMUNICATION SYSTEM. New York; National Electrical Manufacturers Association May, 1940. 24 pages, $8\times10\frac{1}{2}$ inches, paper. 50 cents.

*THE METER AT WORK. BY JOHN F. RIDER, New York: John F. Rider, Inc., 1940. viii + 152 pages, illustrated, $5\frac{1}{4}\times8\frac{1}{2}$ inches, cloth. $1.25.

*ELECTROMAGNETIC THEORY. BY J. A. STRATTON, Associate Professor of Physics, Massachusetts Institute of Technology. New York: McGraw-Hill Book Company 1941. xv + 608 + 7 index pages, illustrated, 6 × 9 inches, cloth. $6.00

PRINTED IN U.S.A.