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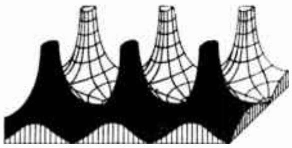
Poles and Zeros



The Members Speak. It has been reported elsewhere that the revised Constitution for the IRE has been approved by the membership and went into effect as of April 20, 1959. We wish here, however, to add our voice in praise of the interest in IRE affairs which was shown by our membership in this balloting.

At the time ballots were distributed the IRE had 40,814 voting members—Fellows, Senior Members, Members, and a few hundred Voting Associates. Almost fifty per cent—19,805—showed their desire to participate in IRE affairs by returning their ballots. This is a phenomenal record among geographically-scattered membership organizations, and further emphasizes much that we have been saying in these columns concerning the dynamic and working nature of this organization with its 103 Sections, 30 Subsections, 28 Professional Groups, and 251 Professional Group Chapters, all providing opportunities for many hundreds to take part each year in organizing and operating professional and technical activities.

Of equal importance to the internal health of the organization was the fact that 96.2 per cent voted "Yes," 2.0 per cent "No," with 1.8 per cent of the ballots technically invalid. Such a vote of confidence in our Board of Directors and IRE operations in general is received by the Board and officers with thanks, and with the realization that it can also be considered as an order to keep on with the work of assuring continued IRE success in professional leadership in this uncharted part of the technical world called electronics.

The May Board Meeting. Most of the Board time on May 13 was spent in carrying out the next step in constitutional revision by adopting a set of By-Laws which would continue our present operations in a frame consonant with the new Constitution.

Beating out the details in thirty pages of semilegalistic phraseology is not easy in a large meeting, since everything from commas to redundancies to principles to semantics to inconsistencies to commas is likely to be encountered in discussion, in about that circuitous order. Many thanks are due to Secretary Haraden Pratt and a small committee of commentators who had previously worked over the document, thereby pointing out a few of the journalistic, legalistic, or lexicographic pitfalls.

Essentially it was desired that present IRE methods and operating principles be carried over to the new By-Laws, and we believe this has been accomplished. Also, in response to work of last year's Policy Advisory Committee the requirements for the various grades of membership have been rewritten and clarified, and these have been followed by the Admissions Committee and Membership Coordinator Ronald McFarlan in a revision of application and reference forms for membership. This revision is designed to simplify the work of the Admissions Committee by providing them with the particular information they desire to measure prospective members against the new membership provisions of the By-Laws. The new forms should be in use in a reasonable time.

As soon as the revised By-Laws have received the blessing of legal counsel, to insure conformity with New York law and the new IRE Constitution, they will be published in these pages for perusal by the membership.

1906 And All That. Appearing in this issue (p. 1253) is a lengthy correspondence, comment, rebuttal, and surrebuttal, on certain aspects of electronic history dating back to the invention of the triode tube. Edited by the writers themselves, this material at one time numbered some hundreds of typewritten pages, and well illustrated the uncertainties inherent in the recording of history. As long as each of us is free to use his own eyes and ears, we must realize that history can only be a collection of subjective observations and semantic interpretations, occasionally made more precise with written record—the latter again subject to the vagaries of translation of our changing languages.

This column is written not to take sides in the various matters discussed, but to comment on how much was accomplished in those early days of the triode. Having read the complete cases as presented, it would appear that, prior to Langmuir's publication of the three-halves power law in 1913, there was little general understanding or knowledge of the volt-ampere curve of a triode. Certainly the reported story of Fritz Lowenstein's negative-grid patent would support this view. Yet today any college sophomore starts at that fundamental point in his education.

The discussion of the discovery of the amplifying properties of the triode, the high-vacuum controversy, or whether the U. S. government (Army *or* Navy) had *an* airplane in 1916, all contribute to a fuller realization of our sophistication of today, of the terrific advances of which electronics has been a part for over fifty years. Be humble in reading these bits of history—could we have done so much with so little?

A New Electronic Application. The Central Florida Section publicizes the use of closed-circuit television in a Havana hotel, to permit following the progress at the roulette tables from the privacy of your own suite. Bets may be placed by telephone and charged to your bill. Such privacy could well be a method to increase business since no feelings of modesty need prevent the patron from betting his pants as well as his shirt!—J.D.R.

Arthur H. Waynick

Director, 1959–1960

Arthur H. Waynick (A'43–SM'46–F'57) was born on November 9, 1905, in Spokane, Wash. He received the B.S. and M.S. degrees in physics from Wayne University, Detroit, Mich., in 1935 and 1936, respectively. In 1937 he attended the Electronics Institute of the University of Michigan, and then studied at Cambridge University in England until the war broke out. He received the D.Sc. degree in communications engineering from Harvard University, Cambridge, Mass., in 1943.

He was employed by the Reno Radio Co., in Detroit, Mich., as a radio engineer from 1922–1935. From 1935–1937 he was an instructor in physics at Wayne University, then until 1939 a demonstrator in physics at Cambridge University, and returned to Wayne University as assistant professor of physics in 1939.

From 1940–1945 he was electronics section head of the Harvard Underwater Sound Laboratory, and for the next three years was electronics section head of the Ordnance Research Laboratory, State College, Pa. Since 1948 he has been director of the Ionosphere Research Laboratory of Pennsylvania State University in University Park, and professor and chairman of the electrical engineering department there. In 1958 he was made program director in engineering science of the National Science Foundation, Washington, D. C.

Dr. Waynick received the Navy Ordnance Development Award and the OSRD Award in 1945, the AIEE Electronics Award for 1951–1952, and the Distinguished Alumni Award from Wayne University in 1957. He was chosen to serve as a Guggenheim Fellow at Cambridge University, 1954–1955.

He has published numerous technical papers and holds a patent for a cathode-ray polar coordinant vector plotter. He is an active member and past chairman and treasurer of the U. S. National Committee of URSI, and has been chairman of the Technical Program Committee of the URSI General Assembly. In 1950, 1952, and 1954 he was a member of the U. S. delegation to the URSI General Assembly.

He is a member of the IRE Editorial Board and has served on the Professional Groups Committee and the Wave Propagation Committees and as chairman of the Professional Group on Antennas and Propagation. He is a past member of the AIEE Basic Science Committee and a member of the AIEE Electronic Instrumentation Committee; the U. S. National Committee of the IGY Panel on Rockets and Satellites and on Ionospheric Physics; the National Science Foundation Advisory Committee on Radio; the Picatinny Arsenal Scientific Advisory Council; and the American Geophysical Union Committee on Cosmic and Terrestrial Relationships. He is an associate editor of the *Journal of Geophysical Research* and a member of the honorary advisory Board of the English edition of *Elektrichestvo.*

Dr. Waynick is a member of the AIEE, the American Geophysical Union, the American Society of Engineering Education, Eta Kappa Nu, and Sigma Pi Sigma.

Scanning the Issue

Related Experiments with Sound Waves and Electromagnetic Waves (Kock, p. 1192)—This month the European journal *Acustica* is publishing a memorial issue in honor of the 60th birthday of Professor Erwin Meyer, Director of the III Physikalisches Institute of the University of Goettingen and one of the outstanding authorities in acoustics and microwaves in Europe. This paper was prepared especially for inclusion in that issue by one of Professor Meyer's former students who is himself a well-known authority in the field. In it the author draws some exceedingly interesting and important comparisons between sound waves and microwaves. Because of the unusual nature of these comparisons and in honor of Professor Meyer, arrangements have been made for simultaneous publication of the paper in this issue of PROCEEDINGS. It is, of course, no surprise that sound and electromagnetic waves behave in a generally similar manner, both being wave phenomena. However, when one considers microwaves in particular, the similarity becomes a much more striking one because the wavelength of audible sound lies in the same centimeter range. This suggests that various devices designed for microwaves, such as waveguides, lenses and radiators, should be dimensionally suited to work equally well with sound waves. This dimensional compatibility has permitted the author to conduct some unusual experiments and take some equally unusual photographs in which he uses sound waves to demonstrate various phenomena and devices that are normally associated with microwaves.

Tunnel Diodes as High-Frequency Devices (Sommers, p. 1201)—Last year Esaki reported that heavily doped junction diodes would exhibit negative resistance when a small forward bias is applied. This phenomenon has now been further investigated and a new and potentially important high frequency device has resulted which may prove capable of operating at a higher frequency than any other semiconductor device. To date it has been observed to have a switching speed of 2 millimicroseconds, to oscillate above 1 kilomegacycle, and to generate harmonics above 4 kilomegacycles. Even more intriguing is the fact that no one yet knows what the ultimate frequency limitation is, except that the theoretical limit is a good deal higher than has been achieved so far. In fact, it appears that the top frequency will be limited by technical problems of fabrication such as the maximum doping achievable or the minimum mounting impedance, or circuit difficulties in exciting the desired mode of operation, rather than any limiting time constant of the physical process within the device itself.

The Cryosar—A New Low-Temperature Computer Component (McWhorter and Rediker, p. 1207)—It has been found that when a semiconductor is subjected to very low temperatures (about 4°K), a small applied voltage will cause a rapid ionization of impurities, resulting in an avalanche breakdown. This phenomenon, known as impact ionization, is leading to new advances in cryogenic devices. Last month we saw how this breakdown process can be controlled so as to amplify millimicrosecond pulses. This month the same phenomenon is employed to produce an important new high-speed computer component. In one form the device functions like a diode; in another, it is bistable and can be used as a memory element, multivibrator or flip-flop. Most interesting of all, a large number of cryosars can be fabricated on one small wafer of germanium. Indeed, it is estimated that 200,000 cryosars could be compressed into one cubic inch, a most timely and significant statistic in view of the widespread current interest in microelectronic circuits.

Parametric Energy Conversion in Distributed Systems (Roe and Boyd, p. 1213)—This paper brings to light an important property of all parametric amplifiers of the traveling-wave type, namely, the tendency to generate harmonics of the fundamental frequency. It is an effect of first-order importance which has been entirely neglected in previous papers. For when the output wave does contain higher-order frequencies, it means that energy is being diverted from the fundamental component in order to generate the higher harmonics, with a resulting decrease in amplification at the fundamental frequency. The degree to which this effect occurs in a practical system depends on its propagation characteristics, and is a subject which will now require further study because of this paper. However, it is shown that, at best, the amount of amplification predicted earlier for traveling-wave type systems will have to be revised downward, and that in special cases a parametric device will not amplify at all, but will act as a good harmonic generator.

A Discussion of Sampling Theorems (Linden, p. 1219)—The convolution theorem of Fourier analysis has proven in the past to be an excellent tool for deriving sampling theorems involving first-order sampling of low-pass functions. The author has extended this technique to a number of other types of sampling, summarizing in a useful way previously published information on sampling theorems and presenting a unifying approach to new problems in sampling. This work will be of considerable interest to engineers in a number of fields, particularly communication and control systems engineers who deal with pulse communication and sampled data control system problems.

An Application of Piecewise Approximations to Reliability and Statistical Design (Gray, p. 1226)—A method has been developed for determining the probability that, due to gradual deterioration of system components, a specified characteristic of the system will deviate beyond acceptable limits. The method requires no special knowledge on the part of the user, and has the added advantage that it can be programmed for a computer. As one PROCEEDINGS reviewer put it: "If they will only use this paper, it could help any circuit designer secure reliable circuit performance over a reasonable period of time."

An Instantaneous Microwave Polarimeter (Allen and Tompkins, p. 1231)—By employing a tri-mode turnstile waveguide junction, the authors have succeeded in building the first simple device that can provide an instantaneous picture on a cathode-ray tube of the polarization of radio waves in the microwave region. While this is a feat primarily involving microwave techniques, this instrument will be of considerable value in the fields of antennas, communications, propagation, radar and radio astronomy.

Thin Film Magnetization Analysis (Chu and Singer, p. 1237)—A simple graphical method has been developed for analyzing complex magnetization conditions that occur in thin films of magnetic materials of the type that are becoming important in computer applications. It will prove very useful in determining hysteresis loop shapes under a variety of conditions and, in particular, in designing memory and logic devices.

Analog Computer Measurements on Saturation Currents, Admittances and Transfer Efficiencies of Semiconductor Junction Diodes and Transistors (Frei and Strutt, p. 1245)—The densities of carriers and diffusion currents in a junction device are usually calculated under the simplifying assumption that a one-dimensional structure exists. In reality, however, the problem is a three-dimensional one, and, moreover, one for which no useful solutions are known. The authors have designed an analog computer for solving the three-dimensional case and derived some curves which will be useful in improving the design of transistors. The curves also show that under certain conditions the one-dimensional approach used in the past has been giving us current densities which are too low by a factor of two.

Scanning the Transactions appears on page 1279.

Related Experiments with Sound Waves and Electromagnetic Waves*

WINSTON E. KOCK†, FELLOW, IRE

Summary—**Various analog situations in acoustics and electromagnetic waves are described. Certain higher order modes of airborne sound waves in tubes possess a transverse or polarized nature, and electromagnetic properties such as cutoff effects, polarization rotation, and circular polarization can be shown for these sound waves. Externally-guided sound waves, similar to radio waves guided by a dielectric rod, are also discussed, as are superdirective acoustic and electromagnetic arrays, space-frequency equivalence in arrays, and experiments in wave diffraction.**

Introduction

One of the areas in which emphasis has been placed in Professor Erwin Meyer's III Physikalisches Institut at the University of Goettingen is that of microwave acoustic analogs.[1] This paper discusses similar analog experiments conducted by the author during the past several years, some at the Bell Telephone Laboratories and the remainder at the Bendix Systems and Research Divisions. All have previously been reported upon orally at various meetings, but have not been published.

Lord Rayleigh stressed quite early the simultaneous treating of sound and electromagnetic wave phenomena. In one paper,[2] he begins with the sentence: "The waves contemplated may be either aerial waves of condensation and rarefaction or electrical waves propagated in a dielectric." Further, he observed that in certain problems of passage of waves through a slit and reflection from a blade the same results are obtained for sound waves or electric waves if the blade or screen is rigid or perfectly conducting, respectively. He thus established that many electromagnetic phenomena can be demonstrated through the use of sound waves, and conversely. A few of the areas in which this analog situation has been useful are described below.

Polarized Airborne Sound Waves[3]

When electromagnetic waves are confined in hollow conducting tubes or waveguides, longitudinal components are created which are not present in a plane progressive free space electromagnetic wave. An analog of this situation exists in the case of airborne sound waves propagating as higher order modes in tubes. Such modes engender a transverse component of particle motion of the air as well as the usual longitudinal component.[4] The higher order mode waves therefore possess properties akin to electromagnetic waves propagating in waveguides. For example, if the size of the tube becomes too small relative to the wavelength, both types of waves experience a "cutoff" effect and propagation cannot occur.

For the first-order transverse mode acoustic wave propagating in a rectangular tube, the transverse dimension must be greater than a half wavelength, and in this mode, the sound pressure is a maximum at both sides of the tube and zero along the center line. In the electromagnetic case the usual transmission mode is the so-called dominant mode and waveguides are generally designed so that only this mode can propagate, all higher modes being beyond cutoff. In the acoustic case a similar dimensioning procedure can prevent still higher order modes from propagating but the normal longitudinal mode must also be suppressed if only the first higher order mode is desired.

Fig. 1 shows an acoustic waveguide in which sound is generated by the small tweeter loudspeaker at the right. This tweeter is coupled into the side wall of a rectangular waveguide so as to generate a substantial component of the first-order transverse wave. The section of waveguide immediately to the left of the tweeter unit has a series of stopped pipes a quarter wavelength long and opening into the center line of this waveguide section. At the frequencies for which these pipes are resonant, a pressure null is forced to exist along this center line; the longitudinal wave is thereby suppressed and the first-order mode (which possesses a pressure null along the center line) is unaffected.

Fig. 2 portrays the sound intensity pattern when such a wave is permitted to radiate from a horn coupled to the mode generator and filter.[5] The pressure null at the center is maintained even after radiation, and two main lobes are generated. In Fig. 3 the phase pattern of this radiation is indicated (method described by Kock and

* Original manuscript received by the IRE, February 16, 1959. This paper, prepared for the July memorial issue of the European Journal, *Acustica*, honoring Prof. Erwin Meyer, director of the III Physik. Inst., Univ. of Goettingen, on his 60th birthday, is being published simultaneously in the Proceedings through the courtesy of Dr. M. Grutzmacher editor-in-chief of *Acustica*.

† Research Laboratories Division, Bendix Aviation Corp., Detroit, Mich.

[1] E. Meyer, "Microwave-acoustic analog experiments at Goettingen," *J. Acoust. Soc. Am.*, vol. 30, pp. 624–632; July, 1958.

[2] Lord Rayleigh, "On the Passage of Waves through Apertures in Plane Screens and Allied Problems," in "Collected Papers," vol. 4, Cambridge University Press, Cambridge, Eng.; 1902.

[3] Material for this section is taken from the demonstration lecture by W. E. Kock and F. K. Harvey, "Polarized airborne sound waves," presented at the forty-sixth meeting of the Acoust. Soc. Am., Cleveland, Ohio; October 16, 1953.

[4] W. P. Mason, "Electromechanical Transducers and Wave Filters," second ed., D. Van Nostrand Co., Inc., New York, N. Y., page 109; 1948.

[5] This photograph was made by employing the techniques for visually portraying sound waves described earlier: W. E. Kock and F. K. Harvey, "A photograph method for displaying sound wave and microwave space patterns," *Bell Sys. Tech. J.*, vol. 30, pp. 564–587; July, 1951.

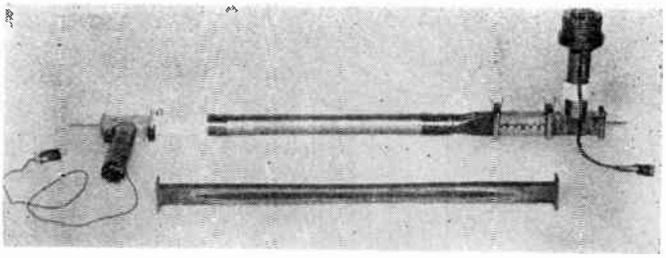


Fig. 1—Equipment for transverse sound wave experiments comprises (top right) generator and longitudinal mode suppressor, (top left) transverse wave receiver, and (in foreground) variable width guide for demonstrating cutoff properties of transverse waves.

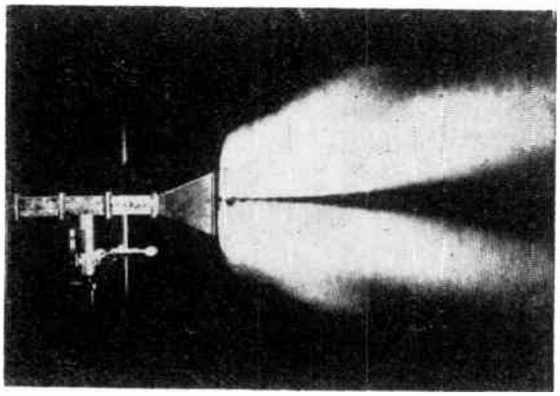


Fig. 2—The radiation pattern of the first-order transverse mode sound wave shows a pressure null along the axis. The generator and mode suppressor at the left are coupled directly to the small horn.

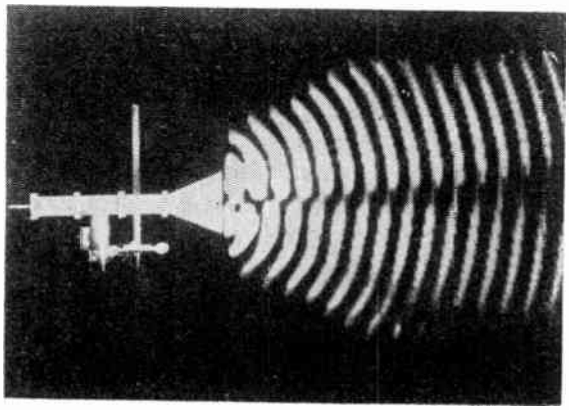


Fig. 3—A phase pattern of the transverse waves of Fig. 2 portrays the progressing wave fronts. The white and dark areas correspond to pressure maxima and minima. At the center line these areas are oppositely phased and cause a transverse air particle motion even in the unbounded medium.

Harvey[5]) and the staggered maxima (bright areas) show that the two lobes are of opposite phase. One sees that, just as longitudinal particle motion exists because of a longitudinal succession of high pressure (bright) areas and low pressure (dark) areas, so transverse particle motive must exist for this mode even in the radiated portion of the wave. Along the center line, high-pressure areas are located opposite to low-pressure areas and this situation generates the transverse air particle motion. Such waves therefore exhibit a "polarized" property in that a receiver similar to the transmitter in Fig. 2 and placed on the center line of radiation can be energized only when oriented in the proper plane of polarization.

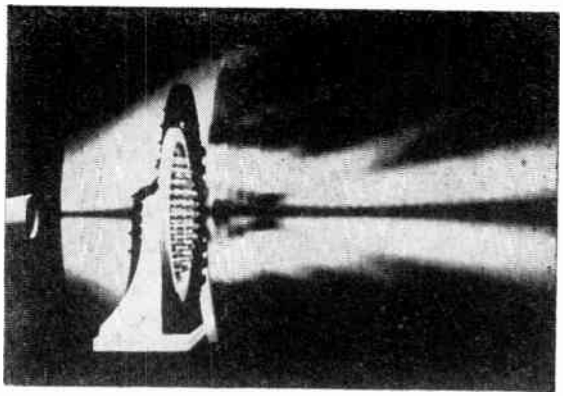


Fig. 4—A lens can be used to narrow the radiated transverse beams of Fig. 2 from the horn at the left.

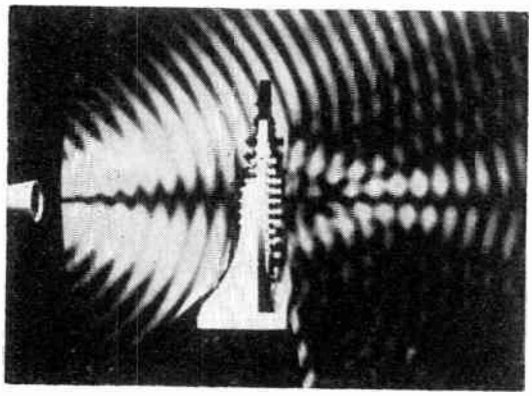


Fig. 5—The phase pattern of Fig. 4 shows that the opposite phase of the two lobes is maintained after passage through the lens.

Also, for the confined wave in Fig. 1, the receiver at the left receives no signal in the orientation shown, but would be energized if it were rotated 90°, *i.e.*, with the tube vertical.

Fig. 4 shows that the two-lobe radiation pattern of the higher mode can be sharpened if desired by the employment of a lens[6] (or reflector). The broad lobes from the horn at the left are converted by the lens into the narrower lobes at the right. Fig. 5 shows that the alternative phase between the upper and lower beams is maintained after passage through the lens.

To demonstrate the cutoff effect the long single waveguide section in the lower part of Fig. 1 is employed. It possesses an accordion-like top and bottom wall so that the transverse width can be reduced by means of a hand clamp. Using the transmitter and receiver shown at the top of the photograph, the cutoff effect for the higher order mode can be demonstrated by reducing the waveguide width to less than a half-wavelength.

Rotation of the polarization (plane of transverse vibrations) of such waves can be accomplished through the use of an analog to the optical half-wave plate, also employed in microwaves. In Fig. 6, free space waves are depicted as entering, from right to left, 1) a dielectric or low velocity medium (top of figure), and 2) a waveguide or high velocity medium (bottom of figure).

[6] This type of obstacle lens was described by W. E. Kock and F. K. Harvey, "Refracting sound waves," *J. Acoust. Soc. Am.*, vol. 21, pp. 471–481; September, 1949.

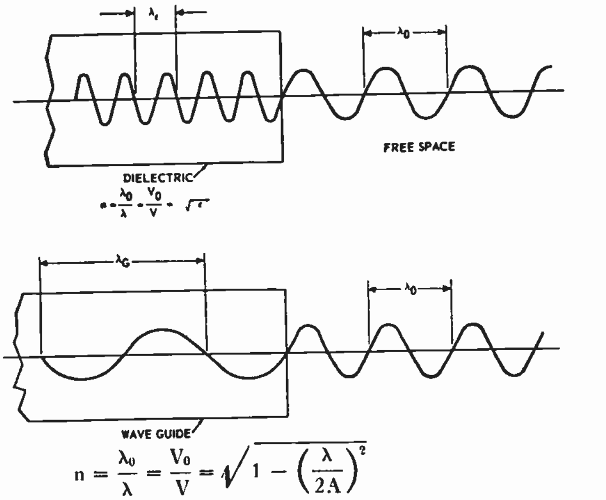


Fig. 6—Free space radio waves entering (top of figure) a dielectric have their wavelength shortened, whereas in a waveguide (bottom of figure), their wavelength and phase velocity are increased. A is the guide width.

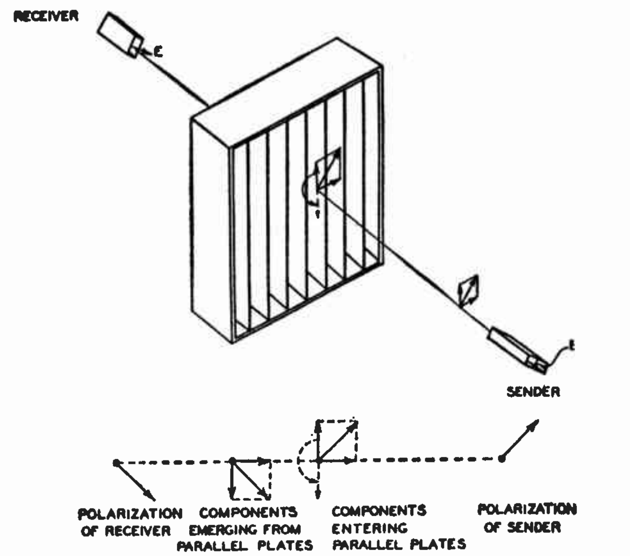


Fig. 7—The greater phase velocity of waves in a parallel plate waveguide enables the structure shown to produce a 90° rotation of polarization.

In the latter case, the observed wavelength λ and the observed phase velocity V are both greater than their free space values λ_0 and V_0 in accordance with the relation indicated (A is the rectangular guide width). This higher phase velocity is made use of in the polarization rotating structure shown in Fig. 7 comprising a series of waveguides placed at 45° to the wave polarization.[7] The phase of the vertical component is advanced relative to the horizontal component (which has free space velocity) and this results in a 90° rotation of polarization.

To accomplish this same result for sound waves, the cylindrical section of guide in the upper part of Fig. 1 is compressed into an elliptical cross section and the axes of the ellipse then placed at 45° to the axes of the rectangular guide.[8] The transverse sound wave entering this elliptical tube can also, as in Fig. 7, be considered as consisting of two components each at 45° to the plane of transverse air motion. When these two components enter the elliptical guide section their velocities of propagation are different, *i.e.*, one of the components experiences a higher phase velocity than the other. For a properly chosen length of this section the relative phase advance can be made to equal 180° and the transverse plane of polarization of the sound wave rotated by 90°.

By shortening the length of the elliptical section, the relative phase advance can be reduced to 90° and under these conditions the transverse sound waves have their plane of polarization continually rotating as they progress. The resultant wave can be compared with a circularly polarized electromagnetic wave. With such a "quarter-wave plate" section of elliptical tube inserted in the transmission path at the top of Fig. 1, the receiving microphone section will be excited in any angular position.

An arrangement similar to that shown at the top of Fig. 1 was used to demonstrate an acoustic analog[9] of the microwave gyrator based on the Faraday effect.[10]

Another transverse acoustic mode of interest is one traveling in a tubular medium having a pressure release surface for its exterior boundary, for example, a mercury delay line. For this wave the pressure is zero at the boundaries and maximum at the center. An airborne variety of this mode can be generated within a rigid tube which has corrugations on its walls. The corrugations can act as quarter wave resonators and force the pressure to be zero at the walls, creating a maximum pressure point at the center of the guide. Again, the phase velocity for this mode is frequency dependent and is greater than the free space velocity.

The higher phase velocity waves in guides has been utilized in the design of lenses for focusing microwaves.[11] An example of such a lens is shown in Fig. 8. Microwaves having their electric polarization parallel to the metal plates shown in the figure experience a higher phase velocity in passing through the lens, and a plane wave arriving from the left is converted to a spherical wave which creates the smaller focal area shown.

In Fig. 9 the same lens has had corrugated structures affixed to alternate plates; this lens was effective in focusing sound waves of proper frequency in the same manner that the lens in Fig. 8 focused microwaves. The corrugations force the sound pressure to become zero near the walls and the waves, in passing through the lens, experience a higher phase velocity. Fig. 10 shows

[7] W. E. Kock, U. S. Patent No. 2,588,249; filed, January, 1946, issued, March, 1952.

[8] A similar technique is employed in microwaveguides. See, for example, G. C. Southworth, "Principles of Waveguide Transmission," D. Van Nostrand Co., Inc., New York, N. Y., p. 327; 1950.

[9] W. E. Kock, "An acoustic gyrator," *Arch. Elek. Übertragung*, vol. 7, pp. 106–107; February, 1953.

[10] C. L. Hogan, "The ferromagnetic Faraday effect at microwave frequencies and its applications—the microwave gyrator," *Bell Sys. Tech. J.*, vol. 31, pp. 1–31; January, 1952.

[11] W. E. Kock, "Metal lens antenna," Proc. IRE, vol. 34, pp. 828–836; November, 1946.

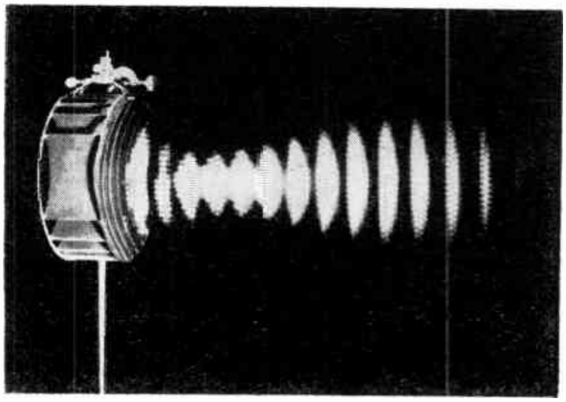


Fig. 8—A parallel plate waveguide structure focuses electromagnetic waves. Because of the increased phase velocity in the guides, the lens must be made concave rather than convex.

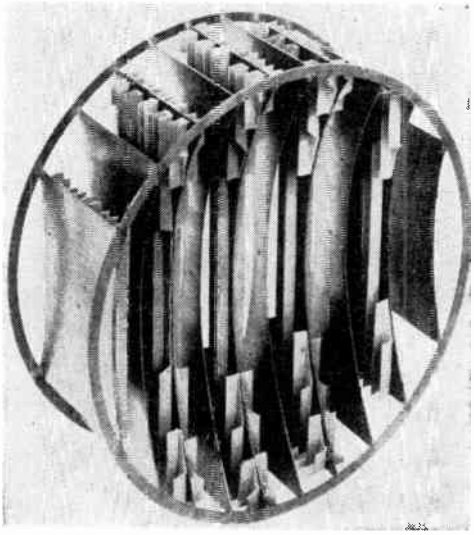


Fig. 9—Quarter-wave resonators affixed to the lens of Fig. 8 cause sound waves to have zero pressure at the resonators. As in the electromagnetic case, the phase velocity of sound waves of proper frequency is increased in passing through the lens and the concave lens causes focusing.

the focusing property of the lens of Fig. 9 for sound waves. In the circled portion of the photograph the phase advance of the sound waves passing through the lens relative to those passing outside the lens is evident. The equivalent photograph for a delay type lens (Fig. 11) shows the retardation of phase in that case.

For a cylindrical corrugated guide, the transverse mode has circular symmetry. The phase velocity again is higher than the free space velocity for those frequencies for which the corrugations are approximately a quarter wavelength deep. In such a guide an interesting experiment can be performed which depends upon the fact that the product of the phase and group velocity of a wave confined in a waveguide is equal to the square of the free space velocity. Because the phase and group velocities are frequency dependent, two acoustic pulses of different frequency should propagate down a corrugated guide with different velocities. This is shown in Fig. 12. In the top photograph, the microphone was placed near the sound source, which was radiating at

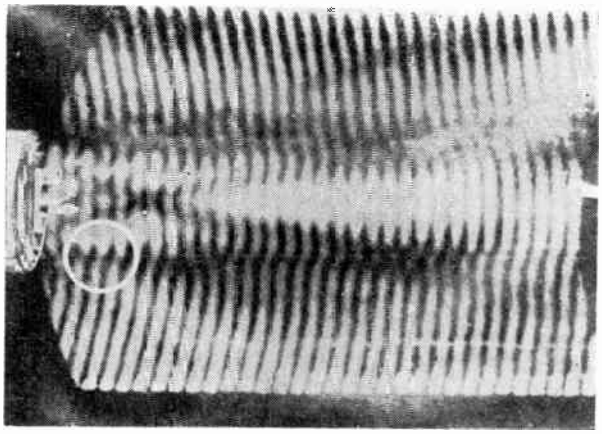


Fig. 10—The bright area in the center of the photograph portrays the focusing effect of the lens of Fig. 9. In the circled portion the phase advance of waves passing through the lens can be observed.

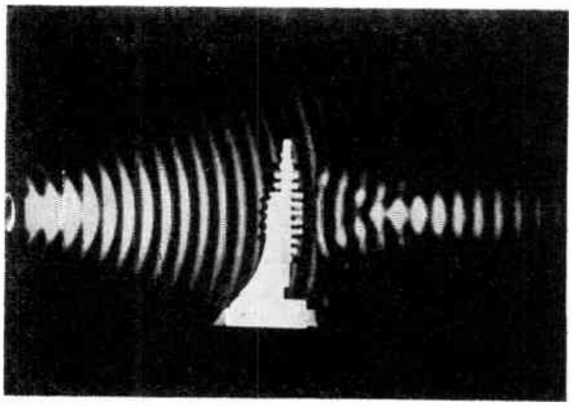


Fig. 11—For an acoustic lens in which the phase velocity is less than free space velocity, a convex shape is required. In this case the phase of waves passing through the lens is retarded.

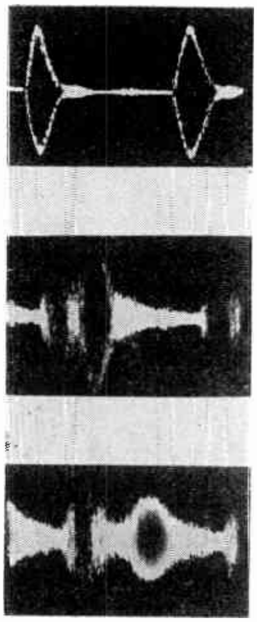


Fig. 12—Two pulses, each comprising two different frequencies, are shown in the top photo as they enter a tube having quarter wavelength corrugations in its wall. The dependence of group velocity upon frequency in such a tube causes the two frequencies present in the pulse to become more and more separated in space and time as they move along the tube (bottom two photos).

7.2 kc and a 10 kc pulse simultaneously, and no visible separation of the pulses is observed. In the second photograph the probe-microphone has been moved 25 inches down the tube (away from the sound source) and the separation of the pulses has begun to become evident. The last photograph, showing the probe moved 52 inches down the tube, displays the pulses well separated.

Externally Guided Waves[12]

It has been known for sometime that dielectric rods can act as waveguides for electromagnetic waves.[13] Thus a rod of dielectric inserted in the open end of a metallic tubular waveguide can guide microwaves a great distance with very little external radiation.[14] If the dielectric rod is made short and is tapered to a smaller dimension, a so-called "end fire" directional radiator is created.[15] With the development of artificial dielectrics for microwave use,[16] analog considerations suggested the possibility of refracting sound waves by these same periodic metallic structures and such an investigation has already been reported.[17] Thus the lens in Fig. 13, designed for microwaves, is capable of focusing sound waves also (Fig. 14). If one compares the lens in Fig. 13 with the usual glass or dielectric lens of optics, it seems reasonable that one of the rows of disks constituting the lens could act as a circular rod of dielectric for waves of proper wave length. If this row of disks, supported axially on a metal rod, is inserted in a waveguide as shown in Fig. 15, both microwaves or sound waves of the proper frequency can be guided by this rod of artificial dielectric. In Fig. 16, the phase fronts of a radiated microwave signal are shown from this structure, and in Fig. 17 the ability of the device to radiate simultaneously both sound waves and microwaves is being demonstrated by F. K. Harvey.

If the disk-studded rod is continued indefinitely, both sound waves and microwaves can be guided along the rod with relatively little radiation. Fig. 18 shows a demonstration of the simultaneous transmission of microwaves and sound waves along a straight and a curved section of this type of transmission line.[18] Fig. 19 shows the radiation pattern of a 15-kc sound wave from the end of the disk radiator of Fig. 15, and Fig. 20 shows the radiation off the end of a curved disk-on-rod waveguide of rather abrupt curvature. Most of the energy is seen to emerge at the end of the rod, however. At the top of the photograph a small amount of energy is seen radiated by the left-hand straight portion of the waveguide. For gradual bends this loss by radiation is less severe. The field around a long straight guide is shown in Fig. 21.

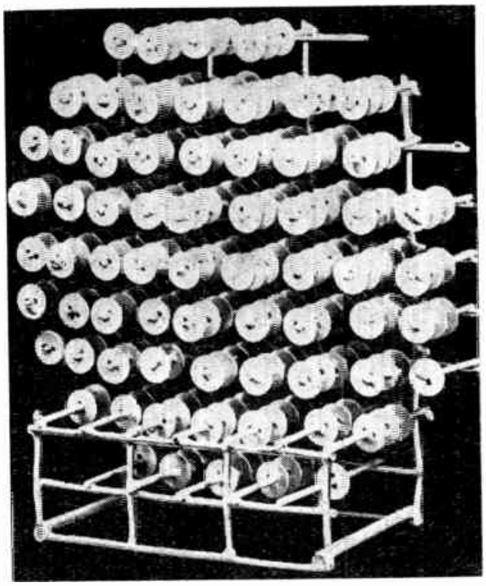


Fig. 13—A structure comprising rows of disks and having an over-all convex shape can act as a lens for radio waves and sound waves.

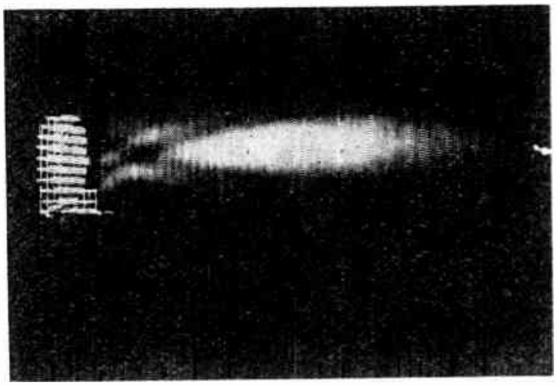


Fig. 14—Sound waves approaching the lens of Fig. 13 from the left are seen to be concentrated or focused into the white area at the right.

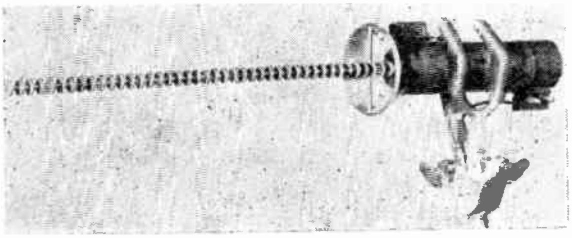


Fig. 15—A row of disks from the lens of Fig. 13, when inserted in the mouth of a tubular waveguide, can act as a directional "end-fire" radiator for either sound waves or electromagnetic waves.

Superdirectivity[19]

In optics, microwaves, and acoustics, maximum directivity of an aperture type radiator has generally been assumed to obtain when uniform excitation exists

[12] Material for this section is taken from a talk presented during the 20th anniversary meeting of the Acoust. Soc. Am., New York, N. Y.; May 6, 1949.

[13] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., Inc., New York, N. Y., p. 425; 1943.

pp. 1188–1192; December, 1949, reports measured attenuations of

[14] For example, C. H. Chandler, "An investigation of dielectric rod as waveguide," *J. Appl. Phys.*, vol. 20, pp. 1188–1192; December, 1949, reports measured attenuations of only 0.00005 db per wavelength in a polystyrene rod for microwave of 1.25 cm. wavelength.

[15] G. E. Mueller and W. A. Tyrrell, "Polyrod antennas," *Bell Sys. Tech. J.*, vol. 26, pp. 837–851; October, 1947.

[16] W. E. Kock, "Metallic delay lens," *Bell. Sys. Tech. J.*, vol. 27, pp. 58–82; January, 1948.

[17] See Kock and Harvey, reference 6.

[18] This photograph was taken at the time of the Acoust. Soc. meeting in 1949.

[19] Material for this section is taken from a talk presented at a meeting of the Long Island Section of IRE, February, 1950.

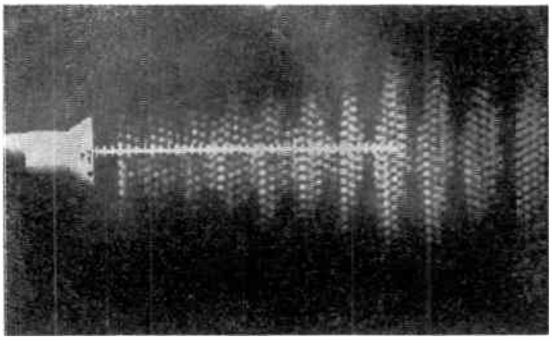


Fig. 16—Waves emerging from the waveguide at the left would normally possess circular wave fronts. The row of disks causes the wave velocity near it to be smaller than free space velocity and the curved wave fronts are thus converted to plane wave fronts. The directivity of a large aperture or "broadside" type radiator is thus achieved by this end-fire process. Radio waves of 3 cm wavelength were used in this photograph.

Fig. 17—The radiator of Fig. 15 can radiate sound waves and electromagnetic waves simultaneously. At the right, a microwave receiver and a microphone detect the presence of both waves. The metallic grid reflects the microwaves but passes the sound waves, and a wooden paddle reflects sound waves but passes microwaves.

over the aperture. Interest has been shown in the possibilities of increasing the directivity of a given size radiator beyond that specified by the usual optical formulas. The first publication on this subject was by Schelkunoff;[20] others who have considered the problem are Hansen,[21] Riblet[22] and Chu.[23] Dolph's application of Tschebyscheff's polynomials to linear arrays[24] was extended

[20] S. A. Schelkunoff, "Mathematical theory of linear arrays," *Bell. Sys. Tech. J.*, vol. 22, pp. 80–107; January, 1943.

[21] W. W. Hansen, M.I.T. RAD. Lab. Rep. T-2.

[22] J. J. Riblet, "Note on the maximum directivity of an antenna," Proc. IRE, vol. 36, pp. 620–623; May, 1948.

[23] L. J. Chu, "Physical limitations of omnidirectional antennas," *J. Appl. Phys.*, vol. 19, pp. 1163–1175; December, 1948.

[24] C. L. Dolph, "A current distribution for broadside arrays which optimizes the relationship between beam width and side-lobe level," Proc. IRE, vol. 34, pp. 335–348; June, 1946.

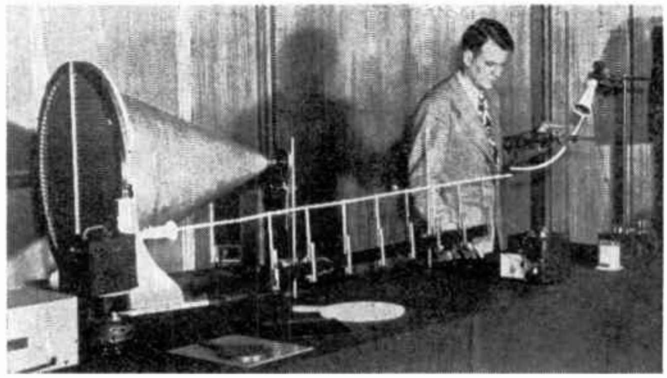


Fig. 18—By extending indefinitely the row of disks of the radiator of Fig. 15, a structure for guiding, externally, both microwaves and sound waves is obtained. When the curved section of guide is axially displaced, the received signal drops rapidly.

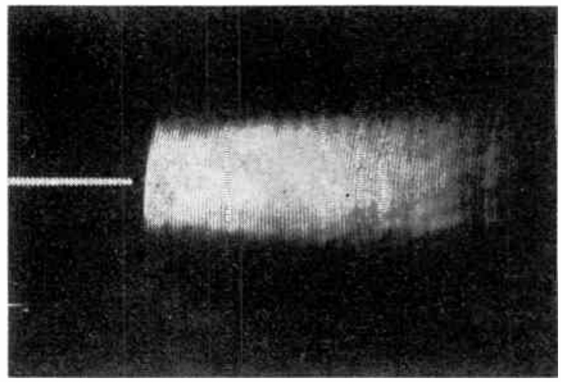


Fig. 19—Sound waves are collimated by the radiator of Fig. 16.

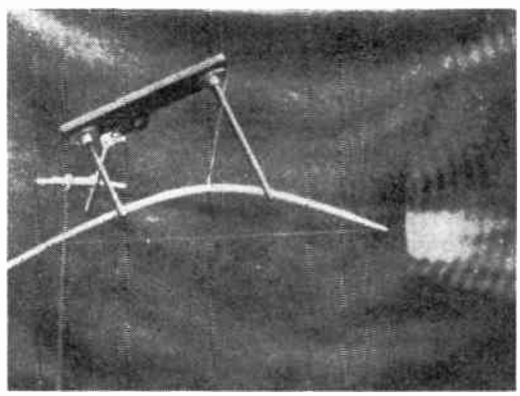


Fig. 20—Most of the sound energy guided by the curved disk transmission line radiates off the end.

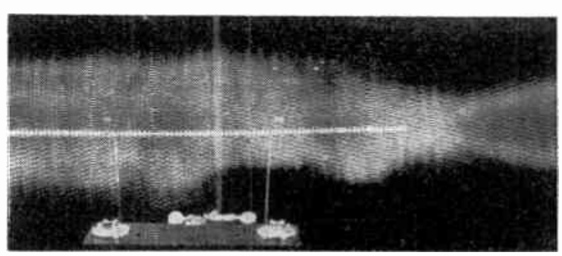


Fig. 21—The microwave field in the vicinity of the open end of an extended section of the disk transmission line is portrayed here.

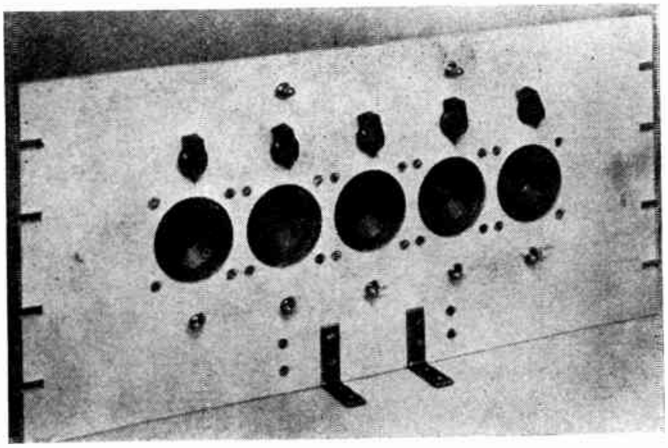


Fig. 22—Five loudspeakers spaced one-quarter wavelength apart can form a superdirective radiator if alternate speakers are reversed in polarity.

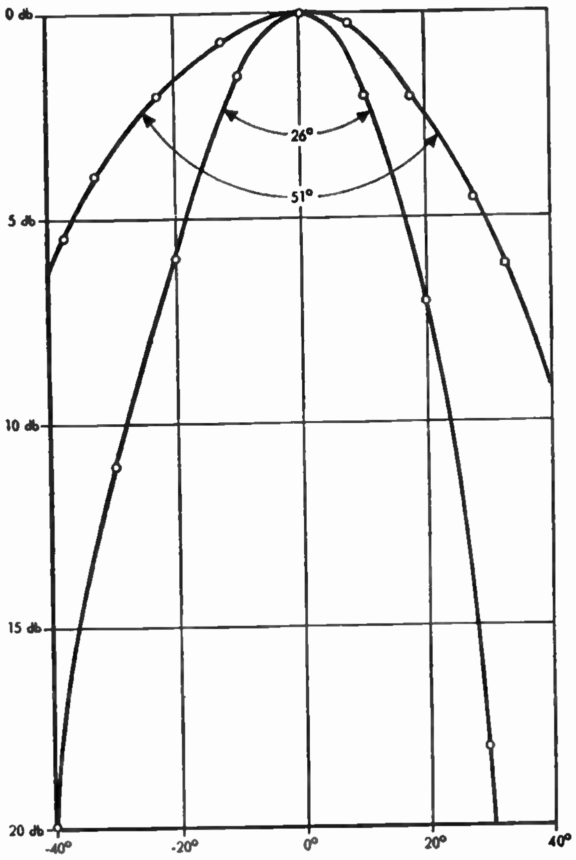


Fig. 23—When the five speakers of Fig. 22 are energized in phase a 51° beam (at the 3-db points) results. When alternate speakers are reversed in polarity, the measured beamwidth drops to 26°.

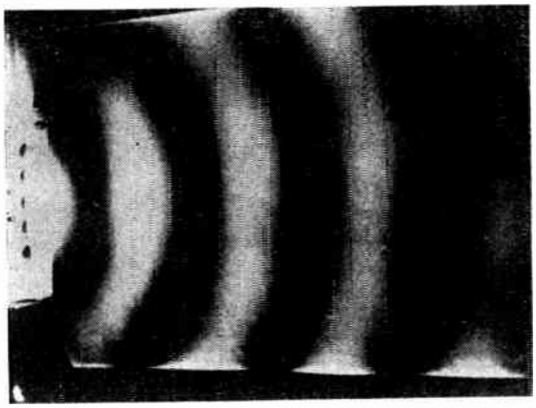


Fig. 24—The in-phase condition for the loudspeakers of Fig. 22 shows the expected curved wave fronts.

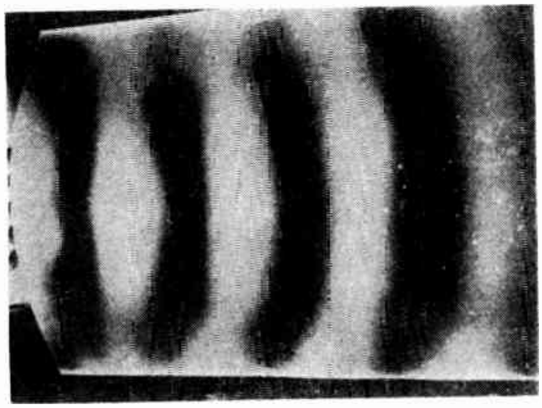


Fig. 25—The reversed-polarity or superdirective condition for the loudspeakers of Fig. 22 yields flatter wave fronts and accordingly, a sharper beam pattern.

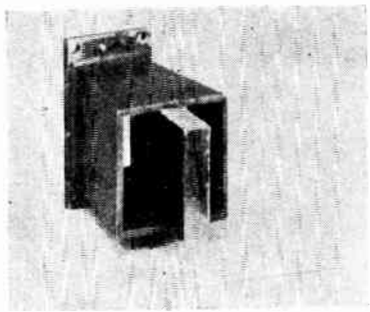


Fig. 26—A three-element superdirective microwave radiator can be created by means of a center section of dielectric which causes a polarity reversal of the energy emerging from this section.

by Pritchard and Rosenberg to superdirective broadside arrays.[25]

Although the principles of superdirectivity could be established experimentally using either electromagnetic or sound waves, the simplest demonstration of the effect was achieved using acoustic techniques. Five small loudspeakers were mounted as shown in Fig. 22, and the acoustic output of each speaker was adjustable by means of the rheostat knobs directly above the units. Two switches, shown at the top of the panel, permitted the polarity of the second and fourth loudspeaker to be reversed relative to the other three. By selecting a frequency such that the spacing between the loudspeaker centers was a quarter wavelength, a five-element array could be made to have either normal directivity (by having all loudspeakers in phase) or to have superdirectivity (by reversing the polarity of loudspeakers two and four). Proper amplitude control was also necessary to achieve the superdirectivity effect.[25]

[25] R. L. Pritchard and M. D. Rosenberg, "Optimum directivity patterns for linear arrays," *J. Acoust. Soc. Am.*, vol. 20, pp. 594–595; July, 1948.

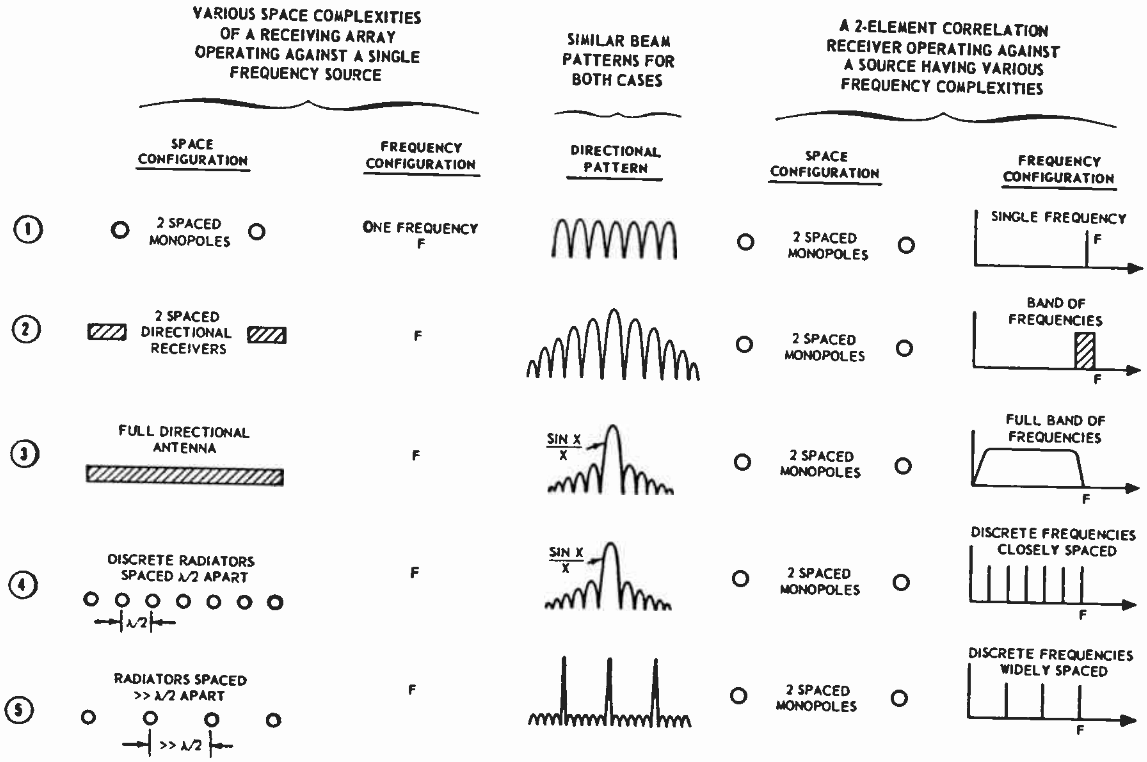


Fig. 27—An equivalence exists in a receiving array between the complexity of its space configuration and the complexity of the frequency configuration of the source.

Fig. 23 shows the beam patterns taken under the two conditions and indicates that a beam sharpening by roughly a factor of two was achieved in the superdirectivity case. The known drawback of power loss also made itself evident. In the superdirective case, the circulating, nonradiating energy between the individual radiators causes viscous losses in the acoustic case and ohmic losses in the antenna in the electromagnetic case. Reversal of the loudspeaker polarity therefore caused an appreciable drop in power radiated even though the beamwidth was sharpened.

The sound portrayal technique was again used to help give an understanding of the reason for the sharpened beamwidth. Thus in Fig. 24 the radiators are in phase and the pronounced curvature of the wave fronts indicates that normal directivity exists. In Fig. 25 (the superdirectivity case) it is seen that the altered polarity has created a flat wave front having a dimension greater than the array length, and the sharpened beam can be considered as being created by an array of about twice the length of the actual array.

Fig. 26 shows a three element microwave superdirective radiator consisting of a waveguide horn into which a section of foil coated polystyrene is inserted. The length of the polystyrene element is such as to cause a 180° relative phase reversal and the thickness is chosen to create the proper energy distribution among the three apertures for achieving superdirectivity. This device functioned in the expected manner.[26]

[26] W. E. Kock, U. S. Patent No. 2,692,336.

Space Frequency Equivalence[27]

Recently, the idea of substituting in a receiving array a complexity in frequency for a complexity in physical structure appeared to have useful significance. This equivalence concept indicates that two widely spaced receivers of small size can achieve a directivity against a multi-frequency source equal to the directivity obtained from a fully extended linear array of twice the length against a single frequency source. Fig. 27 shows a series of space-frequency-equivalence situations. In case one, two spaced monopoles are shown operating at a single frequency; the directional pattern is the well-known rosette pattern having equal multiple lobes. In case two the space configuration is varied so as to give some directionality to each of the two receivers. In the frequency case the two monopoles are retained and their outputs correlated, but the source is now assumed to have a small bandwidth. The patterns for the two cases are again identical and consist of a rosette with sidelobes somewhat suppressed. In case three the directionality of the antenna is made to extend over the full space between the original monopoles; in the frequency case, the source frequency is assumed to extend practically to zero frequency. In this situation a sin x/x directionality pattern is obtained in both situations. In case four the distributed antenna is replaced by a set of discrete radiators spaced a half wavelength apart; in the

[27] The material for this section was taken from W. E. Kock and J. L. Stone, "Space-frequency equivalence," 1957 WESCON Convention Record, pt. 1, p. 216; August 20–23.

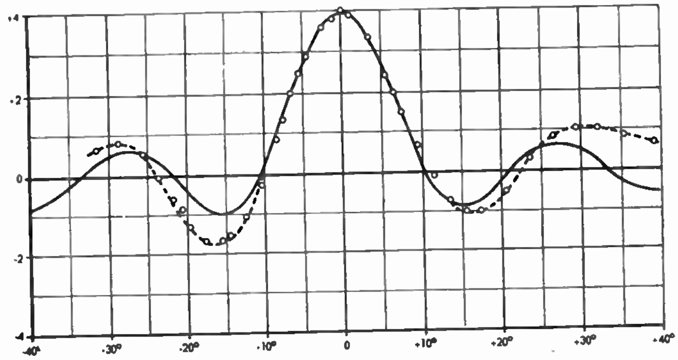


Fig. 28—The correlated outputs of two isotropic receivers spaced 21.5 inches apart against a source comprising frequencies of 300, 900, and 1500 cps creates a beam pattern shown by the dotted curve. The solid curve is the calculated pattern for a six-element linear array 43 inches long against a 1500 cycle source.

frequency case a set of discrete frequencies related by[28] $f_n=(2n+1)f_0$, where $n=0$, 1, 2, etc., replaces the continuous frequency distribution of the source. Again a sin x/x pattern is achieved. In case five the radiator spacing and the frequency spacing are made too large; this results in multiple major lobes.

Testing of this concept for the electromagnetic case provided difficulties because of the broad bandwidths required of the receivers. At audio frequencies this problem was not troublesome and tests could be made to corroborate the theoretical predictions. In one test, a sound source comprising frequencies 300, 900, and 1500 cycles per second was used and two isotropic receivers were placed 21½ inches apart. In Fig. 28 is shown the experimental curve and the calculated pattern for a six-element linear array 43 inches long against a 1500 per cycle source.[29]

Forward Scatter[30]

In echo location devices the target is illuminated by a transmitter and the echo returns to a receiver usually placed at the transmitter location. The fact that the signal scattered back from an object is almost always extremely small compared to the signal scattered or diffracted in the forward direction is apparently not too widely appreciated. This strong signal in the forward direction can be explained in certain cases by Babinet's principle. In Fig. 29 an aperture or hole in a screen upon which a plane wave is falling creates the same diffracted field in the forward direction (at x) as that created by a disk having a size equal to the screen aperture. Now the forward lobe in the case of an aperture is identical with the beam of a radiator having the size of the aperture and radiating a plane wave. It is evident that this forward lobe is always stronger than the echo reflected

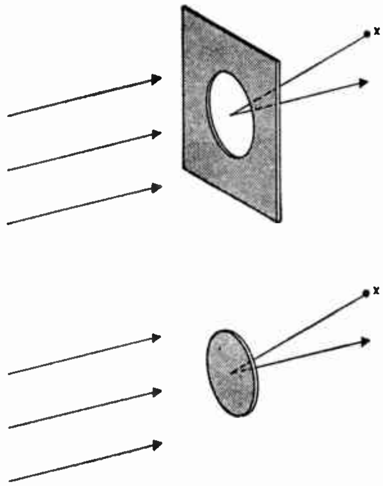


Fig. 29—Babinet's theorem states that the diffracted field from an aperture in a screen (top of figure) is identical with the diffracted field created by an object of equal size (bottom of figure).

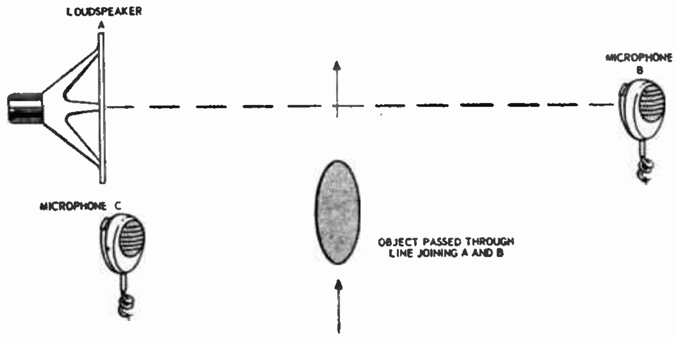


Fig. 30—The signal scattered by an object in the forward direction can be compared with the signal scattered back to the transmitter by using two microphones which have the direct signal cancelled out.

back from an object except in the case where the object is perfectly flat.

Babinet's principle applies strictly only for an infinitely thin screen and for an infinitely thin object replacing the aperture. Thick irregularly shaped objects are difficult to treat analytically, and because of this one cannot conclude theoretically that the forward lobe exists in equal magnitude for a thin, flat disk or for a thick, irregularly shaped object having the same shadow area. Straightforward reasoning suggests, however, that the lobes should be similar since, for objects large with respect to the wavelength, the energy striking the front face of the object is not instrumental in creating the forward lobe.

Again the simplest technique for investigating the forward scatter lobe was an acoustic one. Fig. 30 shows the measurement techniques employed in these tests. In a free space room loudspeaker A, radiating a steady single frequency tone, was placed so that sound could reach microphones B or C. With no object present a signal of proper amplitude and phase was added to the two microphone circuits so as to cause complete cancellation of the directly received signals. When an ob-

[28] W. E. Kock, "Binaural localization and space-frequency equivalence," *J. Acoust. Soc. Am.*, vol. 30, pp. 222–223; March, 1958.

[29] W. E. Kock and J. L. Stone, "Space-frequency equivalence," Proc. IRE, vol. 46, pp. 499–500; Feburary, 1958.

[30] Material from this section is taken from W. E. Kock, J. L. Stone, J. E. Clark and W. D. Friedle, "Forward scatter of electromagnetic waves by spheres," 1958 WESCON Convention Record, pt. 1, p. 86; August 19–22.

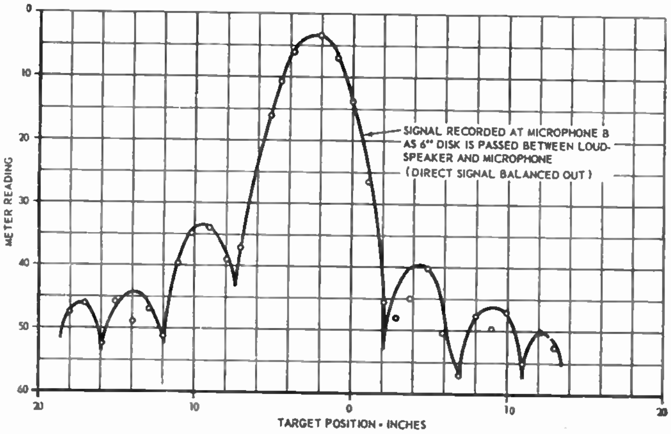


Fig. 31—The beam pattern of the energy scattered in the forward direction is observed to be similar to the beam created by an equivalent aperture in a screen.

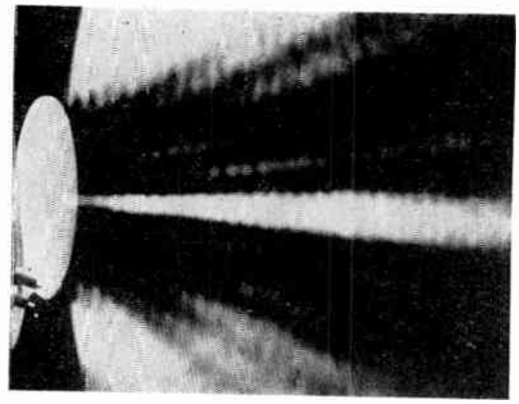


Fig. 32—The lobe scattered in the forward direction can be portrayed visually. In this figure sound waves arrive from the left and the forward lobe is observed in the shadow region.

ject was placed in the position shown, the reflected signal was observed in microphone C and any forward-scatter signal was observed in microphone B.

When a rigid disk was placed in position equi-distant between loud speaker A and microphone B the signals observed in the two microphone circuits were equal. If the disk was then rotated around a vertical or horizontal axis the echo at receiver C was reduced markedly but very little change was observed in the forward lobe until the projected area of the disk was reduced an appreciable amount. When the object was a sphere, the forward-scatter signal was very closely equal to that of the disk of the same cross section but the signal at receiver C was greatly reduced over that produced by the oriented disk. A sphere of sound-absorbing material also created an equal forward-scatter signal but the back-scatter signal was of course extremely small. Directional patterns of the forward lobes of the various objects were measured; they correspond quite well with the calculated values (assuming the object to be equivalent to an equal size hole in a Babinet screen). The pattern of a disk is shown in Fig. 31, and Fig. 32 shows a visual sound protrayal pattern of the forward lobe.

Acknowledgment

Figs. 1–26 were included through the courtesy of Bell Telephone Laboratories. The author is indebted to F. K. Harvey of the Bell Laboratories and to Dr. J. L. Stone of the Bendix Systems Division for collaborative effort in many of the experiments reported upon here.

Tunnel Diodes as High-Frequency Devices*

H. S. SOMMERS, JR.†

Summary—This paper deals with an interesting type of voltage-controlled negative resistance in heavily doped semiconductor junction diodes. The effect, discovered and explained by Esaki, is due to quantum-mechanical tunneling of carriers through the junction. In the present article, experimental and theoretical results are given which show the diode has great promise for frequencies in the kilomegacycle region.

Diodes with a negative conductance of a mho or more have been made; they oscillate above 1 kmc, generate harmonics over 4 kmc, and switch in 2 mμsec. From a gain-bandwidth analysis based on a proposed equivalent circuit, the limiting time constant is shown to be the product of the negative resistance and the junction transition capacitance. According to quantum theory, this product can be varied over a wide range by nominal changes in the free carrier concentration. Germanium diodes with 4.8 by 10^{19} carriers/cm^3 have a measured gain-bandwidth of 1 kmc. Further material development should increase this factor to 10 kmc.

The high negative conductance of the junction coupled with its high shunt susceptance make the device admittance much higher than is normally encountered. As a result, the series impedance of the device mount becomes important. Since the negative resistance is voltage controlled, establishing an operating point requires a voltage supply with internal resistance lower than the magnitude of the negative resistance. Under such conditions, the problem of suppressing parasitic oscillations in the mount and external circuit is serious. The article describes an encapsulation that has been successfully used in the kilomegacycle region and circuit techniques which suppress the parasitic oscillations and increase the effective impedance of the diode.

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† RCA Res. Labs., Princeton, N. J.

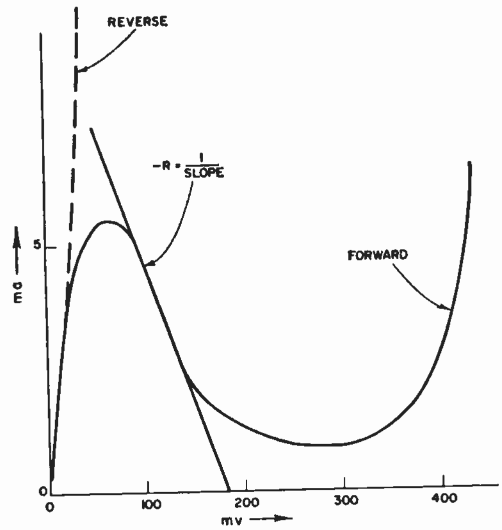


Fig. 1—I-V characteristic of tunnel diode.

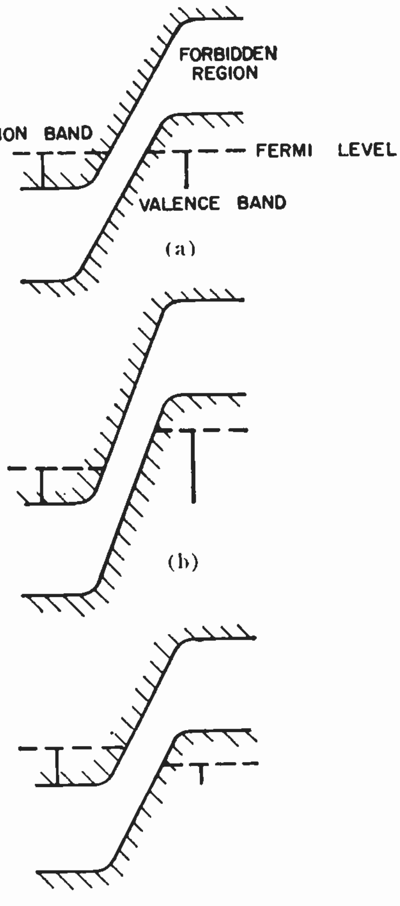


Fig. 2—Energy band of tunnel diode. (a) No applied voltage. (b) Small back bias. (c) Small forwardbias.

I. Introduction

ABRUPT junction germanium diodes made of very highly doped material, reported Esaki,[1] have a negative resistance for small forward bias, an effect he explained as due to quantum tunneling. It is the purpose of the present article to present an approximate equivalent circuit for the device, to give a physical interpretation of the circuit elements, and to point out the limiting time constants and some of the high-frequency possibilities of the tunnel diode.

II. The Effect

Fig. 1 gives the measured dc characteristic of a germanium diode made by alloying a 3 mil dot of In $\frac{1}{2}$ per cent Ga $\frac{1}{2}$ per cent Zn onto a base of 10^{-3} ohm-cm germanium with $2\times10^{19}/\text{cm}^3$ arsenic impurities.[2] For reverse bias, the resistance is small and decreases monotonically with increasing voltage. In the forward direction the current increases to a sharp maximum, drops to a deep and broad minimum, and then increases again. The solid line through the forward characteristic indicates the steepest descent; the reciprocal of its slope is the negative resistance of the diode, $-R$.

Qualitatively the process can be understood from Fig. 2. The material is so heavily doped that the impurities and the Fermi level are in a continuum of states, either adjacent to or part of the normal allowed bands. Calculations based on the density of states of the pure material indicate the Fermi level is about 3 kT above the band edge. With no applied voltage, Fig. 2(a), there is a continuous passage of free carriers through the depletion region by tunneling. An electron approaching the barrier from the left is decelerated by the built-in field, as described by Zener.[3] At the barrier it is in general reflected and returns into the n region. However, the small but finite probability that it will be in the other band when it leaves the barrier is the tunnel effect. This transmitted component is actually very large, perhaps 10^3 amps/cm^2, in spite of the small transmission probability per electron.

In the P region, the electrons are also decelerated as they approach the barrier, and some of them tunnel to the conduction band.[4] These two counter currents, which of course cancel with no applied voltage, have magnitudes indicated by the vertical distance from the Fermi level on one side to the forbidden region on the other (the vertical bars shown in Fig. 2).

Fig. 2(b) shows the case for small reverse bias. The bias appears across the depletion region, displacing the band edges in such a way as to increase the back cur-

[1] L. Esaki, "New phenomenon in narrow Ge p-n junctions," *Phys. Rev.*, vol. 109, p. 603; 1958.

[2] To study this characteristic in the negative resistance region requires suppressing all oscillations, not a trivial task. Our technique is to load the diode with a noninductive shunt resistance so as to overdamp all oscillations; the diode current is deduced by correcting for the drain through the shunt. The problem of suppressing oscillations is further discussed in Section IV.

[3] C. Zener, "Theory of the electrical breakdown of solid dielectrics," *Proc. Roy. Soc.*, vol. 145, pp. 523–529; 1934.

[4] The process can be equally well explained by studying the motion of the electrons on one side of the barrier and the holes in the other. We have found it more convenient to examine the motion of electrons on both sides.

rent, the forward current being roughly constant. For small forward bias, Fig. 2(c), the back current is reduced, again with little change in forward current. The current reaches a maximum as the back component drops to zero; it then declines as the forbidden region begins to block the forward component, and finally increases again as normal minority carrier injection over the barrier becomes important. For good germanium units, the voltages of the extrema are around 50 and 350 mv respectively, roughly independent of the initial doping. Observed current ratios are as high as 15/1 at room temperature.

III. High-Frequency Behavior—Theoretical

The nature of the phenomenon which produces the negative resistance in the tunnel diode suggests the diode might be a high-frequency low-power device. High-frequency performance is to be expected because tunneling is a majority carrier effect with no limitation of minority carrier drift time,[5] while low power dissipation is assured from the low bias voltage (around 100 mv) at which the negative resistance occurs.

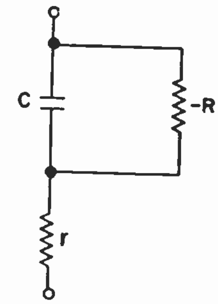
As a basis for calculating the high-frequency performance of a tunnel diode we propose the equivalent circuit of Fig. 3(a). Here $-R$ is the negative resistance of the diode taken from the slope of the dc characteristic at the operating point, C is the junction transition capacitance, and r the dissipative resistance of the diode, including losses inherent in the base and dot and in the soldered connections. In all the units yet made, r is small compared to R. C, on the other hand, is large; for an abrupt junction with 4×10^{19} carriers/cm^3 in the bulk material, it will run about 5 μf/cm^2 of junction area.[6] Since a junction 1.5 mils in diameter will still have a capacitance of nearly 100 $\mu\mu$f, the high-frequency impedance of this device is abnormally low. As shown in the next paragraphs, however, this large capacitance does not prevent high-frequency operation.

Fig. 3(b) is the ac equivalent circuit of a tuned amplifier. The input is across the resistance r_1 while the output is represented by r_2 in series with L_2. C_2, a blocking capacitor, is large enough to have negligible reactance in the bandpass region. The dc power source which keeps the diode in the negative-resistance region is considered to be isolated from the ac circuit.

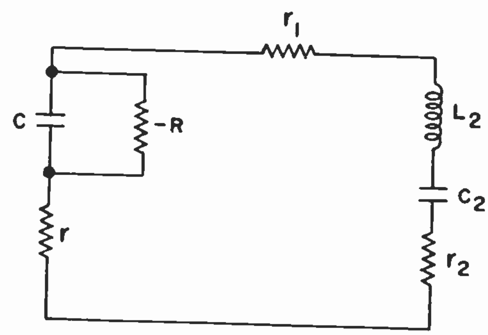
We have analyzed the gain-bandwidth product of this amplifier. For gains much larger than unity we find the approximate expression

$$G\Delta f = 1/(2\pi RC) \tag{1}$$

where G is the mid band voltage gain and $\pm\Delta f$ the width of the resonance curve at half power. Similarly



(a)



(b)

Fig. 3—Equivalent circuit. (a) Tunnel diode. (b) Amplifier.

we find for the highest frequency at which the circuit will sustain self-excited oscillations, defined as f_0,

$$f_0 = (R/r_t)^{1/2}/2\pi RC. \tag{2}$$

r_t is the total ac dissipative resistance of the oscillator circuit. These expressions show the frequency performance of the device can be controlled by changing the RC constant of the depletion region.

IV. Control of Cutoff Frequency

This form of the frequency limit is not unexpected, but in distinction to most devices R is not a dissipative term but the magnitude of the negative resistance. Hence, though RC is independent of area, it is not a constant for a given semiconductor but depends drastically on the doping. While C is a physical capacitance determined principally by the junction area, R can be thought of as proportional to the reciprocal of the transmission coefficient for tunneling through the barrier. As such, it is extremely sensitive to the barrier thickness and hence the free carrier concentration.

Adapting the treatment of Spenke,[7] the tunneling by Zener effect is proportional to the factor

$$\exp - (AE_{vc}[km^+/n^+]^{1/2}). \tag{3}$$

The constant A depends on the units, E_{vc} is approximately the band gap, k the dielectric constant, and m^+ the reduced effective mass of the carriers on the two sides of the barrier (approximately the effective mass of the lighter carrier); n^+ is the weighted average of the

[5] W. Shockley and W. P. Mason, "Dissected amplifiers using negative resistance," discuss the high frequency advantages as well as circuit problems associated with negative resistance diodes, *J. Appl. Phys.*, vol. 25, p. 677; 1954.

[6] See E. Spenke, "Electronic Semiconductors," McGraw-Hill Book Co., Inc., New York, N. Y.; 1958.

[7] E. Spenke, *loc. cit.*

carrier concentration which determines the thickness of the depletion region. It is equal to $n+p/np$ in terms of the majority carrier concentrations, n and p, on the opposite sides of the junction. The implications of (3) can be seen from comparison with the conclusions of Zener in his treatment of breakdown in insulators; where for Zener breakdown the current rises exponentially with applied field, here the tunneling conductance, and hence $1/R$, grow exponentially with increase in doping.

Table I shows some preliminary measurements of the change in RC time constant with doping. The first column gives the electron concentration in the n-type base; the concentration of holes in the alloyed P region is not known but is believed to be somewhat higher. R, the magnitude of the negative resistance, has been measured in only those units which were mounted in special low inductance mounts suitable for high-frequency work (see Section IV). The data indicate that R is about 0.4 times the average resistance connecting the maximum and minimum of current; the latter can be readily measured on any transistor curve tracer. C, which depends mainly on junction area, was measured on a low impedance bridge.

TABLE I
EFFECT OF CARRIER CONCENTRATION ON TIME CONSTANT OF TUNNEL DIODES

Electrons/cm³	R Ω	RC nanosec
2.4×10^{19}	90	4.5
3.6	4.5	0.9
4.8	1	0.05

RC, the device time constant, is given in the last column. As the doping increases from 2.4 to $4.8\times10^{19}/\text{cm}^3$ the time constant of a germanium tunnel diode goes from 5×10^{-9} to 5×10^{-11} sec.[8]

V. Low Impedance Mounting

As already discussed, the tunnel diode has a very high admittance and can only be operated at high frequency if the package in which the diode is mounted has a low series impedance. Junctions with the characteristics of Table I can be operated in the kilomegacycle region if the series inductance of the mount is less than 100 $\mu\mu$h. No pigtail connection to the diode is permissible, which prevents use of normal transistor stems. Even the standard microwave cartridge, though suitable for point-contact diodes, has too much inductance for alloy-junction tunnel diodes.

We have solved the problem of reducing the lead inductance by incorporating the diode into a transmission line. Fig. 4 is a cross section of one such mounting which has permitted operation in the kilomegacycle region. One-eighth inch ribbons of nickel are bonded on either side of ceramic spacers ten mils by $\frac{1}{8}$ by $\frac{1}{8}$ inch. In the $\frac{1}{8}$-inch opening between the ceramics, the diode is mounted with base soldered to the bottom electrode and alloy dot to the top. For mechanical strength, the opening is filled with a plastic such as araldite. The unit is designed to be clamped directly into a section of microstrip line.

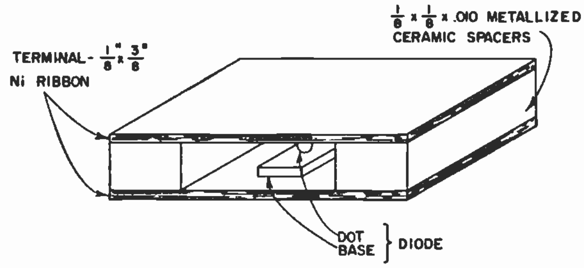


Fig. 4—Microstrip mount of tunnel diode.

VI. Switch

To test its switching response, the tunnel diode was mounted in a transmission line as shown in Fig. 5(a). Either a coax or microstrip line is satisfactory. The output from a pulse generator was capacitatively coupled to the line terminated in the diode. A short circuited delay line connected to a tee reflected an inverted pulse to the diode some ten nanoseconds after the first pulse. The voltage across the diode was viewed on a sampling scope[9] connected to A.

In Fig. 6 we have a drawing of the record on the sampling scope. Fig. 6(a) shows the behavior with no bias across the diode. The direct pulse, about 2 nanosec long, is followed by the inverted reflection some 10 nanosec later. That no switching has occurred is shown by the constancy of the base line between pulses.

Fig. 6(b) is the waveform with the diode biased to point a by the applied voltage V_0 [Fig. 5(b)]. Before the pulse reaches the diode, the diode has the dc voltage of state a; the first pulse switches it to the higher voltage of state b, while the reflected pulse brings it back to a. The important point is that the switching time is less than the rise time of the pulse itself; this particular diode had an RC product deduced from static measurements of 0.5 nanosec.

VII. Self-Excited Oscillator

A tunnel diode will oscillate whenever an operating point is established in the negative resistance region and a suitable ac tank circuit exists. The conditions for these simultaneous occurrences is shown by the load lines of Fig. 5(b). To establish the dc operating point at c, the dc circuit must have a resistance $r_1<R$ so that the load line will intersect the characteristic curve of

[8] The measurements are only semi-quantitative; their accuracy is limited by techniques of alloying and measurement and an imperfect understanding of conditions in the depletion region and the dot.

[9] G. B. Herzog, to be published.

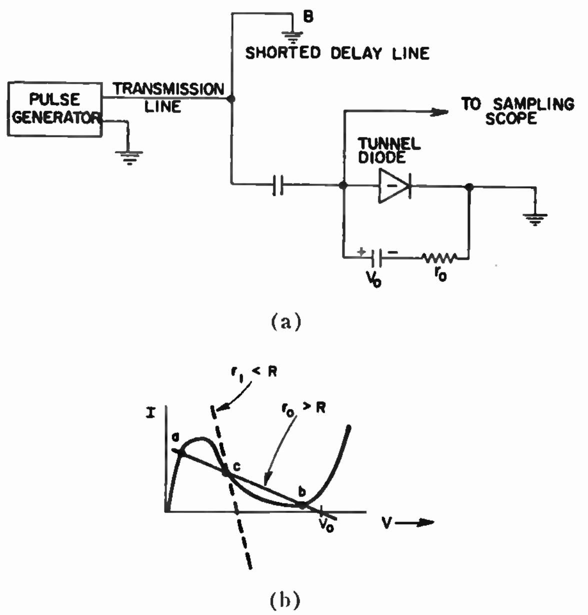


Fig. 5—Switching test. (a) Schematic of pulse test. (b) Diode characteristic with load lines.

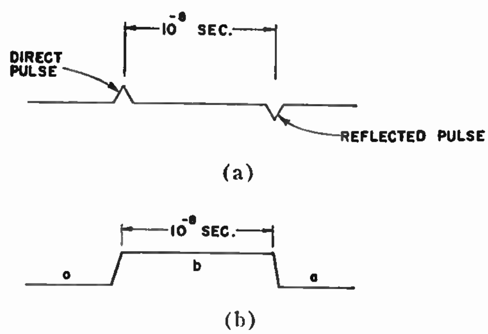


Fig. 6—Diode voltage in switching test. (a) No bias on diode. (b) Bias V_0 applied to diode.

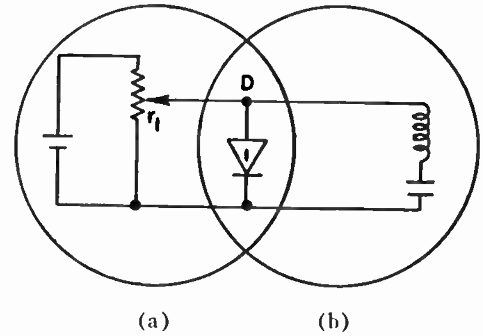


Fig. 7—Oscillator schematic diagram.

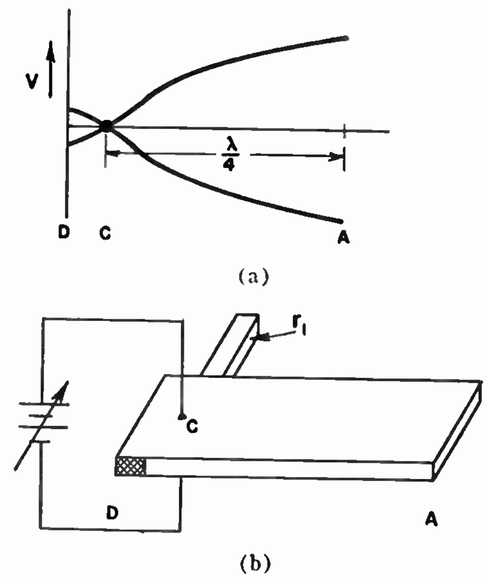


Fig. 8—Quarter wave line. (a) Standing wave pattern. (b) Schematic diagram.

the diode in one point only. For too large a dc series resistance, the load line will resemble r_0 and the dc operating point will be at either a or b, as already described.

The ac tank circuit, on the other hand, must have a multivalued intersection of the load line, such as has r_0. This load line will exist at any frequency where the Q of the circuit exceeds $1/\omega RC$ for the diode. The problem is not of achieving a high Q for the desired mode but in suppressing oscillations for all others.

To illustrate the problem, which is difficult, we have shown a schematic of the dc circuit, Fig. 7(a), and the desired ac tank circuit, Fig. 7(b). The tunnel diode D and its leads are common to both of these. To excite the circuit, which is tuned to around 1 kmc, we must prevent parasitic oscillations of much lower frequency in the circuit involving the inductance of the dc leads which are terminated in r_1.

One solution is to move the frequency of the parasitic oscillations to well above the desired operating frequency, to beyond the cutoff frequency of the diode if possible. We have managed this by placing r_1, a noninductive resistor, at a point so close to the diode that intervening inductance is unimportant compared to the effective inductance of the desired mode; the point of connection is one where the ac swing for the wanted mode is zero so that r_1 will not damp it. Viewed from the dc source, the fact that $r_1 < R$ means the input resistance is positive and no parasitic oscillation will be introduced by the circuit connecting the battery to r_1. This configuration has been achieved, for example, with the quarter wave microstrip line illustrated in Fig. 8(a). The tunnel diode is connected at D; its low impedance permits it to be close to the zero voltage point, shown at C.

Fig. 8(b) shows the microstrip with the diode clamped between the two conducting foils at D and an open circuit at A. The noninductive resistance r_1 is connected at C, either as a short tee arm or between the conducting layers. A suitable noninductive resistance can be made by soldering a wafer of high conductivity germanium between the microstrip conductors. At C the dc leads can also be connected without interfering with the RF mode. This particular oscillator has also the advantage of giving a voltage step-up from the diode to the open end of the line, thus permitting the low impedance diode to be matched into a standard impedance line. Voltage step-ups of greater than 7/1 have been observed with a quarter wave line made of ½-inch wide microstrip of ten mil spacing.

We have studied a number of diodes with low inductance mounts in several different lines. Resonance has been observed by picking up the signal on a radio receiver. For convenience, the oscillation can be modulated by adding an audio modulation to the dc source; this gives both frequency and amplitude modulation. The standing wave pattern has been checked with a travelling probe. We have observed oscillations with a fundamental mode as high as 1.4 kmc.

There are a variety of ways to tune the line. The diode frequency can be pulled several mc by changing the dc operating point. More drastic shifts come from changing the physical length of the line; for very large frequency changes the distance *D-C* will have to be varied with *C-A*, but we have shifted the frequency 30 per cent by changing *C-A* alone. Or the open circuit end can be terminated in a smaller trimmer capacitor. In one line, changing the trimmer from 5 to 17 $\mu\mu f$ with other parameters fixed, lowered the frequency from 400 to 300 mc. This is the kind of change in capacitance one might expect from a parametric diode, indicating the possibility of electrical tuning or modulation over a large range of frequencies.

VIII. Miscellaneous Applications

The tunnel diode is very well suited for use as a self-excited converter, with conversion gain from its negative resistance. It can also be used readily as a negative resistance amplifier with a good noise figure.[10] The very nonlinear I-V characteristic of the tunnel diode permits a large number of uses. For some of these the behavior may be far from ideal but still much the best available because of the shortage of high-speed low-power devices.

The bistable characteristic indicated in Fig. 5(b) permits use as a memory, an important application. If biased to a point beyond the current minimum, *b* of Fig. 5, it becomes a voltage regulator or clipper, as it does in the reverse direction at somewhat lower voltage. Biased to either point *a* or point *b*, the diode, in parallel with a suitable resistance to flatten the characteristic in the negative resistance region, transmits pulses of only one polarity. Around point *b* it is also a constant current source.

When the dc input is terminated in a low resistance, as already described, the dc termination presents a positive resistance to the battery circuit; hence these units can be connected in series without changing their ac characteristics, permitting a constant current source to handle a number of units. Or a group of coupled diodes can be mounted in series, with the appropriate shunting resistors incorporated, to give a higher ac voltage swing, achieving higher power output without lowering the impedance.[11]

IX. Ultimate Frequency

As far as the ultimate frequency limitation is concerned, (2) combined with (3) gives no light. It merely says that higher doping in germanium, or choice of a material with smaller effective mass, will give greatly increased speed. Actually, the present quantum mechanical treatment is based on stationary states of the electrons and gives no account of the time spent in the transition region. Classically the time needed for an electron moving with thermal velocity to move a distance equal to the barrier thickness is about 10^{-13} sec, but it is doubtful if this has any real significance. Of more importance is the dielectric relaxation time, the product of the dielectric constant and the resistivity of the semiconductor, which is the time for the majority carriers to adjust themselves to a change in applied voltage. For highly doped germanium it is also about 10^{-13} sec. It appears that the top frequency of the tunnel diode will be limited by technical problems of fabrication such as the maximum doping achievable or the minimum mounting impedance, or circuit difficulties in exciting the desired mode, rather than any limiting time constant of the physical process.

X. Acknowledgment

The author wishes to thank D. O. North for first pointing out the existence and importance of this effect and for frequent theoretical discussions, C. W. Mueller and H. Nelson for advice on fabricating the diodes, and E. O. Johnson for continuous suggestions about their utilization. The pulse tests were done by G. B. Herzog, who has given his kind permission to publish his results. J. J. Gannon and R. Breinig have been responsible for making and mounting the diodes.

[10] K. K. N. Chang, "Low-noise tunnel-diode amplifier," Proc. IRE, this issue, p. 1268.

[11] Since the voltage swing is constant, the power of a single diode varies as $1/R$.

The Cryosar—A New Low-Temperature Computer Component*

A. L. McWHORTER†, MEMBER, IRE, AND R. H. REDIKER†, SENIOR MEMBER, IRE

Summary—**The cryosar is a high-speed two-terminal computer component whose operation, at liquid helium temperature (4.2°K), is based on impact ionization of impurities in germanium. Two types of cryosars are discussed: the first, fabricated using uncompensated germanium, exhibits a high resistivity ($\sim 10^7$ ohm-cm) until a critical field ($\sim$10 volts/cm) is reached, after which the current increases by as much as seven orders of magnitude; the second, fabricated using compensated *p*-type germanium, has similar electrical characteristics except that a negative resistance region occurs between the high- and low-impedance states, making bistable operation possible. These properties are due to bulk effects, and since both contacts are ohmic, the device is bilateral. The first type of cryosar can perform the functions of an ordinary diode; the bistable cryosar can be used as a memory element, multivibrator or flip-flop. Both types are very fast, the speed being limited by the turn-on time of 10^{-8} seconds or less. Since the active region of each cryosar is limited to the volume directly between its two contacts, a large number of independent cryosars may be placed on one wafer of germanium. Present results point to excellent reliability and reproducibility of the individual elements, making feasible the plating or evaporation of large arrays, possibly integrated into microprinted circuits. If one requires the cryosars alone, it should be possible to fit 200,000 into a cubic inch.**

INTRODUCTION

THE cryosar is a new semiconductor device, intended primarily for high-speed computer switching and memory applications, which utilizes the low-temperature avalanche breakdown produced by impact ionization of impurities. Present cryosars, using germanium doped with Group III and/or Group V impurities, operate at liquid helium temperature (4.2°K). Since the semiconductor returns to its high impedance state by a recombination of the carriers with the ionized impurities, the name of the device was derived from "low-temperature (***cryo***-) ***s***witching by ***a***valanche and ***r***ecombination."

Impact ionization of impurities in uncompensated[1] germanium has been discussed in the literature.[2–5] At liquid helium temperatures, with low applied voltages, germanium may have a resistivity as high as 10^9 ohm-cm. The carriers which were mobile and contributed to the conductivity at room temperature are almost all attached to the impurity centers at these low temperatures. The residual conductivity is due either to those few free carriers generated thermally and by stray radiation, or to a conduction process in the impurity levels themselves, which will be discussed later. As the voltage applied to the germanium is increased, it becomes possible for the free carriers to gain sufficient energy in the electric field to ionize the impurities upon impact. Finally at some critical electric field, the impact ionization rate exceeds the recombination rate and a reversible nondestructive breakdown occurs, similar in many respects to avalanche breakdown in a gas. At the end of the avalanche process, essentially all the impurities are ionized and the resistance changed by as much as seven orders of magnitude.

BREAKDOWN CHARACTERISTICS

Fig. 1 shows a typical voltage-current characteristic for uncompensated germanium, the sample being indium-doped *p*-type germanium in the form of a wafer with ohmic contacts on the opposite faces. Cryosars with such characteristics have previously been proposed for large-scale switching and gating circuits in computers.[6] While they can perform no functions that an ordinary diode cannot do, they have many advantages in terms of mass fabrication, compactness, and reliability, as will be discussed.

A much wider range of applications was recently opened up, however, with the discovery[7] that in compensated germanium there is a region of negative resistance between the high- and low-impedance states, permitting bistable operation. Fig. 2 shows the effect for two samples of indium-doped germanium compensated with antimony, again in the form of wafers with ohmic contacts. As yet there is no explanation for the occurrence of the negative resistance. Experiments have been performed which show that it is not a contact effect and that it is reproducible for germanium with the same doping. Oscilloscope presentations of the voltage-

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† Lincoln Lab., Mass. Inst. of Tech., Lexington, Mass.

[1] By uncompensated germanium is meant germanium which is not intentionally compensated. Of course in this germanium there is always a finite density of compensating centers ranging from 10^{11} cm^{-3} for ultrapure material to around 10^{13} cm^{-3} for transistor-grade material.

[2] N. Sclar and E. Burstein, "Impact ionization of impurities in germanium," *J. Phys. Chem. Solids*, vol. 2, pp. 1–23; March, 1957.

[3] S. H. Koenig and G. R. Gunther-Mohr, "The low temperature electrical conductivity of *n*-type germanium," *J. Phys. Chem. Solids*, vol. 2, pp. 268–283; 1957.

[4] G. Finke and G. Lautz, "On impact ionization in germanium single crystals in the 4.2–10°K temperature range," *Zs. f. Naturforsch.*, vol. 12(a), pp. 223–225; March 1957.

[5] S. H. Koenig, "Rate process and low-temperature electrical conduction in *n*-type germanium," *Phys. Rev.*, vol. 110, pp. 986–988; May 15, 1958.

[6] A. L. McWhorter, "Switching Elements Utilizing Impact Ionization of Impurities," presented at Semiconductor Device Research Conference, Boulder, Colo.; July 15–17, 1957.

[7] R. H. Rediker and A. L. McWhorter, "Low-Temperature Semiconducting Computing Elements," presented at the International Conference on Solid-State Physics in Electronics and Telecommunications, Brussels, Belgium; June 2–7, 1958.

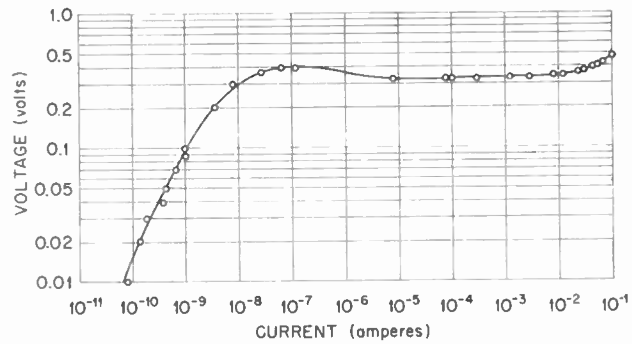

Fig. 1—Voltage-current characteristic at 4.2°K of uncompensated p-type germanium. The thickness of the sample was 0.039 cm, the diameter of the ohmic contacts 0.1 cm. Note that a current variation of nine orders of magnitude is shown.

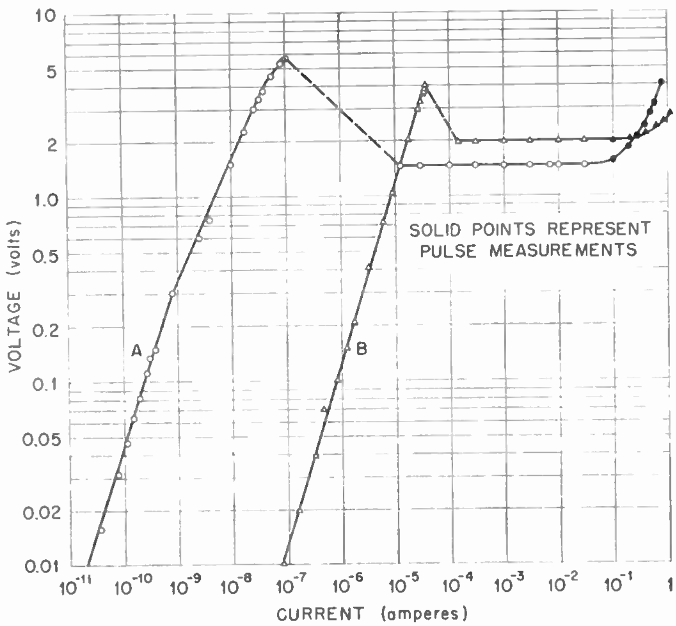

Fig. 2—Voltage-current characteristics at 4.2°K of indium-doped germanium strongly compensated with antimony. In sample A, which is more heavily doped than sample B, the conductivity in the impurity levels is much higher and predominates over the valence band conductivity almost to breakdown. The data are for a sample thickness of 0.043 cm and a contact diameter of 0.1 cm.

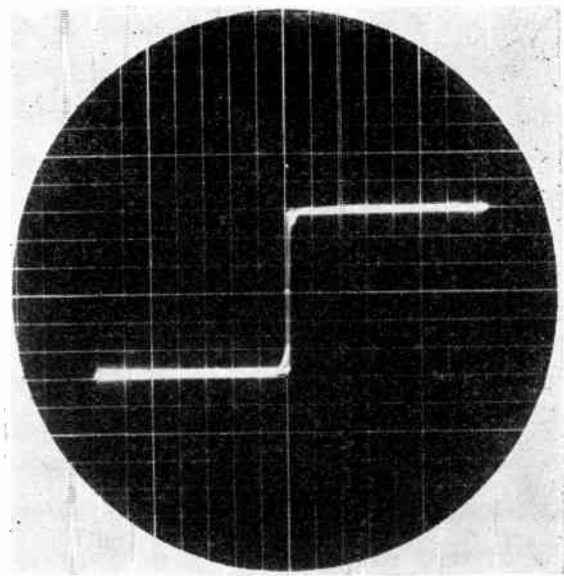

Fig. 3—An oscilloscope presentation of the voltage-current characteristic at 4.2°K of uncompensated germanium. Voltage is the ordinate, current is the abscissa.

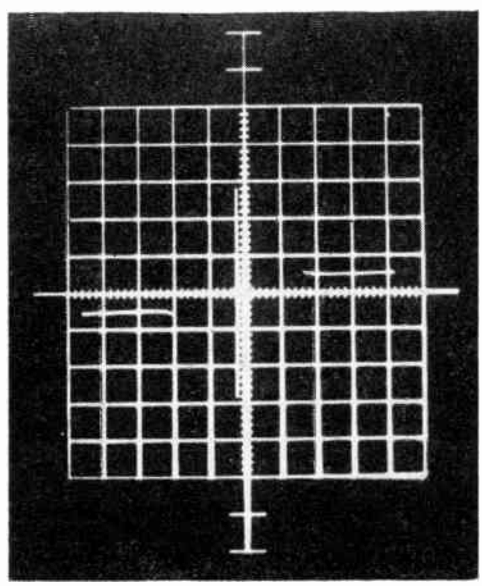

Fig. 4—An oscilloscope presentation of the voltage-current characteristic at 4.2°K of compensated p-type germanium, showing the three stable states: low resistance in either direction and high resistance. The horizontal scale is 20 ma per large division and the vertical scale is 2 volts per large division.

current characteristics of a monostable and bistable cryosar are given in Figs. 3 and 4, respectively. While Fig. 2 is a logarithmic plot, Fig. 4 is linear and shows more clearly the magnitude one can obtain for the negative resistance. The cryosar is of course bilateral since both contacts are ohmic.

These oscilloscope presentations were obtained by using as a source of voltage a 500 μsec sawtooth at a 1-cps repetition rate. The breakdown characteristics were also studied using a series of 10-μsec pulses whose amplitude was varied at a 60-cps rate and whose repetition frequency was about 10 cps. For the data plotted in Figs. 1 and 2, both pulse measurements agreed with each other and with the point-by-point dc measurements for currents smaller than 100 ma. For currents larger than 100 ma, there was some sample heating and only the pulse measurements are plotted.[8]

In impact ionization breakdown, the region where the conductivity is modulated is localized very sharply to the volume where the electric field is above its critical

[8] A heating effect has been postulated (S. H. Koenig and R. D. Brown, "Thermal oscillations in n-germanium," *Bull. Am. Phys. Soc.*, ser. II, vol. 4, p. 27; January, 1959) to explain the much smaller negative resistance region that sometimes occurs in the V-I characteristic for uncompensated germanium. Since the state of the bistable cryosar can be switched with a pulse of 2×10^{-9} seconds' duration, the time constant for a thermal effect to produce the negative resistance in *compensated* germanium must be less than this value.

value. The localization occurs because the neutralizing charge of the ionized impurity atoms is immobile, and also because the mean free path of the high energy carriers is so short. This situation is to be compared with avalanche breakdown in a gas where the discharge spreads to encompass the entire tube. Because of the localization of the conductivity modulation, many independent cryosars may be placed on one wafer of germanium. Fig. 5 shows an experimental array of 25 cryosars, interconnected in matrix fashion for testing purposes, on a wafer 1 cm² in area and 0.050 cm thick. Each cryosar can be turned on or off independently of the state of the adjacent cryosars.

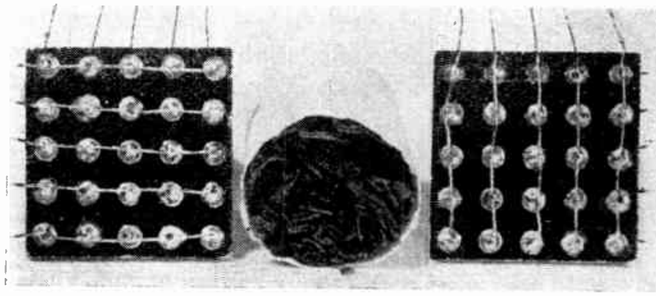

Fig. 5—Top and bottom views of experimental array of 25 cryosars interconnected in matrix fashion. The ohmic contacts to the p-type germanium are indium-alloyed. A cigarette is shown for size comparison.

TABLE I
PROPERTIES OF MATRICES

Matrix No.	From Crystal	Acceptor Density* N_A cm^{-3}	Net Acceptor Density† (N_A-N_D) cm^{-3}	Low-Field Prebreakdown Conductivity (mho cm^{-1})	Current Density at Peak Point (amp cm^{-2})	Breakdown Field (volts cm^{-1})	Sustaining Field (volts cm^{-1})
21	923	10.8×10^{15}	2.8×10^{15}	2.3×10^{-3}	5×10^{-1}	170	140
26	941	4.2×10^{15}	1.4×10^{15}	1.9×10^{-5}	5×10^{-3}	100	40
27	941	4.2×10^{15}	1.4×10^{15}	No data	6×10^{-3}	95	38
37	952	1.65×10^{15}	1.7×10^{14}	5.5×10^{-9}	1×10^{-6}	140	35
45	958	1.65×10^{16}	3.2×10^{14}	4.2×10^{-8}	3×10^{-6}	60	20
49	959	2.5×10^{16}	2.1×10^{16}	2.0×10^{-5}	2×10^{-3}	40	22
54	941	4.2×10^{16}	1.4×10^{16}	1.2×10^{-4}	3×10^{-2}	95	41
56	941	4.2×10^{15}	1.4×10^{15}	2.4×10^{-6}	6×10^{-3}	100	40
68	988	6.8×10^{15}	4.1×10^{15}	2.5×10^{-3}	7×10^{-1}	145	100
69	995	2.8×10^{16}	1.45×10^{16}	2.0×10^{-5}	6×10^{-3}	85	41

* The acceptor density was determined by growing uncompensated indium-doped control crystals, the reproducibility of which was ±10 per cent. This method assumes no interaction between the two types of dopants, which seems reasonable considering the impurity concentrations used.
† The net acceptor density was determined from the room-temperature resistivity.

For the bistable cryosar, both the breakdown field and the sustaining field are functions of the impurity densities in the compensated germanium. Table I lists the properties of a number of matrices of 25 cryosars each. The properties listed were determined by averaging over the 25 elements in each matrix. Variations in the breakdown and sustaining voltages between elements on one matrix have usually been ±10 per cent, and some of this variation can be accounted for by nonuniformities in wafer thickness. Included in this list are four matrices (26, 27, 54, and 56) made from the same crystal (941). While matrix 54 was fabricated using a heavily sandblasted germanium wafer, the other three were fabricated using etched wafers. The reproducibility of the breakdown and sustaining voltage for all four wafers is good. The sustaining voltages for the matrices listed in Table I have been plotted in Fig. 6 as a function of (N_A+N_D). Considering the uncertainty in the values of the impurity densities, the experimental points are consistent with a linear relationship between the sustaining voltage and (N_A+N_D). No such simple empirical relation has been found for the breakdown voltages in compensated germanium.

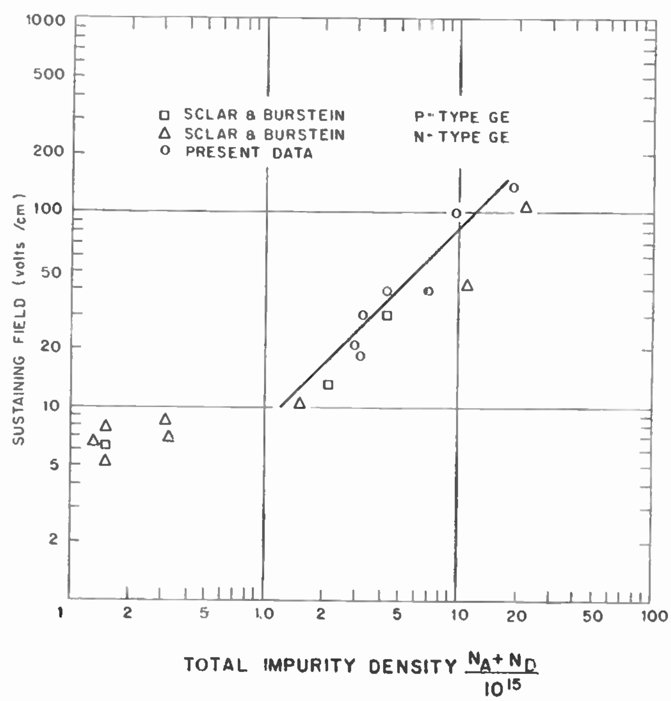

Fig. 6—The sustaining voltage as a function of total impurity density. Also plotted are data from Sclar and Burstein.[2]

Also plotted in Fig. 6 are the breakdown voltages reported by Sclar and Burstein.[2] Since their samples were not intentionally compensated, it has been assumed that the minority impurity density could be neglected with respect to the majority impurity. Their results indicate that the breakdown field for uncompensated germanium is reasonably insensitive to resistivity in the range between 10 and 20 ohm-cm. Our experiments with monostable cryosars agree with this conclusion. For larger impurity densities the breakdown voltage for uncompensated germanium increases with increasing density, and as can be seen in Fig. 6 the values from Sclar and Burstein seem to satisfy the same linear relationship obtained for compensated germanium.

The usual difficulties with junctions and contacts are not encountered in the cryosar since impact ionization is a bulk effect. Although some surface effects have been found while investigating breakdown in bars of compensated germanium, very little has been seen with the wafer geometry. Reproducibility seems to be almost entirely a matter of controlling the impurity densities, which is not difficult as a result of the techniques developed for transistor production. While only 5×5 arrays of cryosars have been built, the experimental results on these indicate that arrays with a density greater by an order of magnitude or more could be achieved and the elements in such arrays would be independent and nearly identical.

Switching Speed

The speed of both the monostable and bistable cryosars is limited by the turn-on time, the time to switch from high to low resistance by the impact ionization process. If the applied voltage barely exceeds the critical voltage necessary for avalanche, the turn-on time may be extremely long—of the order of microseconds or more. As the over-voltage is increased, the turn-on time decreases markedly.[9] In Fig. 7, trace *a* shows the cryosar current as a function of time (as measured by the voltage across the load resistor) after an input pulse 1.8 times the breakdown voltage was applied. The input pulse is shown in trace *b*. The time required for the current to increase from its initial value of 10^{-5} amp to 10^{-2} amp was less than 10 mμsec. This is a typical speed for the matrices of Table I. To explain the delay in the turn-on, Koenig has proposed a model which involves field emission at the contacts.[10] Our present results neither prove nor disprove the model. It is hoped that a further understanding of the phenomenon may permit a reduction of the switching time.

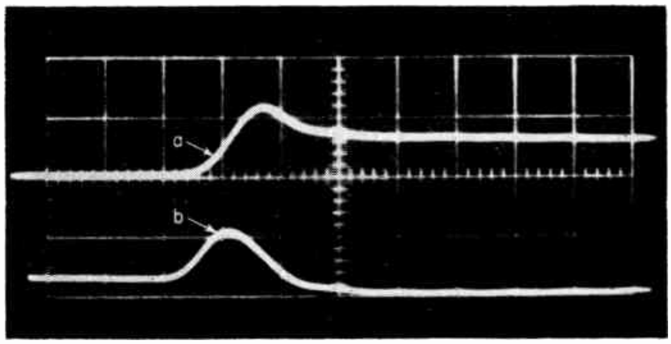

Fig. 7—Trace *a* is the voltage across the bistable-cryosar load resistor, after application of the input turn-on pulse *b*. The horizontal scale is 2×10^{-8} seconds per large division and the vertical scale is 2 volts per large division. The pulse shape is distorted by the vertical response of the oscilloscope; the actual duration of the input pulse is 2.5×10^{-8} sec.

When the voltage across the cryosar is reduced below that necessary to sustain the breakdown, the mobile carriers which had been created by impact ionization recombine with the impurity atoms, and the device again exhibits high impedance. The rate at which recombination occurs depends on the compensation of the sample. If there are N_C compensating impurities, there are always at least N_C majority impurity sites available for carrier recombination. The cross section for recombination of each individual site is sufficiently large,[11] however, that even for what is normally called uncompensated germanium ($N_C \simeq 2\times10^{13}$ cm^{-3}) no turn-off time could be discerned using a Tektronix 545 oscilloscope which should be able to detect times of the order of 10 mμsec. Recombination times of the order of 10^{-10} seconds are predicted for the heavily compensated bistable cryosars. These "de-ionization times" should be compared to the very much larger times for gas discharges.

Prebreakdown Characteristics

As can be seen from Table I, the low-field prebreakdown conductivity varies over five orders of magnitude for the samples of compensated germanium studied. This prebreakdown conductivity can be explained in terms of a mechanism for impurity conduction which has been proposed by Conwell[12] and Mott.[13] For the case of *p*-type germanium of interest here, it is assumed that the electrons in the acceptor levels, which originated from the compensating donor impurities, migrate under the influence of an electric field by tunneling to adjacent empty acceptor sites. Conwell has calculated in a rough way the mobility to be expected on the basis of this model by first determining the diffusion constant, assuming hydrogenic wave functions and uniform spacing

[9] This effect has been used at the RCA Laboratories, Princeton, N. J., to amplify pulses. M. C. Steele, "Pulse amplification using impact ionization in germanium," Proc. IRE, vol. 47, pp. 1109–1117; June, 1959.

[10] S. H. Koenig, "On the Nature of Electrical Conduction in Germanium at Low Temperatures; Non-Equilibrium Bulk and Contact Phenomena," presented at the International Conference on Solid-State Physics in Electronics and Telecommunications, Brussels, Belgium; June 2-7, 1958.

[11] S. H. Koenig, "Recombination of thermal electrons in *n*-type germanium below 10°K," *Phys. Rev.*, vol. 110, pp. 988–990; May 15, 1958.

[12] E. M. Conwell, "Impurity band conduction in germanium and silicon," *Phys. Rev.*, vol. 103, pp. 51–61; July 1, 1956.

[13] N. F. Mott, "On the transition to metallic conduction in semiconductors," *Can. J. Phys.*, vol. 34, pp. 1357–1368; December, 1956.

of the acceptors, and then using the Einstein relationship. For this last step the approximate relation $\mu = qD/kT$ was used, but for the heavily compensated crystals studied here, in which the acceptor levels are fairly full, the exact expression

$$\mu = qD \frac{d}{dE_F} \ln n_a \qquad (1)$$

must be used, where n_a is the concentration of electrons in the acceptor levels, and E_F the Fermi level. This gives for the conductivity

$$\sigma = qN_D \left(\frac{1}{2} \frac{q}{kT} \frac{N_A - N_D}{N_A} \right) (2r_s)^2 \frac{8 r_s E_{act}}{h \kappa a_H^*} \cdot \exp\left(-2r_s/\kappa a_H^*\right), \qquad (2)$$

where $2r_s$ is the spacing between acceptors, a_H^* is the Bohr radius for an electron of mass m^*, κ is the dielectric constant and E_{act} is the impurity activation energy. Taking the experimental value of 0.011 ev for E_{act} of indium, $m^*/m = 0.2$ in computing a_H^* (the same value that must be used to obtain the activation energy from the hydrogen model), and assuming $(4/3)\pi r_s^3 = 1/N_A$, the conductivity at 4.2°K for very low applied fields should be

$$\sigma = 1.1 \times 10^4 [N_D(N_A - N_D)/N_A^2] \cdot \exp(-2.93 \times 10^6 N_A^{-1/3}) \text{ ohm}^{-1} \text{cm}^{-1}. \qquad (3)$$

Fig. 8 shows the experimental values of $\ln [\sigma N_A^2/N_D(N_A - N_D)]$ plotted against $N_A^{-1/3}$ for the seven crystals thus far studied, together with the theoretical curve from the equation above. The agreement is perhaps better than should be expected considering the crudeness of the model and the experimental uncertainty in the impurity concentrations. Experiments on photoconductivity, which have been reported elsewhere,[14] have lent further evidence to Conwell's and Mott's model of impurity conduction.

At higher fields in the prebreakdown region the germanium may become non-ohmic, as in the case of sample *A* of Fig. 2. This may be due both to increasing numbers of carriers in the valence band and a changing mobility in this band.[3]

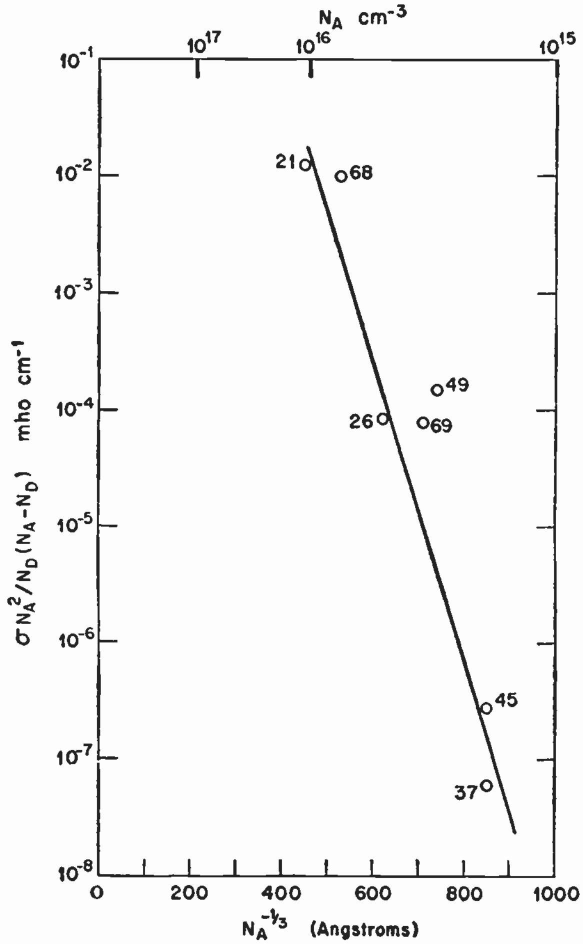

Fig. 8—Comparison of theoretical curve for impurity conduction with experimental points.

Effect of Light

If a bistable cryosar is illuminated so as to produce a photocurrent of, or larger than, the dark current at the peak point, the breakdown voltage is considerably reduced. Thus illumination can be used to control the operation of the bistable cryosar in a way somewhat analogous to the grid voltage in a thyratron. It is possible, for example, to adjust the electrical bias and pulse amplitude applied to a cryosar so that only when the element is illuminated can the pulse switch it from the high-resistance to the low-resistance state. The bias can also be adjusted so that illumination alone will switch the element from its high- to its low-resistance state. It should be pointed out, however, that while cryosar response to electrical pulses is of the order of millimicroseconds, it may require times of the order of milliseconds or longer for the light to affect the cryosar characteristics.

As shown in Table I, by proper selection of germanium the dark current at the peak point can be varied over at least five orders of magnitude. If the dark current at the peak point is larger than the photocurrent that flows as a result of illumination of the element, the breakdown voltage is not sensibly changed by light. Hence by proper selection of germanium resistivity and compensation, the operation of bistable elements can be made insensitive to illumination.

Fabrication Techniques

Satisfactory ohmic contacts have been made to *p*-type germanium, both compensated and not intentionally compensated, either by alloying indium or indium-gallium buttons, or by soldering, using Cerroseal solder and suitable flux. Matrices of elements have also been

[14] A. L. McWhorter, "Effect of light on impurity conduction in germanium," *Bull. Am. Phys. Soc., ser. II*, vol. 4, p. 186: March, 1959.

made by plating indium lines on each face of the wafer, with the lines on one face at right angles to those on the other, and then alloying these lines in to make ohmic contact. No noticeable difference has been observed in the electrical characteristics between cryosars made using the same bulk material and the different types of ohmic contacts discussed above.

Work is under way to make the ohmic contacts by evaporation, which seems to be the technique best suited for the production of large arrays of cryosars.[15] Since only the contacts are evaporated and the characteristics of the cryosar depend on the properties of the *substrate*, the reproducibility of cryosars produced by evaporation should be superior to that of other miniature evaporated computer components whose characteristics depend on the properties of the *evaporant*.

Applications

As described above, the monostable cryosar can perform the functions of an ordinary diode. While it is a bilateral device, its electrical characteristics in either direction are that of a diode in series with a battery. The bistable cryosar can be used as a memory element, or can be used as a multivibrator or flip-flop. It comes closest in static electrical characteristics to the *p-n-p-n* diode. The cryosar is also very fast. Preliminary results indicate that the cryosar will operate at pulse repetition frequencies well above 100 megapulses per second.[16]

Although there may be need for a high-speed replacement for diodes and bistable elements, the usefulness of the cryosar is greatly enhanced by the fact that large arrays with a very high density of cryosars can be built on single wafers of germanium. As indicated in the section on breakdown characteristics, the active region of each cryosar is limited to the volume directly between its two contacts since there are negligible fringing effects. Circuit considerations usually limit the density of cryosars. Since two adjacent cryosars on a wafer of germanium may be interconnected by external circuitry, care must be taken to place them sufficiently far apart so that voltages that are developed between them will not produce an electric field large enough to cause breakdown laterally between the cryosars. For arrays operated in matrix fashion and interconnected as shown in Fig. 5, a spacing between cryosars twice the thickness of the wafer (which is approximately the distance between the contacts of a given cryosar) has been found satisfactory. Thus for this type of application an array of over 625 independent cryosars could be fabricated on a wafer 1 cm² in area and 0.012 cm thick. Each cryosar could have contacts 0.012 cm in diameter, with a minimum separation of 0.028 cm between elements.

The fabrication of such large arrays in a single operation places extremely severe requirements on the reliability and reproducibility of the individual elements, but preliminary results described above with the experimental matrices of 25 elements indicate that these requirements can be met. Since the operation of the cryosar is based on impact ionization of impurities, which is a bulk phenomenon, and only ohmic contacts to the germanium are required, the usual difficulties with junctions and surfaces encountered in transistor-like devices are not present. Note that within the accuracy of measurement the breakdown and sustaining voltages for the elements of matrix 54 are almost identical to the corresponding voltages for matrix 56, although in the first case the cryosars were fabricated on a heavily sandblasted wafer and in the second case they were fabricated on an etched wafer. Only the prebreakdown resistivity was decreased when the surface was heavily sandblasted. Examination of Table I and Fig. 6 shows that reproducibility only requires controlling the impurity densities. Techniques which have been developed to produce uniform resistivity germanium, such as the zone-leveling technique[17] or the floating-crucible technique,[18] should be suitable for producing material which would meet the cryosar reproducibility requirements. There is good reason to expect essentially a 100 per cent yield.

Thus fabrication of arrays of well over 625 nearly identical cryosars on a wafer of germanium 1 cm² in area and 0.012 cm thick should be feasible both from the view of the elements being independent and every element in the array being operational. Wafers of the same thickness can be used to form arrays of nearly identical cryosars as large as necessary. If one used 0.040-cm-thick spacers between wafers of the size described above and required no other components, almost 200,000 cryosars could fit in a cubic inch. The cryosar could, of course, be built into microprinted circuitry.

Since the cryosars prepared in this large array would be scaled down in all three dimensions by a factor of about 8 as compared to the cryosars whose electrical characteristics are plotted in Figs. 1 and 2, maximum power dissipation in each unit could be limited in practical circuits to less than 10^{-6} watts. Thus even if every cryosar was in its low impedance state, 200,000 cryosars would have a maximum power dissipation of 0.2 watt, which could be handled with a very reasonable size liquid helium refrigerator.

One application for which the use of diodes is marginal, and for which the monostable cryosar seems well

[15] *Note added in Proof:* Matrices have been successfully evaporated with densities as high as 635 per cm² (1024 cryosars on a wafer 0.5×0.5 inch).

[16] R. C. Johnston of this laboratory, private communication.

[17] D. C. Bennett and B. Sawyer, "Single crystals of exceptional perfection and uniformity by zone leveling," *Bell Sys. Tech. J.*, vol. 35, pp. 637–660; September, 1956.

[18] W. F. Leverton, "Floating crucible technique for growing uniformly doped crystals," *J. Appl. Phys.*, vol. 29, pp. 1241–1245; August, 1958.

suited, is the function table. This is a circuit in which a binary variable introduced on a set of x conductors is transformed by appropriate diode connections into the specified function of the variable on a set of y conductors. Since many sequential operations are reduced to a few, the use of function tables greatly increases the speed of computation. Feasibility has been experimentally shown for building function tables with suitably connected matrices of monostable cryosars. Because of their compactness, reliability, low cost, high speed, and high off-to-on impedance ratio, cryosars should make large tables practicable.

Since the breakdown voltage of the bistable cryosar can be designed to be light sensitive, a dc bias and pulse amplitude can be chosen so that only illuminated elements will be switched from their high impedance to low impedance state when the pulse is applied. In a matrix of bistable cryosars, connected as in Fig. 5, selective illumination can be used to render only certain elements electrically effective. Such a matrix could be used as a universal function table, since by changing a mask to the incident light the function could be changed. As pointed out previously, however, while cryosar response to electrical pulses is of the order of millimicroseconds, it may require times of the order of milliseconds or longer for the light to affect the cryosar characteristics. Other possible applications of matrices of light sensitive cryosars are the reading of buffer film storage into a computer and pattern recognition.

As digital computers become faster, not only is it necessary to have faster and more reliable devices, but also more compact devices so that lead length and size may be reduced. It will also be advantageous to fabricate the device and the circuit at the same time. The cryosar, both in its bistable and monostable forms, perhaps combined with evaporated superconductive elements, gives promise of meeting the requirements of these higher-speed computers.

Acknowledgment

We wish to thank R. H. Kingston for suggesting originally the use of matrix arrays of impact-ionization switching elements and for other valuable suggestions during the course of the work. We are grateful to R. H. Baker for many stimulating and encouraging discussions, and to R. J. Keyes for help and advice with the photoconductivity studies. Some of the data were taken by C. R. Grant and J. H. R. Ward; the cryosars were fabricated by W. H. Laswell and Mrs. M. L. Barney. We are indebted to P. L. Moody and A. E. Paladino for growing the compensated germanium.

Parametric Energy Conversion in Distributed Systems*

G. M. ROE† AND M. R. BOYD†, MEMBER, IRE

Summary—**Traveling wave type parametric amplifiers have been proposed which utilize transmission lines having reactance varied by a propagating wave. An analysis indicates that in systems with little or no dispersion, pumping will not result in exponential gain of an applied signal but in a conversion of energy to a multiplicity of cross-product frequencies. Two physical models of such wide-band systems are discussed and the special case of a zero-dispersion line is analyzed. The results indicate that although the total energy increases exponentially with distance, the wave becomes extremely rich in high frequencies. Although such systems would not be useful as amplifiers, use as frequency converters is suggested.**

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† Electron Physics Res. Dept., General Electric Res. Lab., Schenectady, N. Y.

Introduction

In order to extend the bandwidth capabilities of parametric amplifiers and frequency converters, considerable attention is presently being given to traveling wave type devices. These include ferromagnetic amplifiers and electron beam devices as well as variable capacitor diode configurations. An added complexity of the distributed system requires not only that frequencies be properly related (as in a resonant system), but that the velocities of propagation of pump, signal, and idler also be related in a prescribed manner. One possibility for maintaining the proper synchronism is to employ a medium with little or no dispersion. It is the purpose of this paper to examine in detail the mechanism of energy transfer for the special case of a dispersionless system. Although the results are for the special case considered, the implications extend to more practical systems. In brief, the analysis indicates that in a system where all frequencies propagate with the same velocity, the parametric or pump energy will not produce exponential amplification of an applied signal but will convert to an infinite set of frequencies comprised of cross-products between the pump and frequencies existing on the line.

In a practical system, of course, impedance or propagating characteristics will eventually limit the frequency content of the wave. It should, however, be emphasized that frequencies which are generated during the pumping process and which lie within the pass bands of the medium should not be neglected. This is particularly important if low dispersion media are to be considered. It would be of interest to consider some physical models which would exhibit the phenomenon described above.

Bridges[1] has proposed an electron beam parametric amplifier in which the electronic reactance at the gap of a resonant cavity is modulated by a pump signal at twice the resonant frequency of the cavity. To extend the bandwidth of this type of amplifier the pump and signal cavities would be replaced by ridge waveguides. This device is a floating drift tube traveling wave klystron and has a sheet beam moving transverse to the direction of wave propagation. Analytically, this traveling wave klystron may be described as a transmission line having a distributed shunt capacitance which is electronically modulated in synchronism with the traveling wave but at twice its frequency. The equivalent transmission line with incremental sections is shown in Fig. 1.

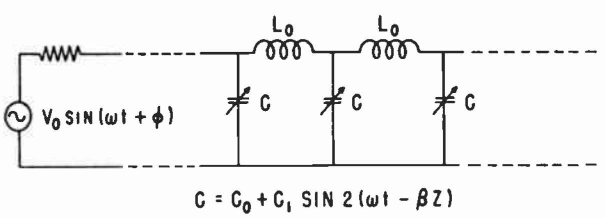

Fig. 1—Equivalent transmission line.

A second physical model of importance is the coaxial or waveguide system which is periodically or continuously loaded with a semiconductor material the capacity of which is a function of voltage. Periodically loading the transmission line with lumped capacitors will, of course, introduce stop bands which will inhibit the creation of some cross-product frequencies. However, the periodicity of the loading may permit many of the frequencies generated to propagate in near-synchronism and thus absorb pump energy.

The equivalent circuit of Fig. 1 has been used in three recent papers by Cullen,[2] by Tien and Suhl,[3] and by Louisell and Quate.[4] These analyses lead to solutions which predict exponential growth of an input wave with distance. Although the difference frequency between pump and signal is considered, the sum frequency and higher components are neglected. While these frequencies may or may not have any intrinsic interest it is possible that they may interact with the pump to generate more of the fundamental component and hence should not be ignored. Neglecting the higher order frequencies implicitly characterizes the system as being dispersive, and this may indeed by the objective of the above-mentioned analyses. It is felt, however, that if low dispersion systems are specified for the purpose of efficient interaction processes, then the possibility of higher frequencies on the system should be recognized.

The equation describing propagation along the circuit of Fig. 1 is

$$\frac{\partial^2 V}{\partial Z^2} - (L_0/l^2)\,\frac{\partial^2(CV)}{\partial t^2} = 0 \tag{1}$$

where L_0/l and C_0/l are the inductance and capacitance per unit length of line. For the purpose of analysis it will be assumed that a signal of frequency ω is applied at the input and that the capacity is varied in synchronism and at twice the frequency. This relation of frequencies is for simplicity but the behavior of more general cases may be implied from the analysis. The variable capacitance is given by

$$C = C_0 + C_1 \sin 2(\omega t - \beta z) \tag{2}$$

where

$$\beta = (L_0C_0)^{1/2}\omega/l. \tag{3}$$

The Form of the Solution

One may ask what happens to the fundamental component of the solution of (1) when we allow the solution to contain two harmonics, three harmonics, etc. In Fig. 2 the dashed curve 1 is the exponential growth curve of Cullen or of Tien and Suhl, obtained by allowing the solution to contain only the fundamental component. Curve 2 allows for a term of frequency 3ω in addition to the fundamental. The amplitude of the fundamental now oscillates instead of growing exponentially. The pumping wave still feeds energy into the system, but a

[1] T. J. Bridges, "A parametric electron beam amplifier," Proc. IRE, vol. 46, pp. 494–495; February, 1958.

[2] A. L. Cullen, "A traveling-wave parametric amplifier," *Nature*, vol. 181, p. 332; February 1, 1958. Cullen's circuit has the inductance variable rather than the capacitance, but this makes no essential change in the mathematical formulation.

[3] P. K. Tien and H. Suhl, "A ferromagnetic traveling wave amplifier," Proc. IRE, vol. 46, pp. 700–706; April, 1958.

[4] W. H. Louisell and C. F. Quate, "Parametric amplification of space charge waves," Proc. IRE, vol. 46, pp. 707–716; April, 1958.

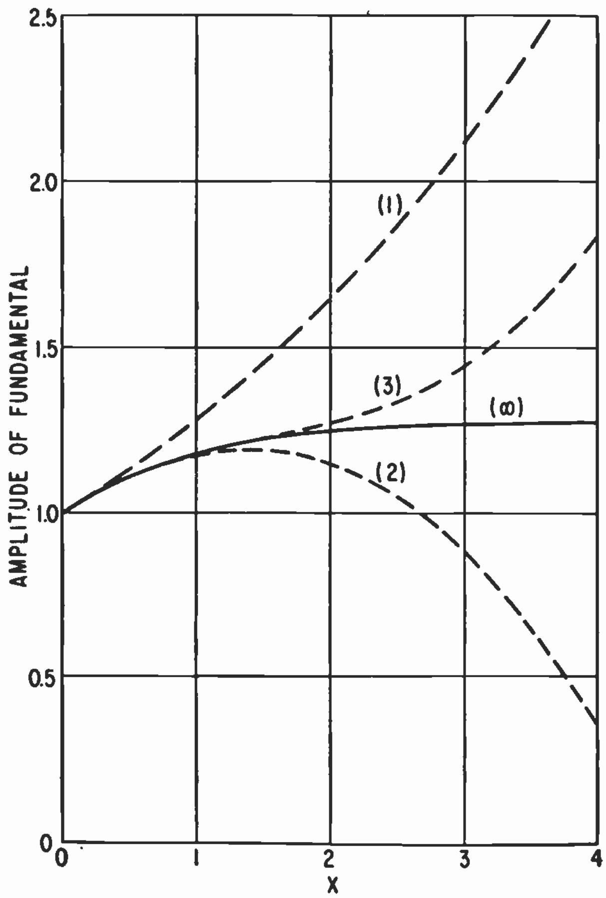

Fig. 2—Growth curve for the fundamental component when the voltage function is allowed to contain 1, 2, 3, and ∞ harmonics. Starting phase $\phi=0$.

large part of the energy present seems to oscillate back and forth between the two harmonics that we have allowed to be present. With three harmonics present (frequencies ω, 3ω, and 5ω) the growth again becomes exponential as shown by curve 3. The rate of growth is, however, much reduced. The solid curve labeled ∞ in Fig. 2 is the exact solution obtained below. The successive approximations shown in Fig. 2 seem to approach the correct solution, but the rate of convergence is so slow that this method of successive approximations is not very useful.[5]

This problem turns out to be one which is easier to solve if the Fourier analysis is made after the solution has been found, rather than in the process of trying to find a solution. In developing the required solution it will be convenient to replace the variables t and z by a phase variable

$$\tau = \omega t - \beta z \tag{4}$$

and a reduced distance variable

$$x = \beta z C_1/C_0. \tag{5}$$

The partial derivative operators of (1) must then be replaced by

$$\frac{\partial}{\partial t} = \omega \frac{\partial}{\partial \tau}$$

$$\frac{\partial}{\partial z} = (C_1/C_0)\beta \frac{\partial}{\partial x} - \beta \frac{\partial}{\partial \tau}\cdot$$

In practical cases the modulation index, C_1/C_0, will be small, and we choose to expand the voltage in powers of this ratio.

$$V = V_0\{F(x, \tau) + (C_1/C_0)H(x, \tau) + \cdots\}. \tag{6}$$

It is unnecessary to carry along terms in the square of C_1/C_0 and higher powers because all of the interesting results appear already in the leading term, $F(x, \tau)$. Note that the variable x depends on C_1/C_0. Even the term in H will be carried along here only to demonstrate that no significant new effect appears when C_1/C_0 is not vanishingly small.

With the above change in variables the original differential equation (1) is replaced by the set,

$$\frac{\partial}{\partial \tau}\left\{\frac{\partial}{\partial \tau}[F \sin 2\tau] + 2\frac{\partial F}{\partial x}\right\} = 0 \tag{7}$$

$$\frac{\partial}{\partial \tau}\left\{\frac{\partial}{\partial \tau}[H \sin 2\tau] + 2\frac{\partial H}{\partial x}\right\} = \frac{\partial^2 F}{\partial x^2}, \tag{8}$$

plus similar equations for the ignorable higher order terms in the expansion (6). Note that the new variables have made it possible to factor out one of the partial derivatives. The remaining first-order equations can be solved by taking Laplace transforms relative to the x variable and then using standard formulas to write solutions for the τ equations in the form of definite integrals. The process of taking the inverse Laplace transform introduces delta functions and derivatives of delta functions which simplify the evaluation of the definite integrals. The solutions obtained in this way are

$$F(x, \tau) = \frac{e^{-2x}\sin\tau\cos\phi + e^{-x}\cos\tau\sin\phi}{[1-(1-e^{-2x})\sin^2\tau]^{3/2}} \tag{9}$$

$$\begin{aligned} H(x, \tau) = {} & \frac{[\tan^{-1}(\tan\tau) - \tan^{-1}(e^{-x}\tan\tau)]}{2[1-(1-e^{-2x})\sin^2\tau]^{5/2}} \\ & \cdot\{e^{-x}[\cos^2\tau - 2e^{-2x}\sin^2\tau]\cos\tau\sin\phi \\ & + e^{-2x}[2\cos^2\tau - e^{-2x}\sin^2\tau]\sin\tau\cos\phi\}. \end{aligned} \tag{10}$$

[5] The successive approximations were obtained as follows:

In (1) we assume a Fourier series solution containing all odd multiples of $(\omega t-\beta z)$. The Fourier coefficients are functions of z and satisfy an infinite set of simultaneous differential equations. The Laplace transforms of the coefficients then satisfy an infinite set of algebraic equations; and the transform of the fundamental amplitude, for example, can be written formally as the ratio of two infinite order determinants. For the nth approximation, each of these infinite order determinants is replaced by a finite order determinant consisting of its first n rows and n columns. This gives, for the transform of the fundamental amplitude, a polynomial fraction of order n, and the n roots of the denominator must be determined numerically in order to evaluate the inverse transform.

It may be verified by direct substitution that these are the solutions of (7) and (8). The constants of integration have been chosen so that for $x=0$, $F=\sin(\tau+\phi)$ and $H=0$. In (10) the multiple valued arctangent functions are to be chosen in such a way that

$$\left[\tan^{-1}(\tan\tau) - \tan^{-1}(e^{-x}\tan\tau)\right]$$

approaches $x\sin\tau\cos\tau$ for x approaching zero.

The above formulas could be used to plot the functions F and H as functions of τ for various values of x. However, considerable computational labor may be saved by noting that for large x, F and H can be approximated by some universal curves. When x is large, both F and H become small except when their denominators are small; that is, when $\sin^2\tau$ is close to unity. Suppose we set

$$\tau = \frac{\pi}{2} + \theta e^{-x}, \tag{11}$$

and let x become large. Then

$$F \cong \frac{e^x[\cos\phi - \theta\sin\phi]}{[1+\theta^2]^{3/2}} \tag{12}$$

$$H \cong \frac{e^x\tan^{-1}\theta}{2[1+\theta^2]^{5/2}}\left\{(1-2\theta^2)\cos\phi - \theta(2-\theta^2)\sin\phi\right\}. \tag{13}$$

Increasing τ by π changes the sign of F and H.

Consider first the case of a starting phase $\phi=0$. The shape curves for large x are plotted in Fig. 3. H is always smaller than F so that in the practical cases when C_1/C_0 is small we can ignore H and concentrate our attention on F. As x increases, the height of the pulse increases exponentially, but the pulse width decreases exponentially. For large x the voltage wave consists of a sequence of alternately positive and negative pips. As each pip moves down the transmission line it becomes higher and narrower (see Fig. 4)[6], as shown schematically in Fig. 5.

The corresponding wave shape for a starting phase $\phi=\pi/2$ is shown in Fig. 6. Here also the voltage wave at large distances is practically negligible except when τ is an odd multiple of $\pi/2$. The main difference is that for $\phi=\pi/2$ each isolated pulse has both a maximum and a minimum.

Harmonic Content

In making an harmonic analysis we will treat C_1/C_0 as small enough so that the pulse shape is essentially given

[6] It is instructive to illustrate the change in waveform with rough pencil sketches. From (7), the change in waveform with distance is given by

$$\frac{\partial F}{\partial x} = -\frac{1}{2}\frac{\partial}{\partial\tau}[F\sin 2\tau].$$

For any F it is easy to sketch qualitative graphs of F, $\sin 2\tau$, their product, and the negative derivative. A sequence of such graphs shows clearly how an initial sine wave can be deformed into a sequence of narrow pips. A typical sketch from such a sequence is shown in Fig. 4.

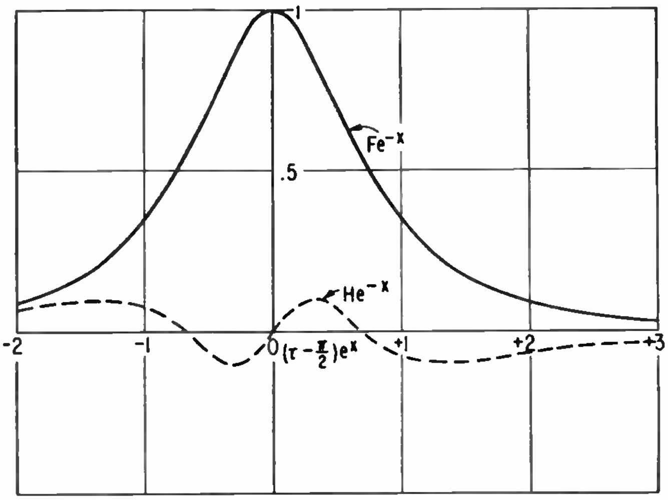

Fig. 3—Shape curves for the voltage pulse functions when x is large. Starting phase $\phi=0$.

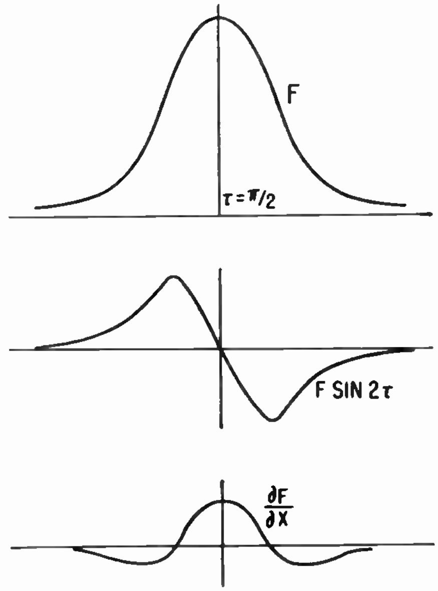

Fig. 4—Sketches for determining the change in wave shape with distance. For any given wave pulse, F, the change in wave pulse with distance, $\partial F/\partial x$, is proportional to the negative τ derivative of the product $F\sin 2\tau$.

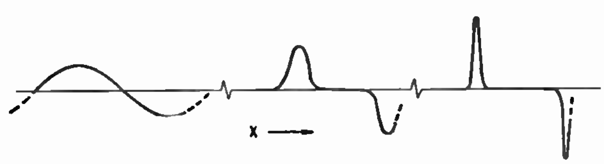

Fig. 5—Change in voltage wave shape with increasing distance.

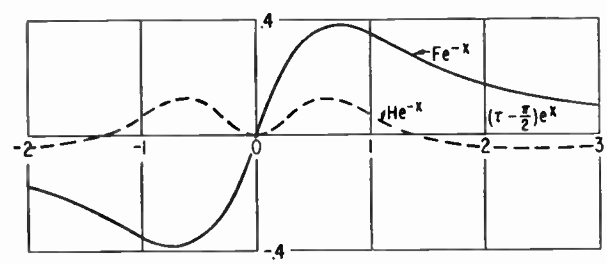

Fig. 6—Shape curves for the voltage pulse functions when x is large. Starting phase $\phi=\pi/2$.

by the function F. Only odd harmonics appear in the Fourier expansion of F.

$$F(x, \tau) = \sum_{n \text{ odd}} \{A_n(x) \cos n(\tau - \pi/2) \cos \phi - B_n(x) \sin n(\tau - \pi/2) \sin \phi\}. \quad (14)$$

The Fourier coefficients are

$$A_n(x) = \frac{4}{\pi}\int_0^{\pi/2} \frac{e^{-2x} \sin \tau \cos n(\tau - \pi/2) d\tau}{[1 - (1 - e^{-2x}) \sin^2 \tau]^{3/2}}, \quad (15)$$

$$B_n(x) = -\frac{4}{\pi}\int_0^{\pi/2} \frac{e^{-x} \cos \tau \sin n(\tau - \pi/2) d\tau}{[1 - (1 - e^{-2x}) \sin^2 \tau]^{3/2}}. \quad (16)$$

If n is not too large these integrals can be expressed simply in terms of the complete elliptic integrals $E(k)$ and $K(k)$. For example, with

$$k^2 = 1 - e^{-2x}, \quad (17)$$

the first few coefficients are

$$A_1 = \frac{4}{\pi}\{K - (K - E)/k^2\} \quad (18)$$

$$B_1 = \frac{4}{\pi} e^{-x}\{(K - E)/k^2\} \quad (19)$$

$$A_3 = \frac{4}{\pi}\{(-3 + 11k^{-2} - 8k^{-4})K + (-7k^{-2} + 8k^{-4})E\} \quad (20)$$

$$B_3 = \frac{4}{\pi} e^{-x}\{(-5k^{-2} + 8k^{-4})K + (k^{-2} - 8k)E\}. \quad (21)$$

The curve marked ∞ in Fig. 2 is a plot of the fundamental component, A_1, based on (18).

In Fig. 7 are plotted several of the coefficients, $A_n(x)$, against the reduced variable ne^{-x}. With this choice of abscissa all of the higher order components lie on the same curve. When n is large in (15) the main contribution to the integral comes from the range τ near $\pi/2$. If we let both n and x become large in such a way that ne^{-x} remains finite, the integral reduces to one of the integral expressions for the Bessel function of imaginary argument. Thus, for x large

$$A_n(x) \cong \frac{4}{\pi} ne^{-x}K_1(ne^{-x}). \quad (22)$$

The solid curve in Fig. 7 is based on this result. For a given, fixed x, Fig. 7 can be treated as a spectrum analysis of the voltage pulse. Note that all of the harmonics approach the same limiting value, $4/\pi$, for x sufficiently large. The curves of Fig. 7 have been replotted on a normal distance scale in Fig. 8. The value of x required to raise the nth harmonic up to the same level as the original fundamental amplitude is approximately $x = \ln(2n)$.

For a starting phase $\phi = \pi/2$ the spectrum is somewhat different. When n and x are large, (16) reduces to

$$B_n(x) \cong \frac{4}{\pi} ne^{-x}K_0(ne^{-x}). \quad (23)$$

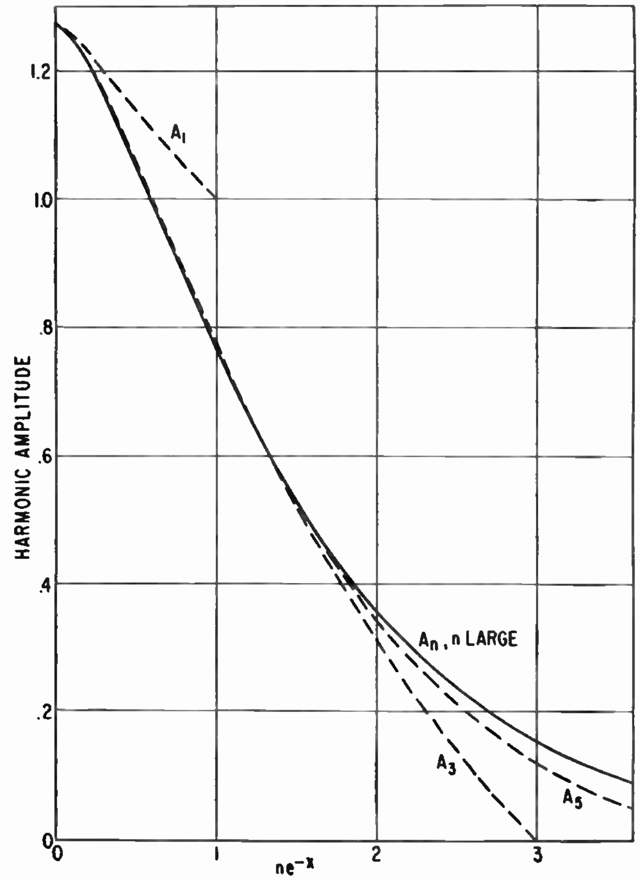

Fig. 7—Harmonic amplitudes as a function of ne^{-x} for starting phase $\phi = 0$.

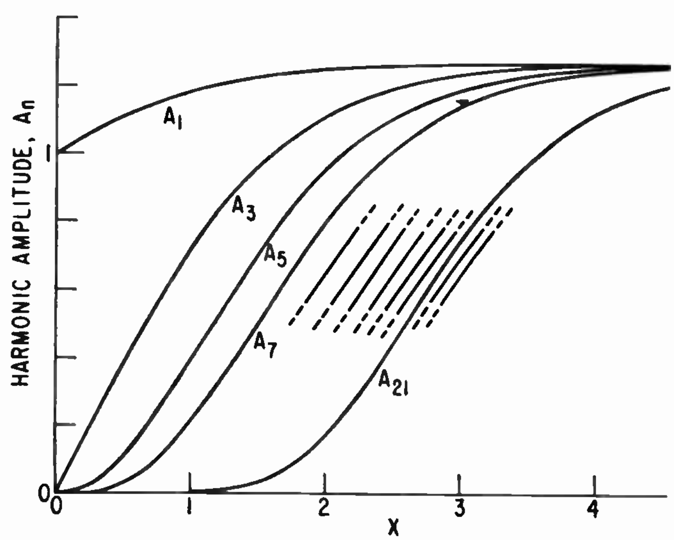

Fig. 8—Harmonic amplitudes as a function of distance, x, for starting phase $\phi = 0$.

This is plotted as a solid line in Fig. 9. Here the fundamental amplitude decreases with increasing distance. The higher harmonics all start out increasing with distance, but eventually decrease with increasing distance. The nth harmonic reaches its maximum amplitude at a distance $x \cong \ln(3n/2)$.

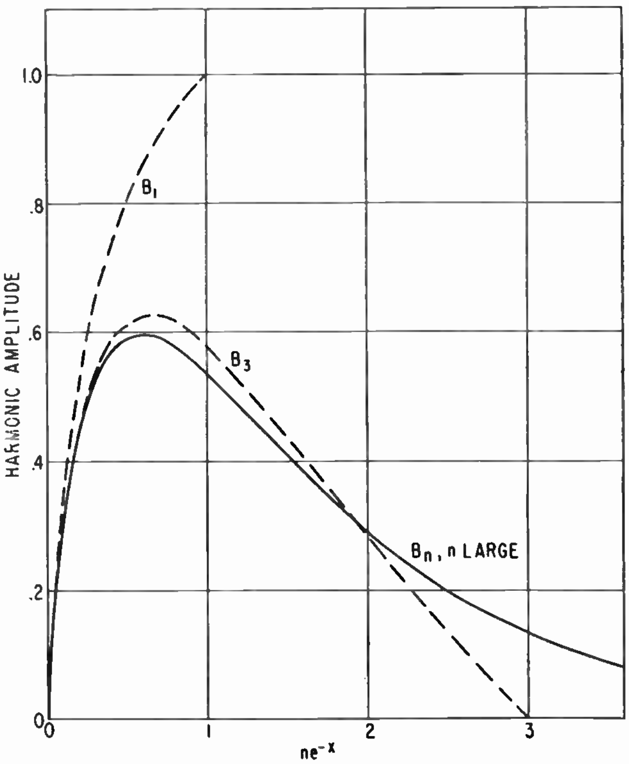

Fig. 9—Harmonic amplitudes as a function ne^{-x} for starting phase $\phi = \pi/2$.

Total Power

The current and voltage are related by

$$i = V - \frac{C_1}{C_0} \frac{\partial}{\partial x} \int^{\tau} V(x, \tau')d\tau'.$$

Hence, for the case of small C_1/C_0, which we have been studying, the current is in phase with the voltage. The power will therefore be proportional to the square of the voltage, or to the square of F. We can define a total power

$$P = \frac{1}{\pi} \int_0^{2\pi} F^2(x, \tau)d\tau$$

which is normalized to unit power at $x = 0$. With (9) for F this integrates to

$$P = \tfrac{1}{4}(3e^x + e^{-x}) \cos^2 \phi + \tfrac{1}{4}(e^x + 3e^{-x}) \sin^2 \phi.$$

Thus, although each of the harmonic amplitudes is bounded, an increasing number of harmonics are excited with increase in distance and the total power increases exponentially. For a starting phase $\phi = \pi/2$ the total power decreases at first, but reaches a minimum and thereafter increases exponentially.

Discussion

From the solution that has been obtained, it is clear that any device which is adquately described by the equivalent circuit of Fig. 1 will act as a good harmonic generator, but will not serve as an amplifier. The energy which is extracted from the pump goes toward the generation of the higher harmonics rather than into increased power in the fundamental. In order to get exponential gain in the fundamental it would be necessary to suppress the higher harmonics. One can do this in the equivalent transmission line by shunting across each of the variable capacitors a tank circuit which is resonant at the fundamental frequency. All of the higher harmonics would then be shorted out and the solutions proposed by Cullen[2] and by Tien and Suhl[3] would be valid.

The beam type device discussed by Louisell and Quate[4] is more difficult to analyze. The finite size of the beam introduces dispersion and the equivalent circuit of Fig. 1 is really applicable only at a particular frequency. By working with the space charge reduction factors Louisell and Quate take proper account of the effects of dispersion, but they follow Tien and Suhl in ignoring the effect of the higher harmonics. When higher harmonics are ignored they find that the net effect of dispersion in changing the relative phase velocities is to reduce the interaction between the pumping wave and the signal wave and make it more difficult to get exponential gain. The greater the dispersion, the higher is the minimum modulation index required for amplification. When the higher harmonics are not ignored we have found that in a system with no dispersion the interaction between the higher harmonics and the pumping frequency prevents any exponential gain in the fundamental component. If some dispersion is introduced into such a system one would expect the change in relative phase velocities to reduce the interaction between the harmonics and the pumping frequency and thereby tend to restore the possibility of amplification in the fundamental component. This result is just opposite to the direct effect of dispersion noted by Louisell and Quate.

Since the various consequences of dispersion can be both favorable and unfavorable for amplification of the fundamental component it is difficult to make any predictions as to the net effect of dispersion. For the actual systems described by Louisell and Quate the amplification will certainly be less than that predicted by their analysis, which ignores the higher harmonics; but it will be better than that predicted by our analysis, which takes no account of dispersion. Whether or not there is any exponential gain left will probably depend on the exact shape of the dispersion curve. A more detailed analysis is required for this type of distributed system.

A Discussion of Sampling Theorems*

D. A. LINDEN†, ASSOCIATE MEMBER, IRE

Summary—**The convolution theorem of Fourier analysis is a convenient tool for the derivation of a number of sampling theorems. This approach has been used by several authors to discuss first-order sampling of functions whose spectrum is limited to a region including the origin ("low-pass" functions). The present paper extends this technique to several other cases: second-order sampling of low-pass and band-pass functions, quadrature and Hilbert-transform sampling, sampling of periodic functions, and simultaneous sampling of a function and of one or more of its derivatives.**

Introduction

Several sampling theorems have appeared in the engineering literature.[1-5] These may be derived in a particularly perspicuous manner by means of the convolution theorem of Fourier analysis. The sampling process is regarded as a multiplication by a periodic sequence of δ-functions, its counterpart in the frequency domain being a convolution by a train of equispaced δ-functions. Interpolation—the recovery of the original signal from its sample values—is viewed in the frequency domain as a process of reconstructing the original spectrum by means of a spectral "window." The corresponding time domain operation consists of the convolution of the sample impulses with the inverse Fourier transform of the window function. This approach has been used by a number of authors[6-8] to discuss the equispaced sampling of low-pass functions. It is the purpose of this paper to present a consistent set of heuristic derivations for a number of additional sampling theorems.

The following transform definitions will be used:

$$F(f) = \int_{-\infty}^{+\infty} f(t)e^{-i\omega t}dt, \qquad \omega \equiv 2\pi f$$

$$f(t) = \int_{-\infty}^{+\infty} F(f)e^{i\omega t}df.$$

It will be convenient to use the notation

$$a(t) * b(t) \equiv \int_{-\infty}^{+\infty} a(\tau)b(t-\tau)d\tau.$$

Following the nomenclature of Kohlenberg,[2] sampling of a time function[9] will be designated as first-order if the sample points are equispaced. Second-order sampling involves two interleaved sequences of equispaced sampling points.

Sampling of Low-Pass Functions

The simplest case is that of a time function $f(t)$ whose spectrum $F(f)$ is limited to $-W \leq f \leq W$. The result of sampling the function at regular intervals spaced τ seconds apart is[10]

$$\hat{f}(t) = f(t)\sum_n \delta(t-n\tau) = \sum_n f(n\tau)\delta(t-n\tau). \qquad (1)$$

The transform of

$$\sum \delta(t-n\tau) \text{ is } \sum \frac{1}{\tau}\delta\left(f-\frac{n}{\tau}\right).$$

Multiplication in the time domain corresponds to convolution in the frequency domain, and the first equality of (1) leads to

$$\hat{F}(f) = F(f) * \sum_n \frac{1}{\tau}\delta\left(f-\frac{n}{\tau}\right)$$
$$= \sum_n \frac{1}{\tau}F\left(f-\frac{n}{\tau}\right). \qquad (2)$$

Apart from the weighting factor $1/\tau$, $\hat{F}(f)$ is seen to consist of replicas of $F(f)$ centered on the spectral lines $\delta(f-n/\tau)$, as illustrated in Fig. 1.[11] The possibility of recovering the original spectrum is insured if $1/\tau \geq 2W$; equality is permissible if $F(f)$ does not contain a δ-func-

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† Stanford Electronics Labs., Stanford University, Stanford, Calif.

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[9] All time functions are assumed to be real unless specifically designated as being complex.

[10] All summations are from $-\infty$ to $+\infty$ unless otherwise stated.

[11] $F(f)$ is in general a complex function and is indicated symbolically in Fig. 1 (a). Weighting factors such as $1/\tau$ will be indicated as shown in Fig. 1 (b).

tion at $f=W$. Assuming that sampling takes place at the lowest permissible rate, one has $1/\tau=2W$. The original spectrum may be recovered by multiplying $\hat{F}(f)$ by the spectral window function $S(f)$ shown in Fig. 1(c). The equivalent operation in the time domain is the convolution of $f(t)$ by the inverse Fourier transform $s(t)$ of $S(f)$, *i.e.*,

$$f(t) = s(t) * \sum_n f(n\tau)\delta(t - n\tau) = \sum_n f(n\tau)s(t - n\tau).$$

Substituting $\tau=1/2W$ and the functional form of $s(t)$,

$$f(t) = \sum_n f\left(\frac{n}{2W}\right)\frac{\sin 2\pi W\left(t - \frac{n}{2W}\right)}{2\pi W\left(t - \frac{n}{2W}\right)}. \quad (3)$$

The low-pass function $f(t)$ may also be subjected to second-order sampling. The two interlaced sampling trains

$$\sum_n \delta\left(t - \frac{n}{W}\right)$$

and

$$\sum_n \delta\left(t - \frac{n}{W} - \alpha\right)$$

will be designated by the letters A and B, respectively. The sampled functions are

$$f_A(t) = \sum_n f\left(\frac{n}{W}\right)\delta\left(t - \frac{n}{W}\right) \quad (4a)$$

and

$$f_B(t) = \sum_n f\left(\frac{n}{W} + \alpha\right)\delta\left(t - \frac{n}{W} - \alpha\right) \quad (4b)$$

and the corresponding spectra are given by

$$F_A(f) = F(f) * \sum_n W\delta(f - nW) \quad (5a)$$

$$F_B(f) = F(f) * \sum_n W\left(\frac{1}{\gamma}\right)^n \delta(f - nW) \quad (5b)$$

$$\gamma \equiv \exp(i2\pi\alpha W) \equiv \exp i\beta. \quad (5c)$$

The results of these convolutions are easily visualized: sketches of the spectra are shown in Fig. 2.[12] Since all time functions involved in this discussion are real, it suffices to consider their spectra for positive frequencies only. The spectral window functions $S_A(f)$ and $S_B(f)$ may be determined by the requirement

$$F_A(f)S_A(f) + F_B(f)S_B(f) = F(f), \quad 0 < f < W. \quad (6)$$

[12] Each spectrum is shown as the sum of two components which correspond to the convolutions of $F(f)$ with different spectral lines of the sampling function.

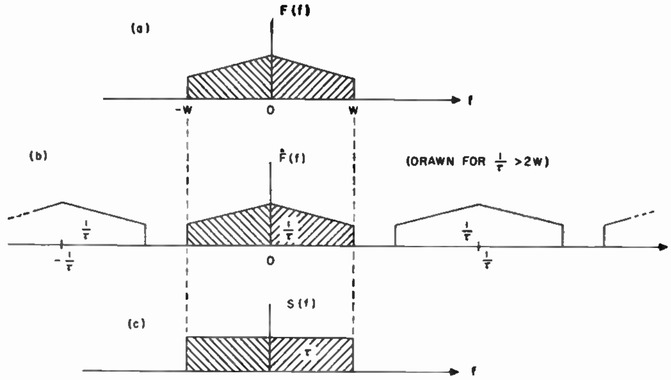


Fig. 1—First-order sampling of low-pass function.

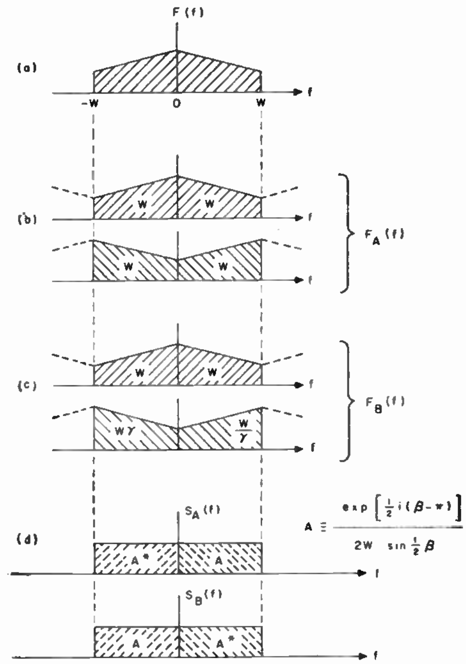


Fig. 2—Second-order sampling of low-pass function.

Inspection of Fig. 2 yields[13]

$$WS_A(f) + WS_B(f) = 1$$

$$WS_A(f) + \frac{W}{\gamma}S_B(f) = 0$$

whence

$$S_A(f) = S_B^*(f) = \frac{\exp\left[i\left(\frac{1}{2}\beta - \frac{\pi}{2}\right)\right]}{2W\sin\frac{1}{2}\beta}; \quad (0 < f < W).$$

[13] It is shown in Appendix I that these equations follow uniquely from (6).

Since the corresponding time functions are real, $S_{A,B}(-f) = S^*_{A,B}(f)$. The inverse Fourier transforms of $S_A(f)$ and $S_B(f)$ are the interpolating functions

$$s_A(t) = s_B(-t) = \frac{\cos(2\pi Wt - \pi\alpha W) - \cos\pi\alpha W}{2\pi Wt \sin\pi\alpha W}. \quad (7a)$$

Finally,

$$\begin{aligned} f(t) &= s_A(t) * f_A(t) + s_B(t) * f_B(t) \\ &= \sum_n f\left(\frac{n}{W}\right) s_A\left(t - \frac{n}{W}\right) \\ &\quad + f\left(\frac{n}{W} + \alpha\right) s_A\left(-t + \frac{n}{W} + \alpha\right). \quad (7b) \end{aligned}$$

With $\alpha = 1/2W$, (7) reduces to (3).

Sampling of Band-Pass Functions

The spectrum is assumed to occupy the range $W_0 \leq |f| \leq (W_0 + W)$, as sketched in Fig. 3(a). In general, second-order sampling must be used,[14] and (4), (5) apply. The results of the convolutions are shown in Fig. 3(b) and 3(c). The spectral window functions $S_A(f)$ and $S_B(f)$ which are required to restore the original spectrum may be computed by a procedure similar to that leading to (6).[15] The result is indicated in Fig. 3(d). The corresponding interpolating functions are[16]

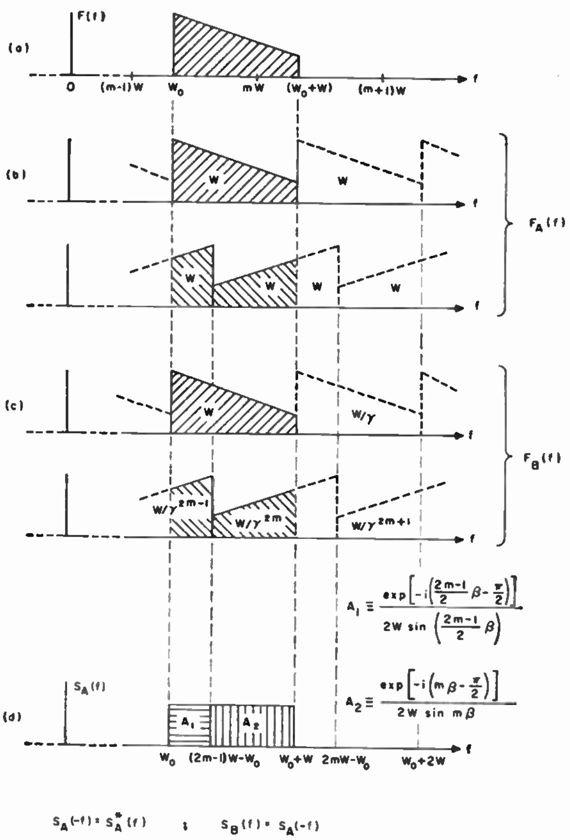


Fig. 3—Second-order sampling of band-pass function.

$$s_A(t) = \left\{ \begin{aligned} &\frac{\cos\left[2\pi m\alpha W - 2\pi(W + W_0)t\right] - \cos\left[2\pi m\alpha W - 2\pi\{(2m-1)W - W_0\}t\right]}{2\pi Wt \sin 2\pi m\alpha W} \\ &+ \frac{\cos\left[(2m-1)\pi\alpha W - 2\pi\{(2m-1)W - W_0\}t\right] - \cos\left[(2m-1)\pi\alpha W - 2\pi W_0 t\right]}{2\pi Wt \sin\left[(2m-1)\pi\alpha W\right]} \end{aligned} \right.$$

$$s_B(t) = s_A(-t) \quad (8)$$

where m is the largest integer for which $(m-1)W < W_0$. Eq. (7b) applies provided that $s_A(t)$ is taken to be the function defined by (8). The separation α between the two interlaced sampling trains is arbitrary, except for the restriction that it may not be an integral multiple of $1/2W$ unless $W_0 = (m-1)W$. In the latter case, a development based on the first-order sampling of (1) and (2) yields the interpolation formula

$$f(t) = \sum_n f\left(\frac{n}{2W}\right) s\left(t - \frac{n}{2W}\right) \quad (9a)$$

$$s(t) = \frac{1}{2\pi Wt}\left[\sin 2\pi mWt - \sin 2\pi(m-1)Wt\right]. \quad (9b)$$

It is interesting to note that the repetitive nature of the spectra $F_A(f)$ and $F_B(f)$ of Fig. 3 offers the possibility of recovering not the original function but a frequency-translated version of it. For example, if the spectral window of Fig. 4 were used, the corresponding time function would represent an upward frequency translation of $f(t)$ by W cps.[17]

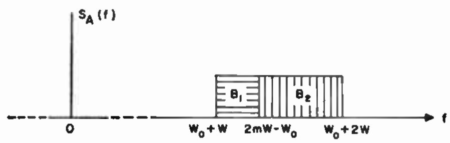
Quadrature and Hilbert Transform Sampling[18]

The sampling operation may be preceded by preparatory processing of the time function. The most obvious example is the representation of a band-pass function in terms of its in-phase and quadrature components, each of which may be sampled separately. Let

$$f(t) = A(t)\cos\left[\omega_0 t + \psi(t)\right] \quad (10)$$

[14] An exceptional case will be discussed at the end of this section.
[15] The only significant difference lies in the fact that the window functions must be computed separately for $W_0 < f < [(2m-1)W - 2W_0]$ and $[(2m-1)W - 2W_0] < f < (W_0 + W)$.
[16] This expression differs from (31) of Kohlenberg, *op. cit.*, only in notation; using $r \equiv 2m-1$, Kohlenberg's result is obtained.

[17] These remarks apply equally well to the low-pass function of Fig. 1. Amplitude modulation could have been achieved by the use of a suitable band-pass spectral window.
[18] Goldman, *op. cit.*



$$B_1 \equiv \frac{\exp\left[-i\left(m\beta-\frac{\pi}{2}\right)\right]}{2W \sin m\beta}$$

$$B_2 \equiv \frac{\exp\left[-i\left(\frac{2m+1}{2}\beta-\frac{\pi}{2}\right)\right]}{2W \sin\left(\frac{2m+1}{2}\beta\right)}$$

Fig. 4—Frequency-translation by use of spectral window.

and let its spectrum $F(f)$ be confined to a frequency band of width W, centered on f_0, as shown in Fig. 5. Providing that $f_0 > W$, the in-phase and quadrature components

$$f_I(t) = A(t) \cos \psi(t) \text{ and } f_Q(t) = A(t) \sin \psi(t) \quad (11)$$

may be obtained by multiplying $f(t)$ by $2 \cos \omega_0 t$ and $-2 \sin \omega_0 t$, respectively, and by filtering out the sum-frequency components. The corresponding spectra are given by

$$F_I(f) = \{F(f) * [\delta(f-f_0) + \delta(f+f_0)]\}_{lf}$$

$$F_Q(f) = \{F(f) * i[\delta(f-f_0) - \delta(f+f_0)]\}_{lf} \quad (12)$$

where the subscript lf indicates that the sum-frequency components have been discarded. These relations are illustrated in Fig. 5. Since $f_I(t)$ and $f_Q(t)$ are band-limited to $-W/2 \leq f \leq W/2$, each may be sampled at the rate of W samples per second. Reconstruction of the original function involves separate interpolations of $f_I(t)$ and $f_Q(t)$, multiplication by $\cos \omega_0 t$ and $\sin \omega_0 t$, respectively, and addition of the results.

First-order sampling of a band-pass function $f(t)$ and of its Hilbert transform

$$f_H(t) = -\frac{1}{\pi}\int_{-\infty}^{+\infty} \frac{f(\tau)d\tau}{t-\tau} = f(t) * \left(-\frac{1}{\pi t}\right) \quad (13)$$

suffices to determine the function. This result is readily obtained by observing that the spectrum $F_H(f)$ of $f_H(t)$ is given by

$$F_H(f) = F(f)\mathfrak{F}\left\{-\frac{1}{\pi t}\right\} = F(f)[-i \operatorname{sgn} f]$$

where $F(f)$ is the spectrum of $f(t)$ and is assumed to be limited to the band $W_0 \leq |f| \leq (W_0 + W)$. The functions $f(t)$ and $f_H(t)$ are now sampled at a rate of W times per second. Using the results of Fig. 3(b), the periodic spectra $\hat{F}(f)$ and $\hat{F}_H(f)$ may be sketched immediately, as

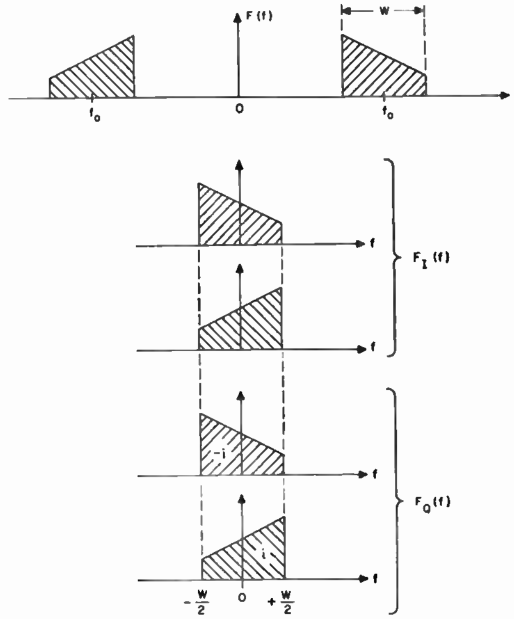


Fig. 5—Quadrature sampling. The direction of cross-hatching distinguishes the positive- and negative-frequency parts of $F(f)$ and the spectral contributions derived from them.

shown in Fig. 6.[19] The required window functions $S(f)$ and $S_H(f)$ may be determined by inspection [Fig. 6(d)], and the corresponding interpolating functions are

$$s(t) = \frac{\sin \pi Wt}{\pi Wt} \cos 2\pi\left(W_0 + \frac{W}{2}\right)t \quad (14a)$$

$$s_H(t) = -\frac{\sin \pi Wt}{\pi Wt} \sin 2\pi\left(W_0 + \frac{W}{2}t\right). \quad (14b)$$

These two functions are Hilbert transforms, as anticipated in the notation. Finally,

$$f(t) = \sum_n f\left(\frac{n}{W}\right) s\left(t - \frac{n}{W}\right) + f_H\left(\frac{n}{W}\right) s_H\left(t - \frac{n}{W}\right). \quad (15)$$

Sampling of Periodic Functions[20]

While the preceding discussion does not exclude line spectra, its results are not particularly useful for periodic functions since the interpolation process is based on an infinite number of samples rather than a finite number of points within one period. The necessary modifications will be outlined for the low-pass case.

Let $f(t)$ be a periodic function of period T, which contains no spectral components above the Nth harmonic, and let the function be sampled at intervals of τ seconds. Fig. 1 applies with $W = N/T$. The inequality $1/\tau > 2W = 2N/T$ cannot be satisfied with the equal sign since this choice would destroy the identity of the spectral line at $f = N/T$. The lowest acceptable rate of equispaced sampling is therefore given by $\tau = T/(2N+1)$. The re-

[19] The similarity between Figs. 5 and 6 is evident. These sketches illustrate the close connection between quadrature sampling and the present procedure.
[20] Goldman, *op. cit.*

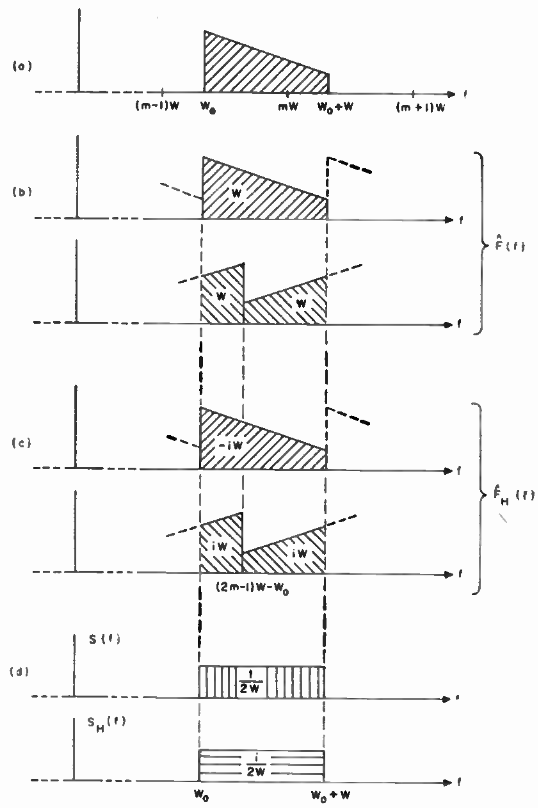


Fig. 6.—Hilbert transform sampling.

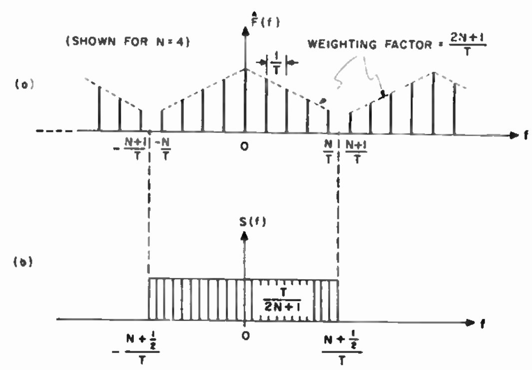


Fig. 7—Sampling of periodic low-pass function.

sulting spectrum is sketched in Fig. 7(a). Since there are gaps between successive replicas of $F(f)$, the spectral window is not uniquely determined. The window function $S(f)$ shown in Fig. 7(b) has the advantage of providing independent sampling since the corresponding interpolating function $s(t)$ has zeros at all sampling points but one. Eq. (3) may now be applied with obvious changes of notation:

$$f(t) = \sum_{n=-\infty}^{+\infty} f(n\tau)s(t - n\tau); \qquad \tau = \frac{T}{2N + 1}, \tag{16a}$$

$$s(t) = \frac{\sin 2\pi\left(\frac{N + \frac{1}{2}}{T}\right)t}{2\pi\left(\frac{N + \frac{1}{2}}{T}\right)t}. \tag{16b}$$

Since $f(t)$ is periodic with period T, (16a) may be written as

$$f(t) = \sum_{n=0}^{2N} f(n\tau)p(t - n\tau)$$

where

$$p(t) = \sum_{k=-\infty}^{+\infty} s(t - kT) = \frac{\sin (2N + 1)\frac{\pi}{T}t}{(2N + 1)\sin\frac{\pi}{T}t}. \tag{17}$$

The last equality is proved in Appendix II.

Sampling of a Function and Its Derivative

Simultaneous sampling of a function and of its derivative yields two periodic spectra from which the original spectrum may be recovered by appropriate spectral windows. It is assumed that the spectrum of $(d/dt)f(t)$ is given by $i2\pi f F(f)$. The procedure will be illustrated for band-limited, low-pass functions. Writing $F_+(f)$ and $F_-(f)$ for the positive and negative frequency parts of $F(f)$, the spectra of $f(t)$ and $f'(t)$ are sketched[21] in Fig. 8(a). The spectra of the sampled functions

$$f_A(t) = f(t)\sum_n \delta\left(t - \frac{n}{W}\right)$$

and

$$f_B(t) = f'(t)\sum_n \delta\left(t - \frac{n}{W}\right) \tag{18}$$

are shown in Fig. 8(b). Using the condition of (6), one obtains in the range $0<f<W$,

$$\begin{aligned} &WF_+(f)S_A(f) + i2\pi fWF_+(f)S_B(f) = F_+(f)\\ &WF_-(f - W)S_A(f)\\ &\qquad + i2\pi(f - W)WF_-(f - W)S_B(f) = 0 \end{aligned} \tag{19}$$

whence

$$\begin{aligned} S_A(f) &= \frac{1}{W}\left(1 - \frac{f}{W}\right), &&0 < f < W\\ S_B(f) &= \frac{1}{i2\pi W^2}, &&0 < f < W. \end{aligned} \tag{20}$$

The interpolating functions are

$$\begin{aligned} s_A(t) &= \left(\frac{\sin \pi Wt}{\pi Wt}\right)^2\\ s_B(t) &= ts_A(t) \end{aligned} \tag{21}$$

so that

$$\begin{aligned} f(t) &= f_A(t) * s_A(t) + f_B(t) * s_B(t)\\ &= \sum_n \left[f\left(\frac{n}{W}\right) + \left(t - \frac{n}{W}\right)f'\left(\frac{n}{W}\right)\right]s_A\left(t - \frac{n}{W}\right). \end{aligned} \tag{22}$$

[21] These sketches are equivalent to (8) of Fogel, *op. cit.*

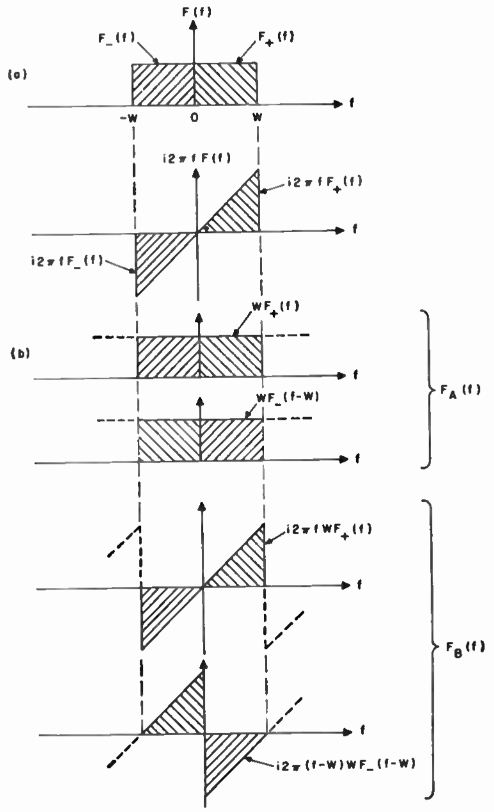


Fig. 8—Sampling of a function and its derivative. The direction of cross-hatching distinguishes the positive- and negative-frequency parts of $F(f)$ and the spectral contributions derived from them.

The more general case of first-order sampling of a real low-pass function and its first R derivatives may be treated by similar methods. The derivation is straightforward, but somewhat lengthy; it is given in Appendix III and leads to the following results. The function and its R derivatives are sampled at intervals of $\tau \equiv (R+1)/2W$ seconds. The spectral window function $S^{(r)}(f)$ for the rth derivative $(r=0, 1, \cdots, R)$ is obtained in $2(R+1)$ segments, each of width $W/(R+1)$, starting at $f=-W$:

$$S^{(r)}(f) = \sum_{n=-(R+1)}^{R} S_n^{(r)}(f).$$

Each segment represents a separate problem; however the following relations reduce the number of functions which must be determined:

$$S_{-(n+1)}^{(r)}(-f) = [S_n^{(r)}(f)]^*$$
$$S_{2m+1}^{(r)}(f) = S_{2m}^{(r)}(f) \qquad (R \text{ odd})$$
$$S_{2m+1}^{(r)}(f) = S_{2m+2}^{(r)}(f) \qquad (R \text{ even}).$$

The $S_n^{(r)}(f)$ for the remaining $(R+1)/2$ (R odd) or $(R+2)/2$ (R even) values of n are found by solving the following sets[22] of equations:

[22] Each set consists of $(R+1)$ equations, corresponding to the $(R+1)$ unknown functions $S_n^{(0)}, \cdots, S_n^{(R)}$.

$$\sum_{r=0}^{R} S_n^{(r)}(f)[i2\pi(f-k/\tau)]^r = \tau\delta_{0,k}, \qquad (f>0)$$
$$k = k_{\min}(n), \cdots, [k_{\min}(n)+R]$$
$$n = 0, 2, 4, \cdots, (R-1) \qquad (R \text{ odd})$$
$$= 0, 2, 4, \cdots, R \qquad (R \text{ even})$$

where $\delta_{0,k}$ is one or zero according as k is zero or nonzero, and where $k_{\min}(n)$ is the smallest integer such that

$$k_{\min}(n) \geq \frac{n-R}{2}.$$

The interpolating functions $s^{(r)}(t)$ are then obtained as the inverse Fourier transforms of the $S^{(r)}(f)$, and

$$f(t) = \sum_{r=0}^{R} \sum_m \frac{d^r f(m\tau)}{dt^r} s^{(r)}(t-m\tau)$$
$$= \sum_m \left[\sum_{r=0}^{R} \frac{d^r f(m\tau)}{dt^r} s^{(r)}(t-m\tau) \right].$$

Appendix I

If the positive and negative-frequency parts of $F(f)$ are designated as $F_+(f)$ and $F_-(f)$, [where $F_+^*(-f) = F_-(f)$], one has in the interval $0<f<W$

$$F_A(f) = WF_+(f) + WF_-(f-W),$$
$$F_B(f) = WF_+(f) + \frac{W}{\gamma} F_-(f-W).$$

Substituting into (6),

$$F_+(f)[WS_A(f) + WS_B(f) - 1]$$
$$+ F_-(f-W)\left[WS_A(f) + \frac{W}{\gamma} S_B(f)\right] = 0.$$

There is, in general, no functional relationship between $F_+(f)$ and $F_-(f-W) = F_+^*(W-f)$; equating to zero the coefficients of $F_+(f)$ and $F_-(f-W)$, one obtains the two equations following (6).

Appendix II

$$p(t) = \sum_{k=-\infty}^{\infty} s(t-kT) = s(t) * \left[\sum_{k=-\infty}^{\infty} \delta(t-kT)\right].$$

The corresponding spectrum is

$$P(f) = S(f) \sum_{k=-\infty}^{\infty} \frac{1}{T}\delta\left(f-\frac{k}{T}\right)$$
$$= \frac{1}{2N+1} \sum_{k=-N}^{N} \delta\left(f-\frac{k}{T}\right).$$

The last equality may be verified by inspection of the window function $S(f)$ shown in Fig. 7(b). Finally,

$$p(t) = \frac{1}{2N+1}\sum_{k=-N}^{N}(e^{i(2\pi/T)})^k = \frac{\sin(2N+1)\dfrac{\pi t}{T}}{(2N+1)\sin\dfrac{\pi t}{T}}.$$

Appendix III

It will be assumed that the spectrum of the rth derivative is $(i2\pi f)^r F(f)$. The function and its first R derivatives are sampled at intervals of $\tau \equiv (R+1)/2W$ seconds. Their spectra are therefore convolved with the impulse function train

$$\frac{1}{\tau}\sum_k \delta\left(f - \frac{k}{\tau}\right). \tag{23}$$

Each of the $(R+1)$ spectra extends from $-W$ to W, and will be divided into $2(R+1)$ intervals of width $W/R+1 = 1/2\tau$, starting at $f = -W$. Let $F_n(f)$ be equal to $F(f)$ in the nth interval and zero outside it, *i.e.*,

$$F(f) = \sum_{n=-(R+1)}^{R} F_n(f). \tag{24}$$

Since $f(t)$ is assumed to be real, $F_{-(n+1)}(-f) = F_n^*(f)$. Fig. 9 shows the spectrum $F(f)$, the numbering of its $(R+1)$ intervals, and the convolving train of impulse functions.

The convolution process is visualized in terms of erecting replicas centered on the impulse functions. It is easily seen that a replica of $F(f)$, centered on the impulse function at $f = k/\tau$, will contribute to the $(2k+j)$th interval the function

$$\frac{1}{\tau}F_j\left(f - \frac{k}{\tau}\right).$$

Using the notation

$$D^k[g(f)] \equiv g\left(f - \frac{k}{\tau}\right),$$

a replica of $F(f)$ centered on $\delta(f - k/\tau)$ will contribute to the nth interval the function $1/\tau D^k[F_{n-2k}(f)]$. Let $F^{(r)}(f)$ be the spectrum obtained from the convolution of $(i2\pi f)^r F(f)$ with the train on impulse functions of (23), and let $F_n^{(r)}(f)$ be its nth segment, *i.e.*,

$$F^{(r)}(f) = \sum_{n=-(R+1)}^{R} F_n^{(r)}(f).$$

Then it follows from the preceding discussion that[23]

$$F_n^{(r)}(f) = \frac{1}{\tau}\sum_{k=k_{\min}(n)}^{k_{\min}(n)+R} D^k[(i2\pi f)^r F_{n-2k}(f)]. \tag{25}$$

Since $F_{n-2k}(f)$ vanishes outside the interval $(-W, W)$,

[23] Eq. (25) is equivalent to (14) of Fogel, *op. cit.*

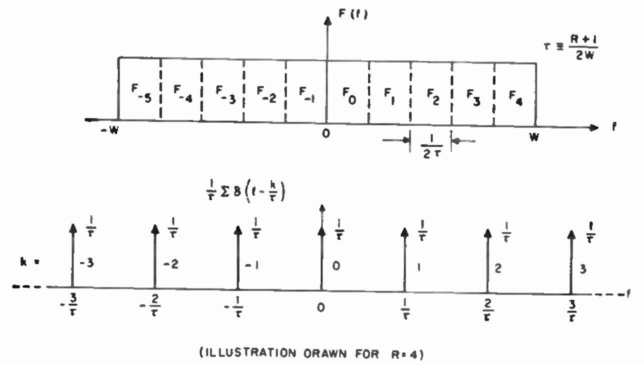


Fig. 9—Sampling of a function and its first R derivatives.

the summation may be restricted to those integral values of k which satisfy the inequality

$$-(R+1) \leq (n - 2k) \leq R.$$

For each n, there are therefore $(R+1)$ values of k, starting with $k_{\min}(n)$; the latter is the smallest integer which satisfies

$$k_{\min}(n) \geq \frac{n-R}{2}. \tag{26}$$

In order to recover $F(f)$ from the $(R+1)$ spectra $F^{(r)}(f)$, each $F^{(r)}(f)$ is multiplied by a spectral window function $S^{(r)}(f)$. One then demands that

$$\sum_{r=0}^{R} S^{(r)}(f)F^{(r)}(f) = F(f). \tag{27}$$

Since $f(t)$ was assumed to be real, the $S^{(r)}(f)$ are spectra of real functions, and it suffices to consider positive frequencies only. Each of the $(R+1)$ positive-frequency intervals must be considered separately so that (27) represents $(R+1)$ separate equations;

$$\sum_{r=0}^{R} S_n^{(r)}(f)F_n^{(r)}(f) = F_n(f); \quad n = 0, 1, \cdots R \tag{28}$$

where $S_n^{(r)}(f)$ represents the nth segment of $S^{(r)}(f)$, with

$$S_{-(n+1)}^{(r)}(-f) = [S_n^{(r)}(f)]^*. \tag{29}$$

Substituting (25) into (28),

$$\sum_{r=0}^{R} S_n^{(r)}(f)\frac{1}{\tau}\sum_{k=k_{\min}(n)}^{k_{\min}(n)+R} D^k[(2\pi f i)^r F_{n-2k}(f)] = F_n(f)$$

$$n = 0, 1, \cdots, R. \tag{30}$$

Interchanging orders of summation,

$$\sum_{k=k_{\min}(n)}^{k_{\min}(n)+R} D^k[F_{n-2k}(f)]\sum_{r=0}^{R} S_n^{(r)}(f)D^k[(2\pi f i)^r]$$

$$= \tau F_n(f). \tag{31}$$

Since the $F_n(f)$ are independent, the coefficient of each $D^k[F_{n-2k}]$ must be identically zero. For each value of n, (31) thus provides $(R+1)$ equations

$$\sum_{r=0}^{R} S_n^{(r)}(f)D^k[(2\pi fi)^r] = \tau\delta_{0,k}$$
$$k = k_{\min}(n), \cdots, [k_{\min}(n) + R]$$
$$n = 0, 2, \cdots, R \tag{32}$$

where $\delta_{0,k}$ is one or zero according as k is zero or nonzero.

Inspection of (26) shows that for odd R,

$$k_{\min}(0) = k_{\min}(1), \qquad k_{\min}(2) = k_{\min}(3), \text{ etc.},$$

while for even R,

$$k_{\min}(1) = k_{\min}(2), \qquad k_{\min}(3) = k_{\min}(4), \text{ etc.}$$

Thus

$$S_{2m+1}^{(r)} = S_{2m}^{(r)} \qquad (R \text{ odd})$$
$$S_{2m+1}^{(r)} = S_{2m+2}^{(r)} \qquad (R \text{ even}).$$

It is therefore sufficient to solve (32) for even values of n so that there are $(R+1)/2$ or $(R+2)/2$ sets of equations, according to whether R is odd or even.

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An Application of Piecewise Approximations to Reliability and Statistical Design*

HARRY J. GRAY, JR.†, MEMBER, IRE

Summary—**If a random variable can be expressed as a weighted sum of other random variables having known distributions which can be approximated piecewise by, for example, polynomials, the distribution of the random variable can be obtained, relatively easily, by the use of the algorithm described in this paper.**

INTRODUCTION

In many systems, such as missile, computer, or control systems, there may arise a need for the determination of the probability of failure due to the gradual deterioration of the system components. Associated with this need is the determination of the probability that a specified characteristic of the system or a part of the system will be outside of acceptable limits on account of a chance unfavorable combination of component values. Examples of specific characteristics might be: the delay of a pulse circuit, the phase margin in a feedback control system, the gain of a linear amplifier—quantities all of which are functions of the values of the components involved such as resistances, capacitances, vacuum tube transconductances, and plate resistances, etc. Denote the characteristic by T and the values of the components involved by $x_1, x_2, \cdots, x_n$.

Then

$$T = T(x_1, x_2, \cdots, x_n). \tag{1}$$

It is often possible to express sufficiently accurately the deviation δT of the characteristic T from some nominal value in terms of the deviations of the component values, δx_i, from their mean values as follows:

$$\delta T = a_1\delta x_1 + a_2\delta x_2 + \cdots + a_n\delta x_n. \tag{2}$$

The numbers, $a_1, a_2, \cdots, a_n$ can be determined either by experiment or by calculation. Eq. (2) may be rewritten:

$$\delta T/T_0 = b_1\delta x_1/x_{10} + b_2\delta x_2/x_{20} + \cdots + b_n\delta x_n/x_{n0};$$
$$b_i = a_i x_{i0}/T_0 \qquad i = 1, 2, \cdots n \tag{3}$$

where $T_0, x_{10}, x_{20}, \cdots, x_{n0}$ are the "mean" values of $T, x_1, \cdots, x_n$. $[T_0 \approx T(x_{10}, x_{20}, \cdots, x_{n0})]$. Eq. (3) can be considered as expressing the percentage change in the characteristic resulting from certain percentage changes in the components involved, as the equality is not affected by multiplying both sides by 100. The problem then becomes one of determining how ξ is distributed knowing how the ξ_i are distributed where

$$\xi = \xi_1 + \xi_2 + \cdots + \xi_n \tag{4}$$

and $\xi = \delta T/T_0$, $\xi_i = b_i\delta x_i/x_{i0}$; $i = 1, 2, \cdots, n$, the mean of ξ_i is zero for $i = 1, 2, \cdots, n$, and the mean of ξ is zero. The ξ_i are assumed to be independent random variables.[1]

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† Moore School of Electrical Engineering, Philadelphia 4, Pa.

[1] The assumption that the means of ξ and ξ_i are zero is not necessary, but simplifies the discussion that follows.

One way of solving the problem has been to assume that the ξ_i are normally distributed.[2,3] It follows, then, that ξ is normally distributed, and the required probabilities may be calculated using tables for the normal curve and the easily calculated standard deviation of ξ.

Another method that has been used[4] when the ξ_i are not normally distributed is to rely on the central limit theorem and to calculate the required probabilities using tables for the normal curve and the standard deviation, σ, of ξ given by $\sigma^2=\sigma_1^2+\sigma_2^2+\cdots+\sigma_n^2$ where the σ_i $(i=1, 2, \cdots, n)$ are the standard deviations of the ξ_i. This procedure is good when the number of the random variables, ξ_i, is large. However, it is difficult to determine how trustworthy the results are. Recourse to Gram-Charlier or Edgeworth's series[5] often gives ridiculous results such as negative values for the calculated probabilities. Recourse to conservative estimates such as Tchebycheff's inequality[5] is often useless as such estimates are often too conservative.

It would be good if a method could be found which would make it possible to compute the distribution of a sum of random variables, given their individual distributions, to any desired degree of precision. It also would be good if the method for doing this was routine or algorithmic in nature so that no special knowledge would be required for its application. An algorithmic method has the advantage that it can be programmed for a digital computer.

It is the purpose of this paper to present such a method. The method makes use of piecewise polynomial approximations. Its application will be illustrated by application to a diode circuit problem. The procedure is justified in the Appendix.

Application of the Algorithmic Method

Consider the diode circuit in Fig. 1. The following will be assumed: $C_0=25$ $\mu\mu$, $C_2=12$ $\mu\mu$, and $C_3=12$ $\mu\mu$, and are assumed to be fixed.

It is also assumed that: 1) all diodes are ideal, 2) all inductances are negligible, and 3) supply voltages having same values come from one common point.

$$\left.\begin{aligned} R_1 &= 46.3k,\\ R_2 &= 18.5k,\\ R_3 &= 10.9k,\\ V_1 &= +70,\\ V_2 &= -70, \text{ and}\\ V_3 &= -5. \end{aligned}\right\} \text{These are nominal values with a tolerance of } \pm 5 \text{ per cent}$$

[2] H. T. Marcy and M. Yachter, "Steady-state systems engineering in automatic process control" in "Conference on Automatic Control," Arnold Tustin, ed., Butterworths Scientific Publications, London, Eng.; 1952.

[3] A. H. Benner and B. Meredith, "Designing reliability into electronic circuits," *Proc. Natl. Electronics Conf.*, vol. 10; pp. 137–145; October, 1954.

[4] C. N. Weygandt, private communication.

[5] H. Cramér, "Mathematical Methods of Statistics," Princeton University Press, Princeton, N. J.; 1946.

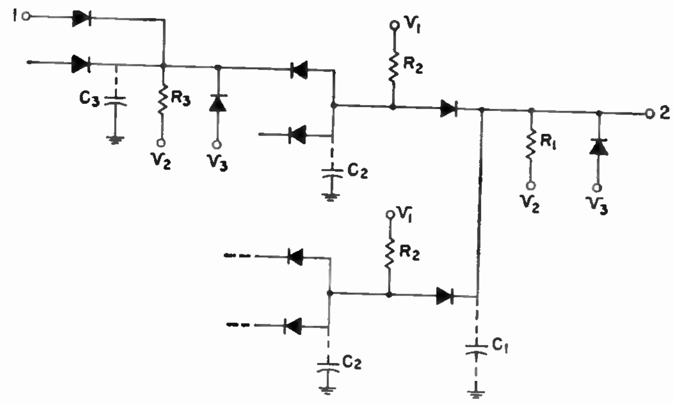

Fig. 1—Diode gating circuit.

Consider T_{r2}, the rise time of point 2, from V_3 to 0 v, assuming a step input at point 1. The nominal (mean) value of T_{r2} is found to be 75 mμsec. The largest possible value of T_{r2} is found to be 100 mμsec.

It is desired to find the probability that δT_{r2} will exceed, say, 20 mμsec. It is found that the equation

$$\delta T_{r2}/T_{r20} = 2.015\delta V_1/V_{10} + 0.814\delta V_2/V_{20} + 0.92\delta V_3/V_{30} + 0.785\delta R_1/R_{10} + 2.11\delta R_2/R_{20}$$

weights the contributions of the various voltages and resistors in the proper ratio and conservatively describes the characteristics of the circuit as far as $\delta T_{r2}/T_{r20}$ is concerned for $0 \le \delta T_{r2} \le 25$ mμsec, where $\delta T_{r2}=T_{r2}-T_{r20}$.

Let

$$\xi_1 = 2.015\delta V_1/V_{10}, \qquad \xi_2 = 0.814\delta V_2/V_{20}, \text{ etc.,}$$

and

$$\xi = \delta T_{r2}/T_{r20}.$$

Then

$$\xi = \xi_1 + \xi_2 + \xi_3 + \xi_4 + \xi_5.$$

Assume $\xi_1, \xi_2, \cdots$, etc., are uniformly distributed, *i.e.*, the frequency function of ξ_i is $f_i(x)$ where

$$f_i(x) = 0: \qquad x < -a_i,\ x > a_i$$

$$f_i(x) = \frac{1}{2a_i}; \qquad -a_i < x < a_i.$$

Then

$$\begin{aligned} a_1 &= 10.15 \text{ per cent}\\ a_2 &= 4.06 \text{ per cent}\\ a_3 &= 4.6 \text{ per cent}\\ a_4 &= 3.92 \text{ per cent}\\ a_5 &= \underline{10.55} \text{ per cent}\\ & 33.3 \text{ per cent (limit of } \xi) \end{aligned}$$

A typical frequency function $f(x)$ is plotted in Fig. 2. If it is assumed that the frequency function is to be approximated from left to right, n points $(x_1, x_2, \cdots, x_n)$ called critical points are selected, and in the interval between adjacent points, the frequency function is approximated by any convenient polynomial. If $f(x)$ is the frequency function of a component, it is convenient to assume wherever possible the simplest polynomial—the equation of a straight line. If the frequency function is the frequency function to be solved for, the critical points will be determined by the algorithmic method as will be the polynomials approximating the final frequency function between the critical points.

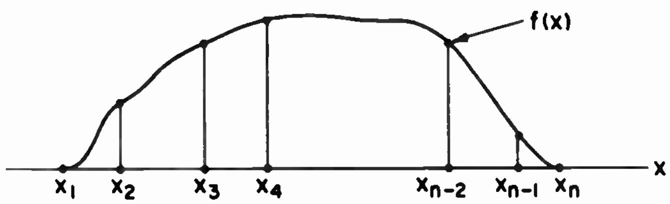

Fig. 2—Frequency function approximated from left to right.

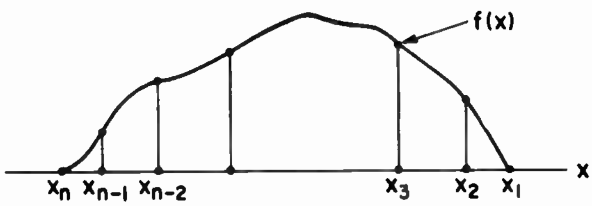

Fig. 3—Frequency function approximated from right to left.

In the example given, it is desired to determine the nature of the frequency function for ξ for values of ξ which are in excess of the mean value and, therefore, lie on the right-hand tail of the frequency function. The algorithmic method will give the complete piecewise polynomial representation of the frequency function for ξ. However, since the right-hand tail of the curve is of interest in the example, the amount of work can be considerably reduced if the frequency functions involved are approximated from *right* to *left* as in Fig. 3.

The function $f_i(x)$ looks like a pulse of width $2a_i$ and amplitude $1/2a_i$. It can be approximated by a single straight-line segment in the interval $(-a_i, a_i)$. The algorithmic method requires that in approximating from right to left, the function be represented by the sum of *semi-infinite* polynomials starting at the critical points and extending indefinitely to the left. In the notation discussed in the Appendix,

$$f_i(x) = (1/2a_i)[g(x, a_i, 0) - g(x, -a_i, 0)] \tag{5}$$

where $g(x, a_i, 0)$, is the zero-order polynomial (a constant) starting at $x = a_i$ having an amplitude of unity. In general

$$g(x, x_0, k) = \frac{(x_0 - x)^k}{k!}; \qquad x < x_0$$
$$g(x, x_0, k) = 0; \qquad x > x_0. \tag{6}$$

Corresponding to $g(x, x_0, k)$, there is a *characteristic function* [(22) Appendix]

$$e^{jtx_0}/(jt)^{k+1} \tag{7}$$

which may replace it in (5) yielding the characteristic function

$$\phi_i(t) = (1/2a_i)\left[\frac{e^{jta_i}}{jt} - \frac{e^{jt(-a_i)}}{jt}\right]. \tag{8}$$

This procedure is justified in the Appendix. The characteristic function of the distribution of ξ can be shown to be[5]

$$\phi(t) = \phi_1(t)\phi_2(t) \cdots \phi_5(t)$$

which is from (8),

$$\phi(t) = \frac{1}{2^5 a_1 a_2 a_3 a_4 a_5}\left[e^{jta_1} - e^{-jta_1}\right]\left[e^{jta_2} - e^{-jta_2}\right]\cdot \cdots \left[e^{jta_5} - e^{-jta_5}\right]\frac{1}{(jt)^5}. \tag{9}$$

Eq. (9) may be multiplied out to yield an expression having the form

$$\phi(t) = \frac{1}{2^5 a_1 a_2 a_3 a_4 a_5}\left[\frac{e^{jtx_1}}{(jt)^5} + c\,\frac{e^{jtx_2}}{(jt)^5} + \cdots\right]. \tag{10}$$

The terms of the form of (7) may be replaced in (10) by the appropriate g functions given by (6) to give the frequency function for ξ. If the reader carries this out, it will be clear to him that the frequency function consists of 31 fourth-degree curves fitted together at the critical points $x = \pm a_1 \pm a_2 \pm a_3 \pm a_4 \pm a_5$ ($2^5 = 32$ points).

Since interest is in the righthand tail of $f(x)$, the larger critical points must be determined. Three of these are (approximating from right to left)

$$x_1 = a_1 + a_2 + a_3 + a_4 + a_5 = 33 \text{ per cent } (\delta T_{r2} = 25 \text{ m}\mu\text{s})$$
$$x_2 = 25.5 \text{ per cent } (\delta T_{r2} = 19.1 \text{ m}\mu\text{s})$$
$$x_3 = 25.2 \text{ per cent } (\delta T_{r2} = 18.9 \text{ m}\mu\text{s}).$$

From (10) and the values of the a_i, one obtains

$$\delta(t) = \frac{3.98 \times 10^{-6}}{(jt)^5}\left(e^{jtx_1} - e^{jtx_2} + \cdots\right),$$

and the carrying out of the substitution of g functions yields

$$f(x) = 0; \quad x > 33.3 \text{ per cent}$$
$$f(x) = \frac{3.98 \times 10^{-6}}{4!}(33.3 - x)^4; \ 25.5 \text{ per cent} < x < 33.3 \text{ per cent}$$
$$f(x) = \frac{3.98 \times 10^{-6}}{4!}\left[(33.3 - x)^4 - (25.5 - x)^4\right] 25.2 \text{ per cent} < x < 25.5 \text{ per cent (etc.)} \tag{11}$$

To calculate the probability that $\delta t_{r2} > 20$ mμs (26.7 per cent), we have from (11)

$$P(x > 26.7 \text{ per cent}) = \int_{26.7}^{33.3} f(x)dx = 0.405 \times 10^{-3}. \quad (12)$$

It is easily found that the standard deviation of x is $\sigma = 9.45$. Using the normal error curve, one obtains

$$P(x > 26.7 \text{ per cent}) = P(x > 2.83\sigma) = 2.34 \times 10^{-3}$$

which is more than five times higher than the value obtained in (12).

In the preceding example, certain simplifications were present such as the fact that the random variables had zero means and were uniformly distributed. Such assumptions are not necessary, in general. It appears that the probability density functions need be such that they may be approximated by pieces of curves corresponding to simple functions, these functions being polynomials in the example.

Appendix

Heuristic Approach

It will be assumed that frequency functions for $\xi, \xi_1, \xi_2, \cdots, \xi_n$ exist and are continuous except possibly at a finite number of points in a finite interval. Then, if the ξ_i are independent random variables, the frequency function of ξ can be obtained by repeated evaluation of the convolution integral[4]: *i.e.*, if $f(x)$, $f_1(x)$, and $f_2(x)$ are the frequency functions of $\xi+\eta$, ξ, and η respectively, and ξ and η are independent random variables, then

$$f(x) = \int_{-\infty}^{\infty} f_1(x-z)f_2(z)dz = \int_{-\infty}^{\infty} f_2(x-z)f_1(z)dz. \quad (13)$$

Thus if three independent random variables have each the rectangular frequency function,[5]

$$\begin{aligned} f(x) &= 0; \qquad x < -\tfrac{1}{2},\ x > \tfrac{1}{2} \\ f(x) &= 1; \qquad -\tfrac{1}{2} < x < \tfrac{1}{2}, \end{aligned} \quad (14)$$

then the sum of two of these random variables has the frequency function (triangular) obtained by the use of (13):[6]

$$\begin{aligned} f(x) &= 0; \qquad x < -1,\ x > 1 \\ f(x) &= x + 1; \qquad -1 < x < 0 \\ f(x) &= 1 - x; \qquad 0 < x < 1. \end{aligned} \quad (15)$$

Application of (13) to (14) and (15) yields the frequency function for the sum of the three random variables (three parabolas):[5]

$$\begin{aligned} f(x) &= 0; \qquad x < -\tfrac{3}{2},\ x > \tfrac{3}{2} \\ f(x) &= \tfrac{1}{2}(x+\tfrac{3}{2})^2; \qquad -\tfrac{3}{2} < x < -\tfrac{1}{2} \\ f(x) &= \tfrac{1}{2}[(x+\tfrac{3}{2})^2 - 3(x+\tfrac{1}{2})^2]; \qquad -\tfrac{1}{2} < x < \tfrac{1}{2} \\ f(x) &= \tfrac{1}{2}[(x+\tfrac{3}{2})^2 - 3(x+\tfrac{1}{2})^2 + 3(x-\tfrac{1}{2})^2]; \ \tfrac{1}{2} < x < \tfrac{3}{2}. \end{aligned} \quad (16)$$

This is a direct procedure and could be applied to (4) except that the labor involved in the evaluation of the convolution integrals is excessive. An indirect procedure involves the use of characteristic functions.[5] If the frequency function of ξ_i is $f_i(x)$, its characteristic function is

$$\phi_i(t) = \int_{-\infty}^{\infty} f_i(x)e^{jtx}dx; \qquad j = \sqrt{-1}. \quad (17)$$

If the ξ_i are independent, the characteristic function of ξ, the sum of the ξ_i is $\phi(t)$ where

$$\phi(t) = \phi_1(t)\phi_2(t) \cdots \phi_n(t), \quad (18)$$

and the frequency function of ξ is given by

$$f(x) = \frac{1}{2\pi}\int_{-\infty}^{\infty} \phi(t)e^{-jtx}dt. \quad (19)$$

However, the evaluation of (19) involves contour integration in the complex plane and is also quite laborious. Nevertheless, the form of (17)–(19) suggests the existence of an operational method.

Consider the function

$$\begin{aligned} f(x, x_0, k) &= 0; \qquad x < x_0 \\ f(x, x_0, k) &= \frac{(x-x_0)^k}{k!}; \qquad x > x_0. \end{aligned} \quad (20)$$

The frequency function for the rectangular distribution is then, from (14),

$$f(x, -\tfrac{1}{2}, 0) - f(x, \tfrac{1}{2}, 0).$$

Eq. (15) may be written

$$f(x, -1, 1) - 2f(x, 0, 1) + f(x, 1, 1),$$

and (16) may be written

$$f(x, -\tfrac{3}{2}, 2) - 3f(x, -\tfrac{1}{2}, 2) + 3f(x, \tfrac{1}{2}, 2) - f(x, \tfrac{3}{2}, 2)$$

The characteristic function for the frequency function given by (14) is using (17)

$$-\frac{e^{jt(-1/2)}}{jt} + \frac{e^{jt(1/2)}}{jt},$$

and for (15), it is either by (17) or by (18),

$$\frac{e^{jt(-1)}}{(jt)^2} - 2\frac{e^{jt(0)}}{(jt)^2} + \frac{e^{jt(1)}}{(jt)^2},$$

and for (16)

$$-\frac{e^{jt(-3/2)}}{(jt)^3} + 3\frac{e^{jt(-1/2)}}{(jt)^3} - 3\frac{e^{jt(1/2)}}{(jt)^3} + \frac{e^{jt(3/2)}}{(jt)^3}.$$

Comparison of the above pairs of frequency functions and characteristic functions suggests the transform pair,

$$f(x, x_0, k) \leftrightarrow (-1)^{k+1}\frac{e^{jtx_0}}{(jt)^{k+1}}. \quad (21)$$

Note, however, that substitution of (20) into (17) yields a divergent integral. A rigorous proof of (21) will be given later.

If, on the other hand, one defines

$$g(x, x_0, k) = \frac{(x_0 - x)^k}{k!}; \quad x < x_0$$

$$g(x, x_0, k) = 0; \quad x > x_0.$$

Then

$$g(x, x_0, k) \leftrightarrow \frac{e^{jtx_0}}{(jt)^{k+1}}. \tag{22}$$

Eqs. (21) and (22) form the basis of the operational method.

Rigorous Approach

Assume that the frequency function in Fig. 2 is piecewise approximated with critical points $x_1 < x_2 < \cdots < x_n$. Let

$$f(x) = P_1(x - x_1)u(x - x_1) + P_2(x - x_2)u(x - x_2) + \cdots + P_n(x + x_n)u(x - x_n) \tag{23}$$

where

$u(x - x_k) = 0;\ x \leq x_k$ and $f(x) = 0;\ x < x_1$ and $x > x_n$

$u(x - x_k) = 1;\ x > x_k.$

The characteristic function is

$$\phi(t) = \int_{-\infty}^{\infty} f(x)e^{jtx}dx = \int_{x_1}^{x_n} f(x)e^{jtx}dx$$

or

$$\phi(t) = \int_{x_1}^{x_n} P_1(x - x_1)u(x - x_1)e^{jtx}dx + \int_{x_2}^{x_n} P_2(x - x_2)u(x - x_2)e^{jtx}dx + \cdots + \int_{x_{n-1}}^{x_n} P_{n-1}(x - x_{n-1})u(x - x_{n-1})e^{jtx}dx$$

or

$$\phi(t) = \int_{x_1}^{x_n} P_1(x - x_1)e^{jtx}dx + \int_{x_2}^{x_n} P_2(x - x_2)e^{jtx}dx + \cdots + \int_{x_{n-1}}^{x_n} P_{n-1}(x - x_{n-1})e^{jtx}dx.$$

Denote

$$\int^{x_s} P_k(x - x_k)e^{jtx}dx = I_k(x_s). \tag{24}$$

Then

$$\phi(t) = I_1(x_n) - I_1(x_1) + I_2(x_n) - I_2(x_2) + \cdots + I_{n-1}(x_n) - I_{n-1}(x_{n-1}). \tag{25}$$

Let

$$I_1(x_n) + I_2(x_n) + \cdots + I_{n-1}(x_n) = K(x_n).$$

Then

$$\phi(t) = -I_1(x_1) - I_2(x_2) - \cdots - I_{n-1}(x_{n-1}) + K(x_n). \tag{26}$$

Comparison of (23) and (25) shows the following correspondences:

$$P_1(x - x_1)u(x - x_1) \leftrightarrow -I_1(x_1)$$
$$P_2(x - x_2)u(x - x_2) \leftrightarrow -I_2(x_2)$$
$$\cdots\cdots\cdots\cdots$$
$$P_{n-1}(x - x_{n-1})u(x - x_{n-1}) \leftrightarrow -I_{n-1}(x_{n-1})$$
$$P_n(x - x_n)u(x - x_n) \leftrightarrow K(x_n).$$

According to (20), let

$$P_0(x - x_0)u(x - x_0) = c_0 \frac{(x - x_0)^k}{k!} u(x - x_0) = c_0 f(x, x_0, k).$$

Then

$$c_0 f(x, x_0, k) \leftrightarrow -I_0(x_0)$$

where

$$-I_0(x_0) = -\int^{x_0} P_0(x - x_0)e^{jtx}dx = -c_0 \int^{x_0} \frac{(x - x_0)^k}{k!} e^{jtx}dx. \tag{27}$$

Evaluation of (27) yields

$$-I_0(x_0) = c_0 \frac{(-1)^{k+1}}{(jt)^{k+1}} e^{jtx_0}.$$

Hence,

$$f(x, x_0, k) \leftrightarrow \frac{(-1)^{k+1}}{(jt)^{k+1}} e^{jtx_0}$$

may be regarded as a transform pair [see (21)].

If the frequency function is approximated as in Fig. 3, where $x_1 > x_2 > \cdots > x_n$ and

$$f(x) = \hat{u}(x_1 - x)\hat{P}_1(x_1 - x) + \cdots + \hat{u}(x_n - x)\hat{P}_n(x_n - x)$$

where $\hat{u}(x_0 - x) = 1 - u(x - x_0)$, then it is easy to show that

$$\hat{u}(x_0 - x)\hat{P}_0(x_0 - x) \leftrightarrow \int^{x_0} \hat{P}_0(x_0 - x)e^{jtx}dx.$$

If one sets

$$\hat{u}(x_0 - x)\hat{P}_0(x_0 - x) = \hat{c}_0 \frac{(x_0 - x)^k}{k!} \hat{u}(x_0 - x) = \hat{c}_0 g(x, x_0, k),$$

then it follows that

$$g(x, x_0, k) \leftrightarrow \frac{e^{jtx_0}}{(jt)^{k+1}} \tag{28}$$

may be regarded as a transform pair [see (22)].

An interesting consequence of (28) is the following pair of relations:

$$\phi(t) = \sum_{i=1}^{n-1} \hat{c}_i \frac{e^{jtx_i}}{(jt)^{\alpha_i}}$$

if

$$f(x) = \sum_{i=1}^{n-1} \hat{c}_i \frac{(x_i - x)^{\alpha_i - 1}}{(\alpha_i - 1)!} \hat{u}(x_i - x); \qquad x > x_n. \quad (29)$$

The second of these relations may be recognized as a general form for a frequency function piecewise approximated from right to left. The first of these relations is recognized as the result of replacement of terms using (28).

The probability that $x > \epsilon$ may be obtained as follows:

$$P(x > \epsilon) = \int_{\epsilon}^{\infty} f(x) dx.$$

Substitution of (29) in the above yields, after integration,

$$P(x > \epsilon) = \hat{c}_i \frac{(x_i - \epsilon)^{\alpha_i}}{(\alpha_i)!} \hat{u}(x_i - x). \quad (30)$$

In particular, if ϵ is a critical point x_s, (30) reduces to

$$P(x > x_s) = \sum_{i=1}^{s-1} \hat{c}_i \frac{(x_i - x_s)^{\alpha_i}}{(\alpha_i)!}.$$

An Instantaneous Microwave Polarimeter*

P. J. ALLEN†, SENIOR MEMBER, IRE, AND R. D. TOMPKINS†

Summary—**Used as a precision dual-balanced mixer with circular waveguide input, the trimode turnstile waveguide junction is the key to a simple microwave polarimeter technique which permits instantaneous viewing of input polarization. Through linear mixing, the relative phase and amplitude of orthogonal components of an arbitrarily polarized input signal are preserved in the IF outputs of the two mixers. After amplification, these two IF signals are applied to orthogonal deflection planes of a cathode-ray tube to obtain an accurate, instantaneous "picture" of input polarization. Circular polarization generates a circle; elliptical polarization, an ellipse; and linear polarization, a line which indicates the plane of polarization. In certain applications, a variation of the method permits direct IF recording of the polarization information. Instrument errors and ways to minimize these are discussed. The polarimeter technique has application in microwave communication, radar, countermeasures, radio astronomy, antenna studies, and in laboratory measurements of polarization.**

Introduction

A KNOWLEDGE of the polarization characteristics of an electromagnetic wave is essential in many fields concerned with guided or free space propagation of radio energy. In investigations such as the radar study of target-echo characteristics for example, detailed knowledge of the polarization characteristics of the signal return is required. At microwave frequencies, the simplest and most common method of measuring polarization has been to employ a rotating linearly-polarized analyzer which samples the signal amplitude at various polarizations. However, this method is lacking in two important respects: relative phase information is not obtained, and amplitude information is collected on a time-sequential basis. For many applications, such as the one cited above, complete *instantaneous* indication of polarization is needed.

An improved method of polarization measurement simultaneously compares the *amplitudes* of two orthogonal linearly- or circularly-polarized components, but unless the relative phase between these components also is obtained, the polarization information is incomplete. The microwave circuitry required to obtain instantaneous phase information, as well as the amplitude data, generally has been quite complex, usually narrow band, and of only moderate accuracy. The new microwave polarimeter technique to be described utilizes both phase and amplitude information to provide an accurate instantaneous presentation of the input signal polarization characteristics, and overcomes the broadband RF circuitry problem with very simple and compact microwave plumbing.

Polarization Measurements

There are a number of ways of completely specifying the polarization characteristics of an electromagnetic wave.[1-3] For example, an arbitrarily polarized plane wave can be described in terms of orthogonal x and y

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† U. S. Naval Res. Lab., Washington, D. C.

[1] J. D. Kraus, "Antennas," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 464–485; 1950.

[2] V. H. Rumsey, *et al.*, "Techniques for handling elliptically polarized waves with special reference to antennas," Proc. IRE, vol. 39, pp. 533–552; May, 1951.

[3] M. H. Cohen, "Radio astronomy polarization measurements," Proc. IRE, vol. 46, pp. 172–183; January, 1958.

components which are located in a reference plane which is normal to the direction of propagation.[2] Thus,

$$E_x = A\cos(\omega t + \phi_1)$$
$$E_y = B\cos(\omega t + \phi_2).$$

Hence the measurement of polarization amounts to establishing the parameters of the above equations either by direct measurement or by computation from other measurements such as axial ratio, orientation angle, etc.

Of the techniques generally used for measuring polarization,[1–6] most involve time-sequential measurement of some parameters. This is particularly true at microwave frequencies where sequential techniques are relatively simple, but where instantaneous (or simultaneous) techniques have required rather complex plumbing, consequently limiting bandwidth and accuracy. Using very simple yet broadband microwave circuitry, the trimode turnstile polarimeter provides accurate and instantaneous presentation of the polarization characteristics of the input signal.

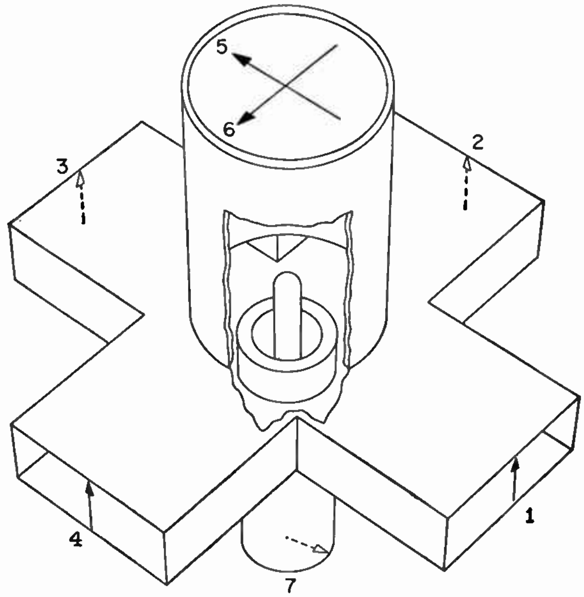

Fig. 1—The trimode turnstile waveguide junction, a unique 7-port hybrid junction.

TABLE I

THEORETICAL COUPLING CHARACTERISTICS OF A TRIMODE TURNSTILE JUNCTION HAVING MATCHED COAXIAL AND CIRCULAR WAVEGUIDE PORTS

Unit Power Input to Port Number	Relative Power Output from Port Number						
	1	2	3	4	5	6	7
7	1/4 (−6 db)	1/4 (−6 db)	1/4 (−6 db)	1/4 (−6 db)	0	0	0 (no reflec)
5	1/2 (−3 db)	0	1/2 (−3 db)	0	0 (no reflec)	0	0
6	0	1/2 (−3 db)	0	1/2 (−3 db)	0	0 (no reflec)	0
1 or 3	1/16 (−12 db)	1/16 (−12 db)	1/16 (−12 db)	1/16 (−12 db)	1/2 (−3 db)	0	1/4 (−6 db)
2 or 4	1/16 (−12 db)	1/16 (−12 db)	1/16 (−12 db)	1/16 (−12 db)	0	1/2 (−3 db)	1/4 (−6 db)

THE TRIMODE TURNSTILE JUNCTION

Before discussing the principle of the polarimeter, the pertinent properties of the trimode turnstile junction will be reviewed. This novel waveguide junction, shown in Fig. 1, is a 7-port hybrid device which couples three different transmission line modes in a variety of ways. Potter[7,8] has shown that the junction can be matched for any two, but not for all three modes simultaneously, and has tabulated the transmission coefficients between ports, which differ for each set of matching conditions. Table I shows the coupling characteristics of a trimode turnstile junction having matched coaxial and circular waveguide ports. Resultant reflected and transmitted components of power are tabulated for unit power input into each port in turn. Broadband performance of an experimental trimode turnstile junction designed to approximate the theoretical case tabulated is shown in Fig. 2. The insert drawing details the matching structure used.

In addition to the examples of junction match chosen by Potter[7,8] the junction can be adjusted to various intermediate conditions which exhibit other properties. In all cases, however, junction symmetry dictates that the circular and coaxial ports will be isolated from one another. The case of matched coaxial and circular waveguides has been detailed in Table I, since this is the junction condition used in the dual balanced mixer of the basic polarimeter.

[4] M. H. Cohen, "The Cornell radio polarimeter," PROC. IRE, vol. 46, pp. 183–190; January, 1958.

[5] S. Suzuki and A. Tsuchiya, "A time-sharing polarimeter at 200 mc," PROC. IRE, vol. 46, pp. 190–194; January, 1958.

[6] K. Akabane, "A polarimeter in the microwave region," PROC. IRE, vol. 46, pp. 194–197; January, 1958.

[7] R. S. Potter, "The Analysis and Matching of the Trimode Turnstile Waveguide Junction," NRL Rep. No. 4670; December 19, 1955.

[8] R. S. Potter, "Multiple Mode Excitation of the Trimode Turnstile Waveguide Junction," NRL Rep. No. 4802; August 10, 1956.

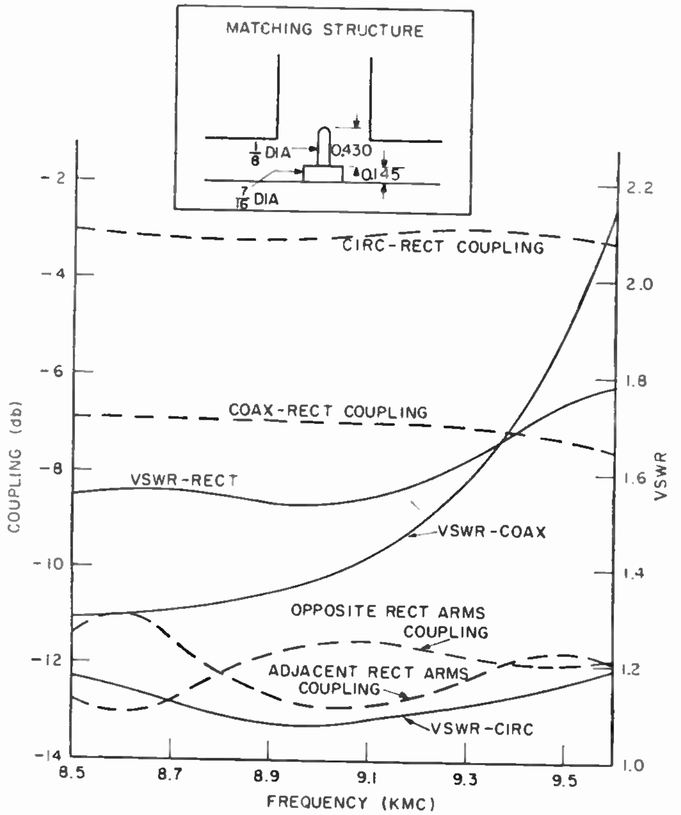

Fig. 2—Experimental approximation of the trimode turnstile junction characteristics listed in Table I (matched coaxial and circular waveguides). Insert drawing shows detail of matching structure used.

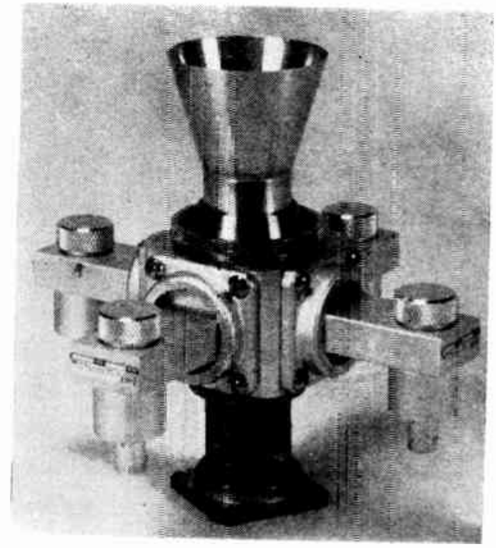

Fig. 3—X-band model of the trimode turnstile dual balanced mixer used in the microwave polarimeter. Local oscillator power is fed into rectangular waveguide at bottom while signal enters circular waveguide at top.

The Dual Balanced Mixer

The unusual coupling properties (Table I and Fig. 2) of the trimode turnstile junction make it ideally suited to a precision dual balanced mixer. The simplicity of the new mixer is evident from Fig. 3 which shows an experimental model employed in the microwave polarimeter. The rectangular waveguide at the bottom of the photograph is an end-on waveguide-to-coax transition to port 7 of Fig. 1 and serves as the local oscillator (LO) input. The dual mixer is unique in that it has a circular waveguide signal input and thus can accommodate *any* input polarization.

Consider what happens when an electromagnetic wave of arbitrary polarization is introduced into the circular waveguide arm of a trimode turnstile mixer in which the coaxial and circular ports are matched. Referring to Fig. 1, let arrows 5 and 6 represent orthogonal components of this input signal, which are not necessarily equal nor in phase. On entering the junction, component 5 will divide equally, but out of phase, between arms 1 and 3. Similarly, component 6 will divide equally, but out of phase, between arms 2 and 4.

It should be apparent from Fig. 1 how an LO input to port 7 will divide equally and *in* phase between the four rectangular arms 1, 2, 3, and 4. Mixer crystals attached to all four rectangular arms, equidistant from the junction, thus will be excited *in* phase by the LO input, while *opposing* crystals will be excited by equal *out-of-phase* components of the signal input to the circular waveguide. The outputs of opposing crystals can be combined in a number of ways to obtain balanced mixer operation, but it is convenient to use reversed matched crystals connected in dc series with a simple shunt connection for single-ended IF output, as advanced by Riblet.[9]

The Trimode Turnstile Polarimeter

Under conditions of linear mixing, the phase and amplitude relationships of the orthogonal components of the input signal to the circular waveguide will be accurately preserved in the IF signal outputs of the two balanced mixers, regardless of the IF chosen. The high degree of accuracy of this conversion operation, which can be achieved because of symmetry of the trimode turnstile mixer, is most important in its application to a precision polarimeter. The simple manner in which these IF signals are utilized to make visible the polarization characteristics of the input signal is illustrated by the block diagram of Fig. 4. Typically, the two IF signals are applied to orthogonal deflection planes of a cathode-ray tube, after the necessary amplification, to obtain a Lissajous figure which is a pictorial representation of the input signal polarization. A circularly polarized input signal will generate a circle, elliptical polarization will generate an ellipse which portrays the axial ratio and orientation of the input signal, and linear polarization will generate a line oriented to indicate the plane of polarization. This presentation is instantaneous, the Lissajous patterns being painted at the IF rate. Measurement of the polarization parameters of the inci-

[9] H. J. Riblet, "The short-slot hybrid junction," Proc. IRE, vol. 40, pp. 180–184; February, 1952.

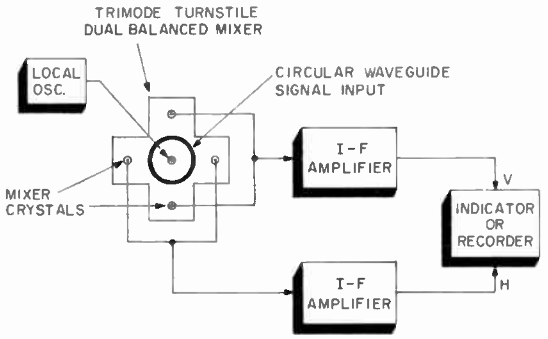

Fig. 4—Block diagram of the microwave polarimeter illustrating its simplicity.

dent wave can be made directly from the Lissajous pattern by means of established techniques.[10,11]

The commercial availability of high-gain, direct-coupled, XY type oscilloscopes with wide-band amplifiers providing identical phase and gain characteristics in both x and y planes, simplifies the presentation problem for many applications. Two types of commercial oscilloscopes have been used in the experimental work, a Weston model 983, and a Tektronix type 536. The Weston instrument has identical x- and y-axis amplifiers with a 3-db bandwidth of 4.5 mc and relative phase shift within 2° to 1 mc. The Tektronix 536 employs identical plug-in amplifiers for both axes. For the polarimeter application, type B, wide-band, high-gain amplifiers have been used. Bandwidth is greater than 10 mc, with less than 1° of phase error. An experimental embodiment of the polarimeter diagrammed in Fig. 4 is shown in Fig. 5. With exception of the trimode turnstile junction and conical horn, all necessary equipment is commercially available.

Excellent presentations have been obtained using separate reflex klystrons for signal source and for LO. These are tuned to produce an IF beat which is within the passband of the scope amplifiers. Satisfactory results were obtained without using automatic frequency control, as evidenced by Fig. 6 which shows a multiple exposure of crt presentations obtained with circular, elliptical, and linear polarizations.

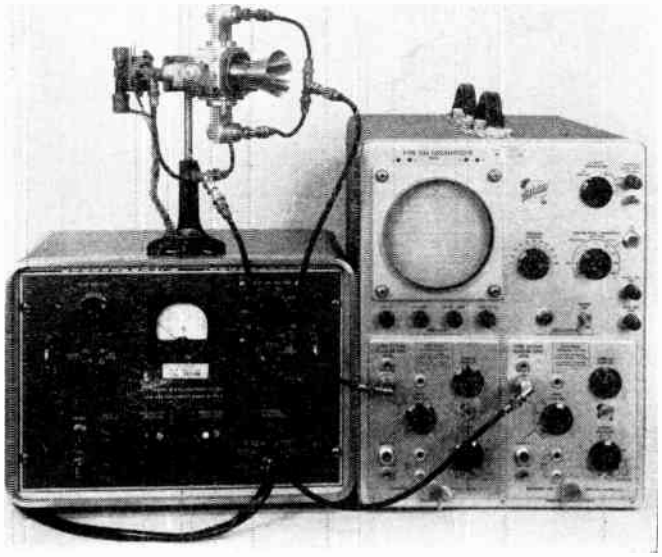

Fig. 5—Experimental setup of the trimode turnstile microwave polarimeter.

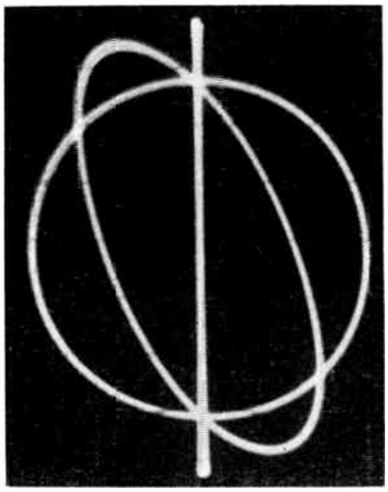

Fig. 6—Triple exposure showing polarization "pictures" obtained with circular, elliptical, and linear input polarizations. Photo was taken using polarimeter similar to that shown in Fig. 5.

Polarimeter Accuracy

By virtue of the precise symmetry of the dual mixer, phase and amplitude errors arise due only to reflections from the individual mixer crystals. Thus, a signal component reflected from a mismatched mixer crystal will couple to all other ports of the trimode turnstile in accordance with Table I (for an input to arm 1). Any signal coupled to the coaxial port will be absorbed in the LO padding device used in that arm, while any component coupled out the circular arm will either be reradiated, or absorbed in the matched source. Components of the reflected signal which couple to the adjacent arms can introduce crosstalk in the orthogonal mixer channel. The level of this crosstalk will depend on the magnitude and phase of mixer reflections, the adjacent-arm coupling factor or isolation (which in this case is −12 db), and the degree of in-phase cancellation achieved by mixer balance. For the assumed case of a mixer VSWR of 1.5:1 and 15-db cancellation by the balanced mixers, the crosstalk between the two IF channels will be down more than 40 db, corresponding to a maximum voltage error from this source of less than 1 per cent.

The major source of error, however, is due to that portion of a reflected signal which couples to the mixer on the *opposing* rectangular arm. The magnitude of this error can be estimated as follows. Consider opposing arms of a trimode turnstile which initially are excited

[10] H. D. Webb, "Accurate oscilloscope phase shift measurements," *Electronic Design*, vol. 4, pp. 34–35; January, 1956.

[11] J. F. Sodaro, "Phase shift by CRO," *Electronics*, vol. 26, pp. 192–194; May, 1953.

from the junction with signals of equal magnitude. In addition to the signal in each mixer arm, there will be an undesired component reflected from the mixer on the opposite arm. The resultant signal in each arm may be written in the form

$$E(1 + K_n e^{j\theta_n})e^{j\omega t}$$

where E is the amplitude of the desired signal, K_n is a function of the reflection from the opposite arm and the coupling coefficient between opposite arms, and θ_n is the phase factor of the reflected signal. If it is assumed that the crystals are operating as linear converters, and that their conversion coefficients k_c are identical, the outputs from a pair of crystals can be added to give a signal of the form

$$k_c E\left(1 + \frac{K_2 e^{j\theta_2} + K_1 e^{j\theta_1}}{2}\right).$$

Consideration of this expression shows that the maximum amplitude error occurs at zero phase error and has a magnitude of

$$\frac{\pm K_1 + K_2}{2} \times 100 \text{ per cent.}$$

The maximum phase error θ will have a value of $\tan^{-1}(K_1+K_2)/2$, and will be coupled with a small amplitude error of magnitude $\sqrt{1+(K_1+K_2)^2}$.

For the mixer characteristics assumed above (*i. e.*, maximum VSWR per crystal 1.5:1 and identical conversion coefficients), the cross-coupled components K_n are down 26 db. In the worst case, then, the maximum error from this source in the IF output of either balanced mixer will be ± 5 per cent in amplitude, or $\pm 3°$ in phase. However, both the amplitude and phase errors are not maximum simultaneously.

Errors of this magnitude are quite tolerable for many applications of the polarimeter. However, further improvement in accuracy of the polarimeter is possible. It has been stated that the principal source of error is due to cross-coupling of signal components reflected from the mixer crystals. The lower the crystal mismatch, the smaller the error. However, assuming that the practical limit of broadband crystal match has been achieved, the addition of identical ferrite isolators to each of the four mixer arms will serve to absorb crystal reflections, and—assuming low input VSWR to the isolators—cross-coupling will be minimized with consequent reduction in IF phase and amplitude errors. Isolator reverse loss of 20 to 30 db is adequate; and if forward losses are low, small differences will not introduce appreciable error in the polarimeter.

When using isolators in the mixer arms, the principal source of error will be due to differences in electrical lengths of the isolators in the forward direction. Although the phase error in the output of either of the balanced mixers will be only one-half the differential phase error between the two isolators used in that mixer (for small differential phase errors, say less than 10°), this might become significant if the dispersive characteristics of the isolators differ appreciably. By using isolators which are electrically identical, a high degree of accuracy is possible with this polarimeter technique. Isolators having at least 20-db reverse attenuation with an input VSWR of 1.1:1 or less will insure that the ratio of desired signal to cross-coupled signal will be greater than 40 db. Under these conditions, phase and amplitude errors due to reflections from the crystals will be negligible.

Accurate presentation of the polarization information contained in the IF signals depends, of course, on the gain and phase equality of the two IF amplifiers employed. Where signal level is adequate, the XY oscilloscope may be all that is required. For very small signals, such as radar return, etc., considerably more gain will be required. This is a special circuitry problem, the requirements depending on the application, and will not be treated here. In practice, unless extreme care is exercised, polarimeter accuracy may be seriously limited by the amplifying and presentation system employed. The object of this paper is to show a simple solution to the RF circuitry problem which is the *basic* source of error in the polarimeter.

Operation with Single Mixers

If the application of the polarimeter is such that a 3-db signal loss can be tolerated, high accuracy is possible by using but a single mixer in each plane, with the spare rectangular arms terminated in "perfect" matched loads. In this manner the opposite-arm reflection problem is eliminated. Crosstalk, due to adjacent-arm coupling, then becomes the major problem, since the cancellation feature of the balanced mixers is no longer present. However, it is possible to obtain greater isolation between adjacent arms of the junction than indicated in Table I by changing the configuration of the matching structure in the junction to alter the matching conditions. To attain this change in coupling requires, of course, a change in other characteristics of the junction as prescribed by the unitary requirements of the junction scattering matrix.

Potter and Sagar[12] have shown that complete isolation can be achieved between adjacent rectangular waveguide arms in an *ordinary* turnstile junction. In applying this principle to the trimode turnstile junction, a good compromise has been obtained between adjacent-arm isolation and other properties of the junction. The junction can be designed to provide adjacent-arm isolation of 20 db or more over a 12 per cent band as shown by the experimental data plots in Fig. 7. This degree of isolation was achieved in an experimental model by accepting an appreciable mismatch of the coaxial port. As

[12] R. S. Potter and A. Sagar, "A new property of the turnstile waveguide junction," *Proc. NEC*, vol. 13, pp. 452–458; 1957.

a consequence, the coaxial-to-rectangular arm coupling is reduced to −10 db (cf. Figs. 2 and 7). This mismatched condition can be tolerated in this application by using an attenuator pad or ferrite isolator in the LO arm. The insert drawing in Fig. 7 details the junction matching configuration employed, and shows the disk iris used in the circular waveguide arm to keep input VSWR under 1.1:1 over the 12 per cent band.

With adjacent-arm isolation of 20 db and using single mixers whose maximum VSWR is 1.5:1, amplitude error should not be more than 2 per cent, without the use of isolators in the mixer arms. If identical isolators are added to the two mixer arms, the accuracy of the microwave portion of the polarimeter would then be determined primarily by quality of input match to the isolators, and a high degree of accuracy should be possible.

Determining Sense of Polarization

Although the trimode turnstile polarimeter provides an instantaneous pictorial representation of the input polarization, one bit of information is not presented explicitly. This is the "sense" of polarization, or the direction of rotation of the electric vector in the case of elliptically or circularly polarized input. This information is actually contained in the IF signals and determines the direction of motion of the spot on the crt as it continually replots the polarization locus at the IF rate, but cannot be discerned visually except when the IF is a few cps or less. Novel techniques have been used for determining the direction of spot rotation[13,14] but are difficult to employ at megacycle frequencies especially if the IF is not constant.

The sense of input polarization can be determined from knowledge of the relative phase[1,15] between the two IF signals, or by determining the direction of rotation of the spot on the crt. Theoretically, this might be done by slowly tuning the LO through the signal frequency and noting the direction of spot motion as the IF approaches zero beat. In practice, however, it is difficult to observe near-zero-beat between two free-running klystron oscillators. In this regard, it is important to realize that the direction of spot motion *reverses* in passing through zero beat between a "positive" and a "negative" IF. Thus, for consistent results in determining "sense" by this method, the LO should be maintained on the same side of the signal frequency (*i.e.*, *either* above or below the signal frequency).

A properly oriented quarter-wave plate attached to the circular waveguide signal input will resolve the right- and left-handed components of the input signal so that "right" goes to one mixer channel and "left" goes

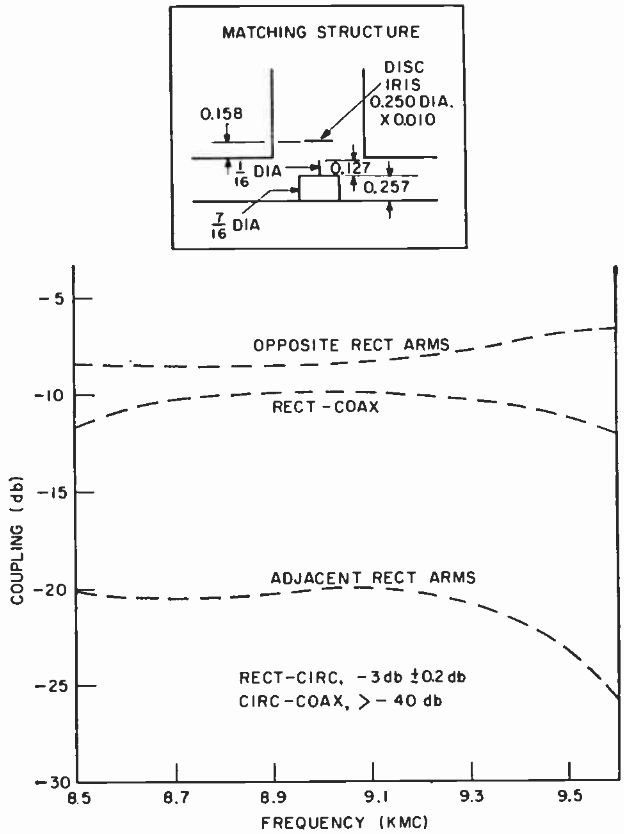

Fig. 7—Experimental performance of a trimode turnstile junction adjusted for 20-db isolation between adjacent rectangular waveguide ports. The increased isolation improves polarimeter accuracy when single mixers are used. Insert drawing shows detail of matching structure used.

to the other mixer channel. In this manner the relative amplitudes of the two can be directly measured, and the sense of the input polarization determined. The quarter-wave plate should be removed for direct observation of input polarization. In some applications it may be of advantage to use a separate dual mixer of this type, sensitive to right- and left-hand circular polarizations, to indicate continuously which component dominates.

Shifted Carrier Operation of the Polarimeter

For many laboratory applications of the polarimeter, it may be convenient and of advantage to use the same signal generator or klystron for both LO and signal source. This can be done by employing some form of single-sideband (SSB) generator to shift the frequency of the RF component used for the polarized signal. A block diagram of such a system is shown in Fig. 8. This method eliminates the need for a second signal generator and provides a constant IF output at the frequency used to modulate the SSB generator. By this technique the IF can be made a low audio frequency, and the IF outputs of the two mixers then can be recorded directly on a continuous chart or on magnetic tape for subsequent study.

The SSB generator can take many forms, one of the

[13] E. R. Mann, "A device for showing the direction of motion of the oscilloscope spot," *Rev. Sci. Instr.*, vol. 5, pp. 214–215; 1934.
[14] J. R. Haynes, "Direction of motion of oscilloscope spot," *Bell Lab. Rec.*, vol. 14, pp. 224–225; 1936.
[15] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 947–950; 1943.

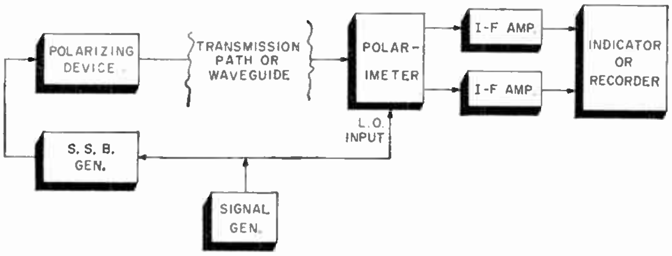

Fig. 8—One signal generator can be used for both LO and signal source in a closed polarimeter setup by incorporating a single-sideband generator to shift the carrier frequency of one or the other. This method is convenient for recording since a constant, low-frequency IF is readily obtained in this manner.

simplest being a mechanically driven continuous phase shifter.[16] Other types, which are electronically driven, employ ferrite devices, crystal diodes, traveling-wave tubes, etc., to accomplish the desired frequency translation. The SSB generator can, however, be the source of appreciable error in the polarimeter if it introduces any amplitude modulation in the IF output. Since the LO power at the mixer is much larger than the signal power, amplitude modulation effects will be minimized, when employing single mixers, if the SSB generator is inserted in the signal path rather than in the LO path. However, when using balanced mixers, the SSB generator should be in the LO arm, to take advantage of the in-phase cancellation obtained in the mixers. The stable low-frequency IF possible with the SSB method simplifies the problem of determining the polarization sense. Thus, the frequency shift introduced by the SSB generator can be reduced to the point where spot motion can be observed directly, or the phase relationship of the two IF signals can be measured directly or monitored using a commercial phase meter.

[16] A. G. Fox, "An adjustable wave-guide phase changer," Proc. IRE, vol. 35, pp. 1489–1498; December, 1947.

Conclusions

A simple new microwave polarimeter technique which permits instantaneous viewing of the polarization of an input signal has been described. Utilizing a trimode turnstile junction in a precision dual balanced mixer for circular waveguide, the new polarimeter is capable of high accuracy over a wide frequency band, achieved through symmetry of the RF plumbing. Amplitude errors of less than 2 per cent over a 12 per cent band are possible with present designs.

The compact size of the trimode turnstile polarimeter makes it a convenient instrument for both laboratory and field use in measuring polarization. Because of its instantaneous character, it is particularly valuable where signal input is variable or intermittent. The technique should have immediate utility in such fields as antennas, propagation, ferrite devices, radar return studies, signal intercept, countermeasures, communication, and radio astronomy.

Thin Film Magnetization Analysis*

K. CHU† AND J. R. SINGER†, MEMBER, IRE

***Summary*—A "Graphical Method" is used for analysis of the magnetization direction in terms of the magnetic energy, and for prediction of the hysteresis loop shape of ferromagnetic thin films. Three major magnetization conditions are discussed: 1) condition of magnetization with an 180° field applied: 2) condition of magnetization with a 90° field applied; and 3) condition of magnetization with both the 90° and the 180° field applied. Corresponding to these magnetization conditions, the hysteresis loop shapes are predicted and constructed showing close identity to those experimentally observed. These conditions are chosen because of their importance to computer applications.**

In the graphical method, the principle of superposition applies. Using this method, the magnetic energy relationships are readily established. The general expression for the total free energy equation is:

$$E_t = E_k + E_\sigma + E_{H\theta}$$

where E_k is the anisotropy energy, E_σ is the magnetostriction energy, and $E_{H\theta}$ is the magnetization energy. Since the method of construction of the total free energy curve as well as the hysteresis loop is simple and mechanical, a quite complex magnetization condition with a multiplicity of fields of different magnitudes and directions may be simplified and handled by graphical means with ease. Thus, the present scheme should be a practical and useful tool for analysis and engineering design of magnetic devices and systems utilizing ferromagnetic thin films.

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† Elec. Engrg. Dept., University of California, Berkeley.

Introduction

THE preparation and magnetic properties of thin films (~1000Å thick) of vacuum evaporated 80-20 Permalloy have been discussed by Blois.[1] When the film is deposited on a heated glass substrate (microscope slide) in the presence of an external magnetic field in the plane of the film, the film develops a preferred magnetic axis in the field direction. The uniaxial direction of easy magnetization in films is apparent from the domain patterns by means of the well known Bitter Technique,[2,3] and the longitudinal Kerr magneto-optic method.[4,5,6] A rotational magnetization process instead of the usual 180° domain wall movement occurs during the process of magnetization reversal because it is energetically favored in the case of the thin films. A theoretical argument[7] lends weight to this conjecture. Conger[8] showed by a hysteresis loop tracer that 80-20 Permalloy thin films have rectangular hysteresis loops. Using an M-loop tracer, the process of magnetization reversal has been shown to be by a rotation in the plane of the film.[8]

It is the purpose of this paper to analyze the magnetization phenomena and to predict the hysteresis loop shape by examining the total free energy expression in detail. In this connection, a "Graphical Method" is introduced which is then used for analysis of the magnetization phenomena and prediction of the hysteresis loop shape of magnetic thin films.

The Graphical Method

Since most of the magnetic thin films thus far developed possess a uniaxial easy direction of magnetization, we may assume that for these materials the anisotropy energy may be represented by[9]

$$E_k = k_0 + k_1 \sin^2 \theta \tag{1}$$

where k_0 and k_1 are anisotropy constants, and θ is the angle between the magnetization vector and the axis of easy direction of magnetization. The unit for E_k is erg/cm³.

The magnetostriction energy may be expressed as[9]

$$E_\sigma = (3/2)\lambda_s\sigma \sin^2 \theta \tag{2}$$

where λ_s is the magnetostriction expansion occurring between the demagnetized state and saturation, σ is the tension, and θ is the angle between magnetization and tension.

Since we are interested in the free energy, the lowest energy state may be set at zero and the magnetization energy is[9]

$$E_{H\theta} = -HI_s[\cos(\theta_0 - \theta) - 1] \tag{3}$$

where H is the field strength, I_s is the saturation magnetization, θ_0 is the angle between the direction of easy magnetization and the direction of the applied magnetic field, and θ the angular displacement of the magnetization vector from its original easy magnetization axis.[10]

The energy curves representing (1) and (3) are shown in Fig. 1. These energy curves describe the magnitude of the energy components and their relation to the magnetization direction.

The general expression for the total free energy equation may be written as

$$E_t = E_k + E_\sigma + E_{H\theta} \tag{4}$$

In the construction of the total free energy curve, the principle of superposition applies. The total free energy curve is obtained by adding the corresponding energy components of the magnetic energy terms proper. The slope of the E_t curve is, by definition, $\partial E_t/\partial\theta$. Therefore, at any point along the curve of E_t at which the slope is equal to zero, a possible relative minimum energy point exists.

A Nomogram

Once a particular type of magnetic material is chosen for the design, the terms E_k and E_σ in (4) are fixed. The only variable parameter in the equation will be $E_{H\theta}$. This variable appears whenever an analysis involves an external magnetic field. For convenience in further analysis a nomogram with a family of the $E_{H\theta}$ curves appears in Fig. 2. This was constructed by inserting discrete values of H at equal intervals into (3). For each value of the applied H, an $E_{H\theta}$ curve was plotted. A family of the $E_{H\theta}$ curves results, which is called the nomogram (see Fig. 2). This nomogram of the magnetization energy may be used for the evaluation of the magnetization condition by superimposing it onto the $E_k + E_\sigma$ curve. In the event that more complicated magnetization conditions appear than those considered here, the present method is still applicable. The present treatment will use simple examples to aid in the description.

[1] M. S. Blois, Jr., "Preparation of thin films and their properties," J. Appl. Phys., vol. 26, pp. 975–980; August, 1955.

[2] H. Williams, R. Bozorth, and W. Schockley, "Magnetic domain patterns on single crystals of silicon iron," *Phys. Rev.*, vol. 75, pp. 155–178; January, 1949.

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[4] C. A. Fowler and E. M. Fryer, "Magnetic domains in thin films of nickel-iron," *Phys. Rev.*, vol. 100, pp. 746–747; October, 1955.

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[7] C. Kittel, "Theory of the structure of ferromagnetic domains in films and small particles," *Phys. Rev.*, vol. 70, pp. 965–971; December, 1946.

[8] R. L. Conger, "High frequency effect in magnetic films," in *Proc. AIEE Conf. on Magnetism and Magnetic Materials*, pp. 610–619; February, 1957.

[9] R. M. Bozorth, "Ferromagnetism," D. Van Nostrand Co., Inc., New York, N. Y.; 1951.

[10] K. Chu, "Investigation of ferromagnetic thin films," Engineering Library, University of California, Berkeley; 1958.

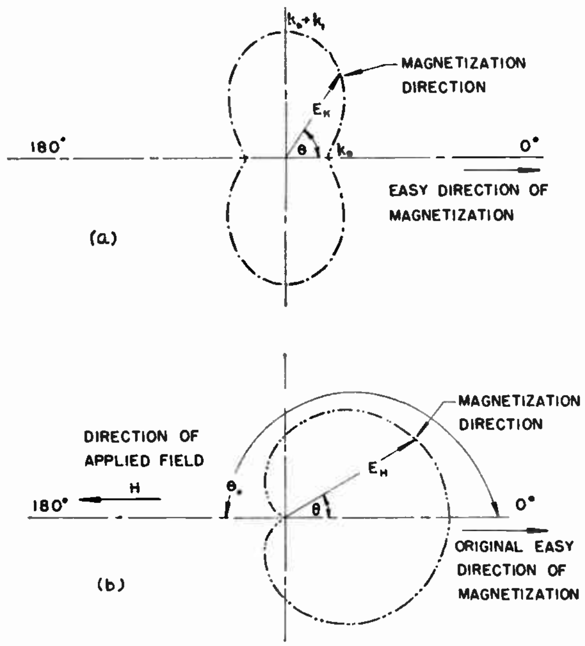


Fig. 1—Energy curves for (a) the anisotropy energy: $E_k=k_0+k_1 \sin^2\theta$, and (b) the magnetization energy:

$E_{H\theta}=-HI_s[\cos(\theta_0-\theta)-1]$. (Polar plot.)

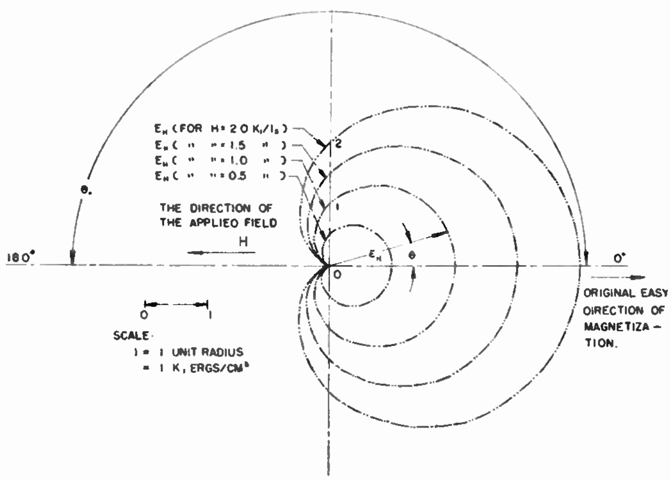


Fig. 2—A nomogram of the magnetization energy: $E_{H\theta}=-HI_s[\cos(\theta_0-\theta)-1]$. (Polar plot.)

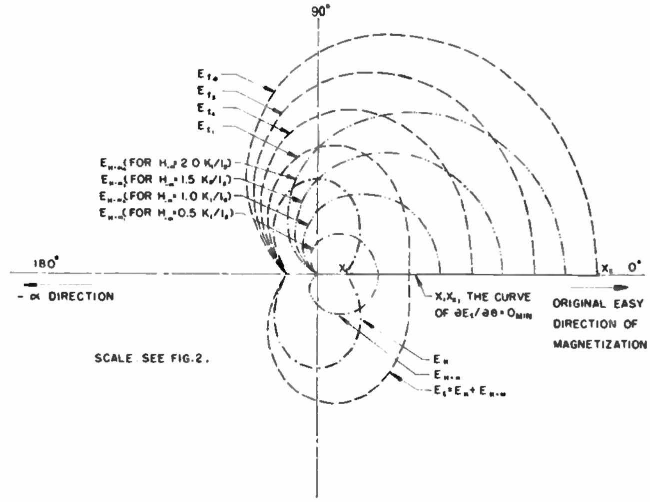
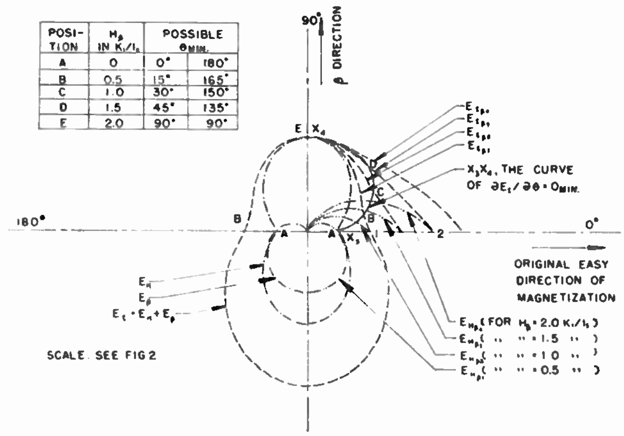


Fig. 3—The energy curves for (1), condition of magnetization with a 180° field applied. (Polar plot.)



POSITION	H_β IN K_1/I_s	POSSIBLE θ_{MIN}	
A	0	0°	180°
B	0.5	15°	165°
C	1.0	30°	150°
D	1.5	45°	135°
E	2.0	90°	90°

Fig. 4—The energy curves for (2), condition of magnetization with a 90° field applied. (Polar plot.)

Graphical Analysis of the Magnetization Phenomena in Terms of the Magnetic Energy

For simplicity, let E_σ be small and negligible compared to E_k. This corresponds to the Permalloy (80-20) case. We now consider several magnetization conditions:

Condition of Magnetization with a 180° Field Applied

A graphical analysis of the magnetization condition is presented in Fig. 3. The E_t curves may be obtained by the following procedure. First, superimpose the nomogram of the $E_{H\theta}$ curves onto the E_k curve in a manner such that the direction of the applied magnetic field is 180° from the original direction of the magnetization vector. Secondly, add up the corresponding energy components. The total energy equation may be written as:

$$E_t = E_k + E_{H_{-\alpha}} \tag{5}$$

where $-\alpha$ denotes that the applied field is 180° from the original direction of the magnetization vector. From Fig. 3, one may observe that:

a) In the absence of an external magnetic field, the position of stability is at 0° which is a relative minimum energy point. The relative maximum energy point is at the 90° position.

b) When $H_{-\alpha}$ is applied and the value of $H_{-\alpha}$ is gradually increased, the minimum energy point is unchanged until a value of $H_{-\alpha}=2k_1/I_s$ is reached. Then, rotation of the magnetization vector takes place. The 0° position is no longer the minimum energy point, and the new minimum energy point is now at the 180° position. This shows why a rectangular hysteresis loop may be obtained by the rotational process in magnetic thin films.

c) Symmetry about the 0°–180° line is observed. Thus, it is only necessary to plot half of the curves as shown.

Condition of Magnetization with a 90° Field Applied

A graphical analysis of the magnetization condition is shown in Fig. 4 (p. 1239) in which,

$$E_t = E_k + E_{H_\beta} \tag{6}$$

where β denotes that the direction of the applied field is 90° from the initial direction of the magnetization vector. In the graph, the following may be noted:

a) Upon the application of a magnetic field H_β, it may be seen that the minimum energy point is shifted away from its 0° position. As the value of H_β increases, the position of the relative minimum energy point changes such that the direction of the magnetization vector is gradually rotated towards the 90° position. This rotational magnetization process is indicated in Fig. 4 by the curve of $\partial E_t/\partial\theta = 0_{\min}$.

b) It is of interest to note, from the curve of $\partial E_t/\partial\theta_{\min}$, and will be discussed below, that this case of the rotational magnetization phenomenon corresponds to a zero area hysteresis loop.

c) Upon the collapse of the applied field, the magnetization vector will rotate to and rest at its nearest minimum energy point. For example, in the case of $H_\beta = 1.5k_1/I_s$, the magnetization vector is at about the 45° position. When $H_\beta = 0$ again, the nearest new minimum energy point is the 0° position. Therefore, the magnetization vector will rotate back and rest at this nearest minimum energy point. It is of interest that the magnetization process is reversible. This reversible magnetization process is referred to as the "non-destructive read-out mode" in the magnetic memory. It is so named because the information stored (*i.e.*, the "0" state or the "1" state of a bit) can be detected and its original polarity retained after detection.

Now, examine the special case of $H_\beta = 2k_1/I_s$. We note that the magnetization vector is at the 90° position. In this case, when $H_\beta = 0$ again the magnetization vector has two nearest minimum energy points, namely the 0° position and the 180° position. Two directions are equally probable for rotation. It may be noted that under some conditions a poly domain state is observed after removal of the transverse field.

d) In order to design a reliable non-desctructive readout system, we must discard the minimum energy point at the 90° position as a possible site for the magnetization vector because of the uncertainty. That is, the magnetization vector should not rotate that far. As a matter of fact, the region from 90° to 60° may be regarded as a poor region for non-destructive operation. The reason is explicit and is indicated in Fig. 4. From the graph, it may be seen that within this region, the energy difference between the minimum energy point at the 60° position and the maximum energy point at the 90° position is very small. The energy difference, ΔE, is in the order of 0.01 K_1/I_s. ΔE may be viewed as a margin of safety to insure that the magnetization process will be reversible. For the minimum energy point at D of Fig. 4, $\Delta E \simeq 0.08k_1/I_s$.

e) Symmetry about the 90°–270° line is observed. Thus, the curves to the left of the line may be omitted. Further investigations indicate that the curves in the fourth quadrant may also be omitted and neglected because the values of the total energy components in it are very much higher than those in the first quadrant. Therefore, the magnetization vector will not be found in the fourth quadrant because it is energetically unfavorable. Thus symmetry properties and careful reasoning often simplifies the analysis considerably.

Condition of Magnetization with Both the 90° and the 180° Field Applied

The graphical analysis of the magnetization conditions are shown in Figs. 5–8 in which

$$E_t = E_k + E_{H_\beta} + E_{H_{-\alpha}}. \tag{7}$$

We consider the case of a dc biasing field (*i.e.*, a field of constant amplitude) 90° to the uniaxial axis. In order to study the effect of this transverse dc bias on magnetization, we will use four different values of dc biasing. These are separately applied and analyzed. These values are shown in the respective figures.

Fig. 5 was constructed by first superimposing onto the E_k curve a curve of E_{H_β} with $H_\beta = 0.5k_1/I_s$. This simulates the condition of an applied 90° dc bias of magnitude of $0.5k_1/I_s$. Then, superimpose onto this combined energy curve $E_{t\beta_1}$ a nomogram of the $E_{H_{-\alpha}} =$ 0.25, 0.50, 0.75, and $1.0k_1/I_s$, respectively. This simulates the application of a switching field with increasing field strength. The total free energy curve represents the condition of magnetization under the influence of both fields. From Fig. 5, the following may be noted: a) When $H_{-\alpha} = 0$ and $H_\beta = 0.5k_1/I_s$, the minimum energy point is not at the 0° position but at about the 15° position. b) When $H_{-\alpha}$ is applied and increased, the minimum energy point shifts upward as indicated by the curve of $\partial E_t/\partial\theta = 0_{\min}$. c) When $H_{-\alpha} = 1.0k_1/I_s$, the minimum energy point is suddenly rotated to the position where it is about 171°, and this is the new minimum energy point. d) It may be observed that it is energetically unfavorable for the magnetization vector to exist in the third and fourth quadrant. Thus, the portions of the curves below the 0°–180° line are omitted and neglected.

Fig. 6 was constructed similar to Fig. 5. In this case, an $H_\beta = 1.0k_1/I_s$ was applied. From Fig. 6, we note: 1) When $H_{-\alpha} = 0$, the minimum energy point is at about the 30° position, which is farther away from the 0° position than for the case of $H_\beta = 0.5k_1/I_s$. 2) When $H_{-\alpha}$ is gradually applied and increased, the minimum energy point shifts upward along the curve of $\partial E_t/\partial\theta = 0_{\min}$. 3) When $H_{-\alpha} = 0.5k_1/I_s$, the magnetization vector is suddenly rotated to about the 157° position.

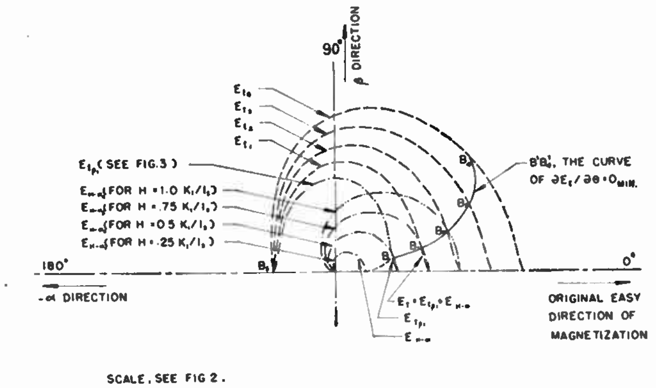


Fig. 5—The energy curves for (3), condition of magnetization with both the 90° and the 180° field applied; the applied 90° field is a dc biasing field of $H_\beta = 0.5k_1/I_s$.

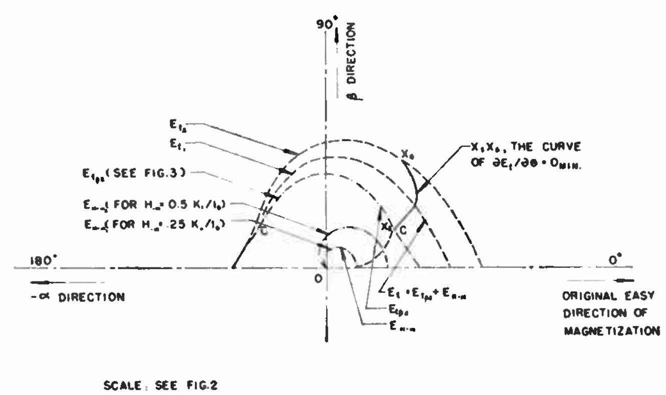


Fig. 6—The energy curves for (3), condition of magnetization with both the 90° and the 180° field applied; the applied 90° field is a dc biasing field of $H_\beta = k_1/I_s$. (Polar plot.)

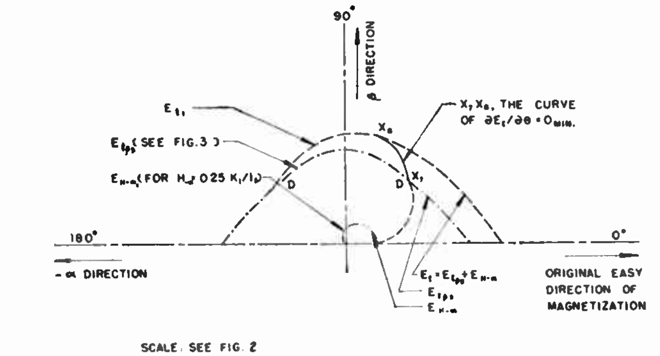


Fig. 7—The energy curves for (3), condition of magnetization with both the 90° and the 180° field applied; the applied 90° field is a dc biasing field of $H_\beta = 1.5k_1/I_s$. (Polar plot.)

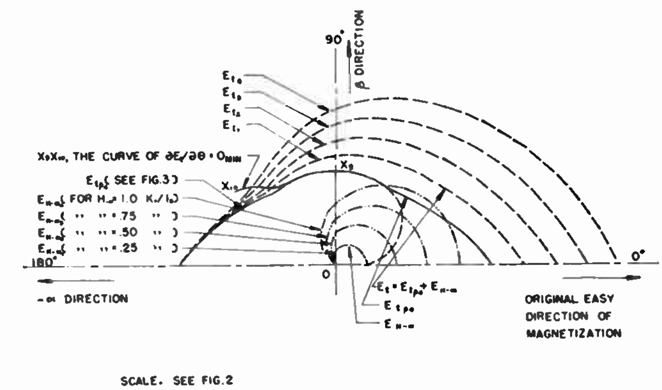


Fig. 8—The energy curves for (3), condition of magnetization with both the 90° and the 180° field applied; the applied 90° field is a dc biasing field of $H_\beta = 2k_1/I_s$. (Polar plot.)

Fig. 7 was constructed in a similar way. In this case, an $H_\beta = 1.5k_1/I_s$ was applied. Before the application of $H_{-\alpha}$, the minimum energy point is at about the 45° position. With the application of an $H_{-\alpha}$ of about $0.25k_1/I_s$, the magnetization vector is immediately rotated to a new minimum energy point at about the 137° position.

Fig. 8 was constructed similarly. In this case, an $H_\beta = 2k_1/I_s$. From Fig. 8, the following is of interest: a) Before the application of $H_{-\alpha}$, the minimum energy point is at the 90° position. b) When an $H_{-\alpha} = 0.25k_1/I_s$ is applied, the magnetization vector is immediately rotated to the 130° position which is the new minimum energy point. c) As $H_{-\alpha}$ further increases, the minimum energy point shifts along the curve of $\partial E_t/\partial\theta = 0_{\min}$. d) The condition of Fig. 8 differs from those of Figs. 5–7 in this respect: when the $H_{-\alpha}$ field vanishes, the magnetization vector will always rotate back to the 90° position where the energy is minimum as long as the dc bias of $H_\beta = 2.0k_1/I_s$ is applied.

Prediction of the Hysteresis Loop Shape

The "Unit Magnetization Circle"

Magnetic thin films with a single domain have a constant magnitude of magnetization. It is, therefore, possible to construct a "Unit Magnetization Circle" as shown in Fig. 9. The radius of the "Unit Magnetization Circle" is arbitrarily chosen; its magnitude is equal to the normalized value of saturation magnetization of the magnetic material, whatever it may be. Thus, it is named the "Unit Magnetization Circle."

Now, let the uniaxial easy magnetization axis be along the 0° axis, and the position of the magnetization vector at the 0° position. This 0° position is one of the two possible minimum energy stable points when no external field is applied. The other minimum energy point is at the 180° position. These two points of saturation magnetization are designated by A on the "Unit Magnetization Circle."

Now, let a pickup coil be placed with its axis along the 0° axis. The pickup coil will always be so positioned unless otherwise indicated. The application of a switching field of magnitude $H_{-\alpha} = 2.0k_1/I_s$ will cause saturation in the opposite direction. The total flux change corresponds to a change in the flux density from 0° to 180° which is of the magnitude indicated by the component AA along the 0°–180° line of Fig. 9.

However, if a 90° field of constant amplitude is applied to the magnetic material (*i.e.*, if $H_\beta \neq 0$) the minimum energy point will not be at the 0° position. The functional dependence of the minimum energy point on H_β, when $H_{-\alpha} = 0$, can be seen in Fig. 4. It can also be seen in Figs. 5–8, and it is tabulated in Fig. 4. In Fig. 9, the positions of the magnetization vector corresponding to θ_{min} are located on the "Unit Magnetization Circle" as indicated by B, C, D, and E, respectively.

The pickup coil sees only the component of the moving magnetization vector that is parallel to the axis of the pickup coil. This component can be obtained by drawing a line perpendicular to the 0°–180° line from the individual points on the "Unit Magnetization Circle." The projected points B', C', D', and E' on the 0°–180° line correspond to the effective values of the flux density B on the B-axis of the B-H loop as seen by the pickup coil under different applied values of H_β before and after the application of the switching field $H_{-\alpha}$ (*i.e.*, when $H_{-\alpha}=0$). These points are shown in Fig. 9.

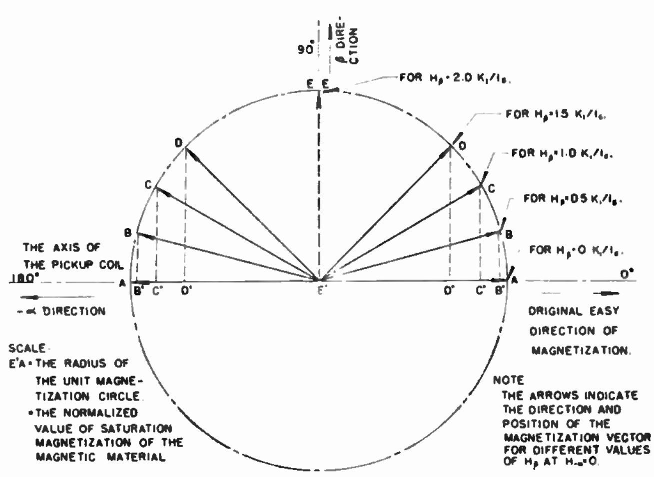


Fig. 9—The "Unit Magnetization Circle." (Polar plot.)

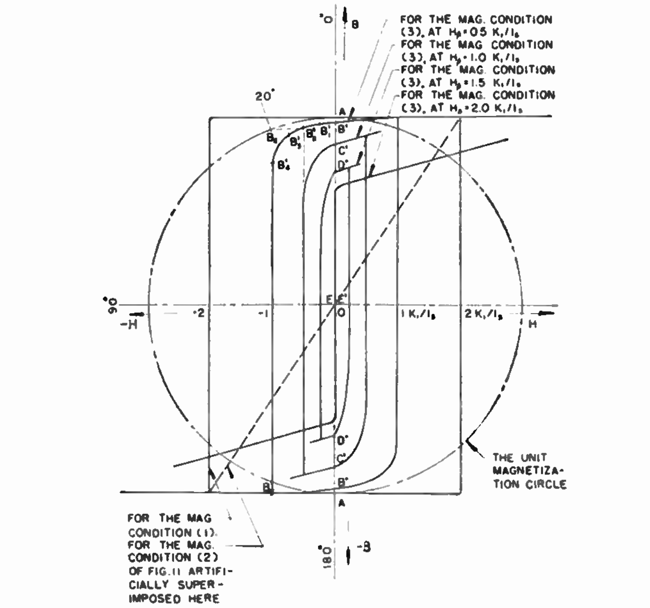


Fig. 10—Hysteresis loops for the different magnetization conditions.

The Hysteresis Loop

Theory used to obtain a theoretical hysteresis loop:

On the total free energy curve, the position of the magnetization vector is determined by the condition of $\partial E_t/\partial\theta=0_{min}$. The curve of $\partial E_t/\partial\theta=0_{min}$ expresses the functional dependence of θ_{min} on H. From this curve, for a known value of H, a corresponding value of the θ_{min} is determined.

Now, for a rotational magnetization process and for the axis of the pickup coil along the uniaxial magnetization axis, the component of the moving magnetization vector that cuts the pickup coil can be expressed by $B=I_s\cos\theta$. In the case when the axis of the pickup coil is 90° to the uniaxial magnetization axis, $B=I_s\sin\theta$. These relationships are expressed by the "Unit Magnetization Circle." From the "Unit Magnetization Circle," for a known value of θ, a corresponding value of B as experienced by the pickup coil can be determined.

By the above methods, the values of H and B are separately derived corresponding to a given value of θ. Therefore, for each value of the θ_{min}, a pair of values of H and B can be determined. From a series of data of the corresponding pairs of H and B thus obtained, the relationship of B to H is defined. When plotted, it is the B-H loop, or the hysteresis loop.

Method of construction:

Fig. 10 is a plot of the B-H loops, or the hysteresis loops. The units for both H and B are in units of k_1/I_s. The saturation magnetization value on the B-axis is set equal to the normalized radius of the "Unit Magnetization Circle." The two maximum points for saturation magnetization are designated by A on the B-axis of the B-H loop in Fig. 10. Other points such as B', C', D', and E' are also transferred from Fig. 9 to the "Unit Magnetization Circle" in the background of Fig. 10, and are reproduced on the B-axis of Fig. 10.

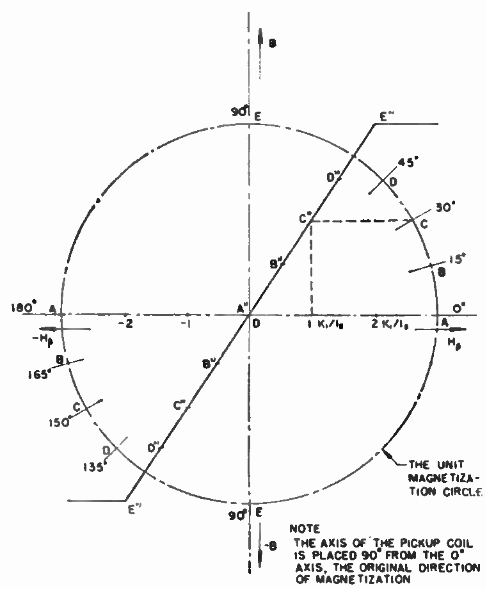


Fig. 11—The hysteresis loop for (2), condition of magnetization with a 90° field applied.

The construction of the B-H loop for the magnetization condition (1) will be presented first. In this case, $H_\beta=0$. Since no rotation of the magnetization vector takes place until a value of $(H_{-\alpha}) = 2k_1/I_s$ is reached,[11] the appropriate plot consists of two horizontal lines past points A and A, and two vertical lines through the two points at which $H_{-\alpha} = +2k_1/I_s$ and $-2k_1/I_s$. The loop thus constructed by the joining of two pairs of parallel lines is a rectangular hysteresis loop which is in agreement with that experimentally observed.[8]

We now go on to the construction of the B-H loop for the magnetization condition (2). In this case, $H_{-\alpha}=0$ and the axis of the pickup coil is placed 90° to the uniaxial magnetization axis. Fig. 4 tabulates the relationships of H_β and $\theta_{\min}$. Corresponding to each value of the $\theta_{\min}$, a point was plotted on the "Unit Magnetization Circle" of Fig. 11. These points are designated as A, B, C, D, and E, respectively. The points on the B-H loop were plotted by drawing a horizontal line and a vertical line through the corresponding values of B and H. The point of intersection determines the position of a point of the B-H loop. For instance, C'' was plotted by intersecting the horizontal line through C and the vertical line through $H_\beta = k_1/I_s$. Other points such as A'', B'', D'', and E'' were obtained by the same procedures used for locating C''. It may be seen that the B-H loop constructed in this way is a straight line within the range of $H_\beta = +2k_1/I_s$ and $-2k_1/I_s$. This behavior of the B-H relationship for the magnetization condition is in agreement with that shown by Smith.[12]

Finally, we shall consider in detail the construction method of the B-H loop for the magnetization condition (3). First, consider the case for $H_\beta = 0.5k_1/I_s$. From Figs. 5 and 9, it may be seen that for $H_{-\alpha}=0$, the magnetization vector is at $\theta_{\min}=15°$, which position is designated by B on the "Unit Magnetization Circle" of Fig. 9. Its component along the axis of the pickup coil as seen by the pickup coil is denoted by B' on the 0° axis as shown in Fig. 9.

Now, locate the point B' on the B-axis of Fig. 10. The point B' may be viewed as being plotted by drawing a horizontal line through B and a vertical line through $H_{-\alpha}=0$ from which the point of intersection is the location of B'.

For $H_{-\alpha}=0.25k_1/I_s$, Fig. 5 shows that $\theta_{\min}\simeq 16°$. Therefore, locate a point B_1 on the "Unit Magnetization Circle" at $\theta=16°$ in Fig. 10. A corresponding point B_1' is plotted by drawing a horizontal line through B_1 and a vertical line through $H_{-\alpha}=0.25k_1/I_s$. The point of intersection of these two lines is the location of B_1'. The second point of the B-H curve is thus located. Points B_2, B_3', and B_4' were constructed in a similar way.

For $H_{-\alpha}=1k_1/I_s$, Fig. 5 shows that $\theta_{\min}=40°$. Now, for a very small increase of $H_{-\alpha}$, Fig. 5 indicates that a sudden rotation of the magnetization vector occurs, and that the new minimum energy point is now at $\theta_{\min}\simeq 171°$. Therefore, locate a point B_5 on the "Unit Magnetization Circle" at $\theta=171°$ in Fig. 10. A corresponding point B_5' is plotted by drawing a horizontal line through B_5, a vertical line through $H_{-\alpha}=1.0^+k_1/I_s$. The point of intersection of these two lines is the location of B_5'. The sixth point of the B-H curve is thus located.

Now, consider the case when the switching field is released after performing the operation of reverse magnetization. That is, let $H_{-\alpha}=0$, again. Then, Figs. 5 and 9 show that the magnetization vector is now at $\theta_{\min}=170°$ which position is designated by B on the "Unit Magnetization Circle" of Fig. 9. A corresponding point B' is plotted on the -B-axis of Figs. 9 and 10. The seventh point of the B-H curve is thus located.

If a curve is drawn to join these seven points B', B_1', B_2', B_3', B_4', B_5', and B', we obtain a hysteresis curve which is exactly one-half of the hysteresis loop. The other half of this hysteresis loop has the same shape.

The constructed hysteresis loop shape is almost an exact duplicate of that obtained by experiment using the B-H loop tracer. Other hysteresis loops for other cases of the magnetization condition (3) were plotted using the same construction method as used in obtaining the hysteresis loop just described.

Summary and Conclusions

A "Graphical Method" was introduced for interpretation and analysis of the different magnetization phenomena in terms of the magnetic energy, and for prediction of the hysteresis loop shape of ferromagnetic thin films. By use of this method, the magnetic energy relationships are readily established.

Three major magnetization conditions were discussed: 1) condition of magnetization with a 180° field applied, 2) condition of magnetization with a 90° field applied, and 3) condition of magnetization with both the 90° and the 180° field applied. Corresponding to these magnetization conditions, the hysteresis loop shapes were predicted and constructed showing close identity to those experimentally observed. These conditions were especially chosen because of their significant importance to computer applications. Examples of these are:

Condition 1) simulates the "write mode" and the "erase mode" in the *NRZ* (non-return to zero) magnetic recording systems which include application on the existing devices such as magnetic tape, drum, disc, and core. Under modified conditions, it may simulate the "nondestructive read-out mode" in the high speed magnetic core memory.

Condition 2) simulates the "nondestructive read-out mode" in the high speed magnetic core memory. Considerable importance is attached to an investigation of nondestructive sensing by both methods, not only from the theoretical standpoint, but also from the practical standpoint of performance and design economy of a system. One reason for this is that by using magnetic thin films, it is just as easy to apply a 90° field as a

[11] Some authors use $H_{-\alpha} = H_k = 2k_1/I_s$, the anisotropy field.

[12] D. O. Smith, "Magnetic relaxation in thin films," in *Proc. AIEE Conf. on Magnetism and Magnetic Materials*, pp. 625–636; February, 1957.

180° field which is not true in the case of conventional magnetic cores.

Condition 3) is the magnetization process in cross fields. It is of interest from the viewpoint of the total energy consumption in a switching process. It is seen to require less total energy input to perform the switching operation than that required with a single field. The functional dependence of $H_{-\alpha}$ on H_β is: $H_{-\alpha}=(2k_1/I_s)\cos\theta-H_\beta\cot\theta$[10] where θ is the minimum energy point when both fields are applied. This process offers greater flexibility in choosing the value of the electrical time constant L/R. This, in turn, influences the problem of power dissipation, power needs, and input and output waveforms. This cross-field magnetization method may lead to new magnetic recording techniques for future computers.

In the "Graphical Method," the principle of superposition applies. The total free energy curve is obtained by adding up the corresponding energy components of the magnetic energy terms proper. The general expression for the total free energy equation may be written as $E_t=E_k+E_\sigma+E_{H\theta}$, in which E_t is the total free energy, E_k the anisotropy energy, E_σ the magnetostriction energy, and $E_{H\theta}$ the magnetization energy.

Since the position and direction of the magnetization vector rests where the total free energy is minimum, the total free energy curve immediately establishes a relationship of the magnetization vector with respect to the physical geometry of the magnetic material. This relationship holds for magnetic thin films which are a single domain, have the magnetization vector in the plane of the film surface, and will rotate in the plane of the film surface upon magnetization reversal.

From the total free energy curves, the important relationship of the angular displacement of the magnetization vector as a function of the applied magnetic field is established. This relationship is represented by the curve of $\partial E_t/\partial\theta=0_{\min}$. This functional relationship, with the aid of the "Unit Magnetization Circle," permits the prediction and construction of the hysteresis loop.

The "Unit Magnetization Circle" relates the position of the magnetization vector to the position of the pickup coil. It expresses the effect of the angular displacement of the magnetization vector on the output of the pickup coil as experienced by the pickup coil.

By use of the "Graphical Method," the general shape of the anisotropy energy curve, or the combined anisotropy and magnetostriction energy curve of an unknown magnetic thin film may be experimentally determined and reconstructed.[10] This curve, regardless whether it possesses a simple and analytic equation mathematically or not, can readily be used for investigation and analysis under other magnetization conditions.

Since the method of construction of the total free energy curve as well as the hysteresis loop is simple and mechanical, a quite complex magnetization condition with a multiplicity of fields of different magnitudes and directions may be simplified and handled by graphical means with ease. Thus, the "Graphical Method" should be a practical and useful tool for analysis and engineering design of magnetic devices and systems utilizing ferromagnetic thin films.

Analog Computer Measurements on Saturation Currents, Admittances and Transfer Efficiencies of Semiconductor Junction Diodes and Transistors*

A. H. FREI† AND M. J. O. STRUTT†, FELLOW, IRE

Summary—**Under the usual simplifying conditions, the solution of the diffusion equations at low current densities in junction diodes and transistors is straightforward, if a one-dimensional structure is assumed. In reality, however, such structures are three dimensional, with rotational symmetry round an axis. No useful solutions of such diffusion equations are known for these cases. An analog computer was designed, allowing for diffusion, space, and surface recombination in cases of rotational symmetry. With this computer, saturation currents of junction diodes and ac admittances and transfer efficiencies of transistors were obtained and represented by curves. Actual junction diodes were manufactured in three batches according to specifications in concordance with the computer curves. Comparing measurements of saturation currents and admittances of the three batches of diodes with computer results, yielded fair coincidence. Some unexpected features of the computer curves are that the ratios of saturation currents obtained from the analog computer and from one-dimensional solutions exceed unity and may for instance be as high as 2. Similar figures hold for ac admittances. Some choices of transistor dimensions, leading to high transfer efficiencies, are suggested by the computer curves.**

I. Introduction

We consider first the case of low current density in a one-dimensional p-n junction. The carrier densities as well as the diffusion current density may be calculated at dc and at ac. In the n-conductor outside the depletion layer, the corresponding equations are

$$p(z) = p_n + \frac{p(z=0) - p_n}{1 - \exp\left(\frac{-2l}{L_p}\right)}\left[\exp\left(\frac{-z}{L_p}\right) - \exp\left(\frac{-2l}{L_p}\right)\exp\left(\frac{z}{L_p}\right)\right]$$

$$s = (p(z=0) - p_n)\frac{eD_p}{L_p}\operatorname{cth}\left(\frac{l}{L_p}\right)$$

$$p_a(z)\exp(j\omega t)$$

$$= \frac{p_n \exp\left(\frac{eU_0}{kT}\right) eU_1 \exp(j\omega t)}{kT\left(1 - \exp\left(\frac{-2l}{L_p}\sqrt{1+j\omega\tau_p}\right)\right)}\left[\exp\left(\frac{-z}{L_p}\sqrt{1+j\omega\tau_p}\right) - \exp\left(\frac{-2l}{L_p}\sqrt{1+j\omega\tau_p}\right)\exp\left(\frac{z}{L_p}\sqrt{1+j\omega\tau_p}\right)\right]$$

$$s_a \exp(j\omega t) = \frac{e^2 D_p}{kTL_p} p_n \exp\left(\frac{eU_0}{kT}\right)\sqrt{1+j\omega\tau_p}\cdot\operatorname{cth}\left(\frac{l}{L_p}\sqrt{1+j\omega\tau_p}\right)U_1\exp(j\omega t). \quad (1)$$

We refer to the list of symbols at the end of this paper. This one-dimensional case may be approximated by a resistance-capacitance model, as shown in Fig. 1. This simple analog computer yields results which were shown to be within 2 per cent of those obtained from (1) [1].

We now consider a three-dimensional p-n junction diode of rotational symmetry. If we consider the differential equations, describing diffusion and recombination (space and surface) in the present case, it is found that no straightforward solution may be obtained which is similar to the one-dimensional case. Some of the equations pertaining to the present case are summarized in Fig. 2.

Again, an impedance network embodying all the features may be applied to the present case, shown in Fig. 2. The elements of the impedance network may be evaluated from the given differential equations, as was shown previously [2].

II. Results Obtained from the Analog Computer for the p-n-Junction Diode of Rotational Symmetry

The analog computer under consideration has been designed so as to allow variation of axial and of radial dimensions. The influence of surface recombination (see surface recombination velocity s_{rec} of Fig. 2) is represented by suitable leakage resistances in the computer.

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† Swiss Federal Institute of Technology, Zürich, Switzerland.

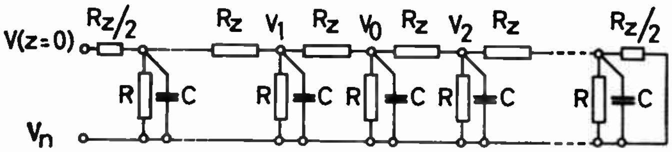


Fig. 1—Part of resistance-capacitance chain for the one-dimensional case.

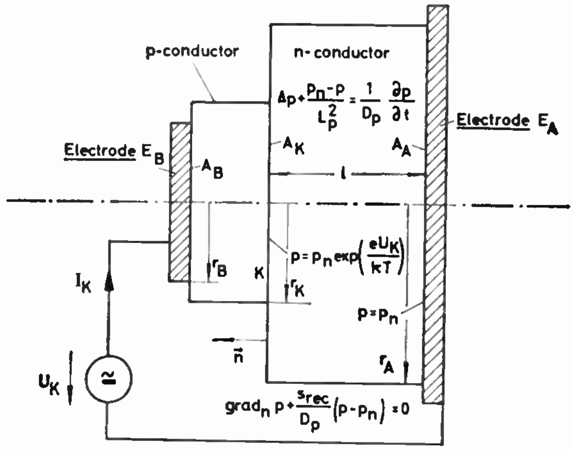


Fig. 2—Schematical representation of a p-n-junction diode of rotational symmetry round the axis. The equations are derived from the laws of diffusion and of recombination of holes in the n conductor and on its surface.

Various surface recombination velocities may be accounted for by the insertion of corresponding resistances.

Comparing current densities in the cases of rotational symmetry and of one dimension only, equal voltages are applied to the diodes in both cases. A current factor may be defined in comparing the one-dimensional and the rotational case. This factor is different for dc and for ac. Its definition is as follows:

$$k_2 = \frac{\overline{s_R}}{s} = \frac{\int_A s_R dA}{As}$$

$$k_{2a} \exp(j\varphi_0) = \frac{\overline{s_{Ra}}}{s_a} = \frac{\int_A s_{Ra} dA}{As_a}. \qquad (2)$$

Here, $\overline{s_R}$ is the mean dc current density at the border of the depletion layer in the axial direction of the rotational case. The value s is the dc current density of the one-dimensional case, according to (1). If we consider ac, we obtain the current factor $k_{2a} \exp(j\varphi_0)$. This is the ratio of the mean ac current density $\overline{s_{Ra}}$ in the rotational case to the ac current density s_a of the one-dimensional case, according to (1). Hitherto, these current factors were generally assumed to be unity. We shall prove that this assumption may in some cases be far from the truth.

Referring to the impedance network computer described previously [2], the currents between mesh points of the network are proportional to the diffusion currents in the junction diode through corresponding elements of cross section. The factor of proportionality, k_1, is given by

$$2\pi r d\, s_R = \frac{2\pi e D_p R_0 d}{k_0} i_N = k_1 i_N. \qquad (3)$$

Here, i_N is the network current, mentioned above, and s_R is the current density of the semiconductor. This (3) is valid at dc as well as at ac. In order to obtain the total diode current, (3) must be integrated over the entire cross section of the frontier between the depletion layer and the n semiconductor. This integration yields

$$k_2 = \frac{2\pi R_0 L_p d}{A} \frac{1}{\mathrm{cth}\left(\frac{l}{L_p}\right)} Y_N$$

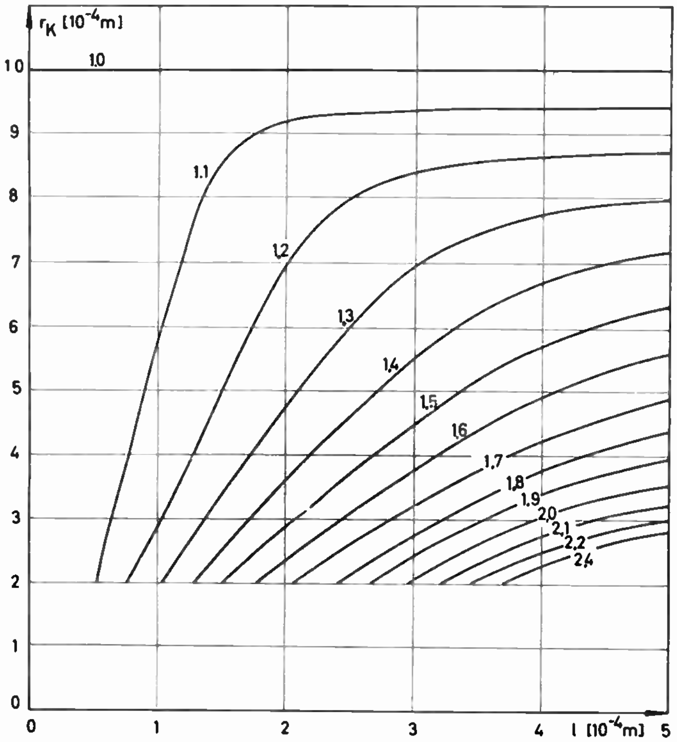
$$k_{2a} = \frac{2\pi R_0 L_p d}{A} \frac{1}{\left|\sqrt{1 + j\omega\tau_p}\right|} \cdot \frac{1}{\left|\mathrm{cth}\left(\frac{l}{L_p}\sqrt{1 + j\omega\tau_p}\right)\right|} \left| Y_{Na} \right|$$

$$\varphi_0 = \mathrm{arc}\,(Y_{Na}) - \mathrm{arc}\left(\sqrt{1 + j\omega\tau_p}\,\mathrm{cth}\left(\frac{l}{L_p}\sqrt{1 + j\omega\tau_p}\right)\right). \qquad (4)$$

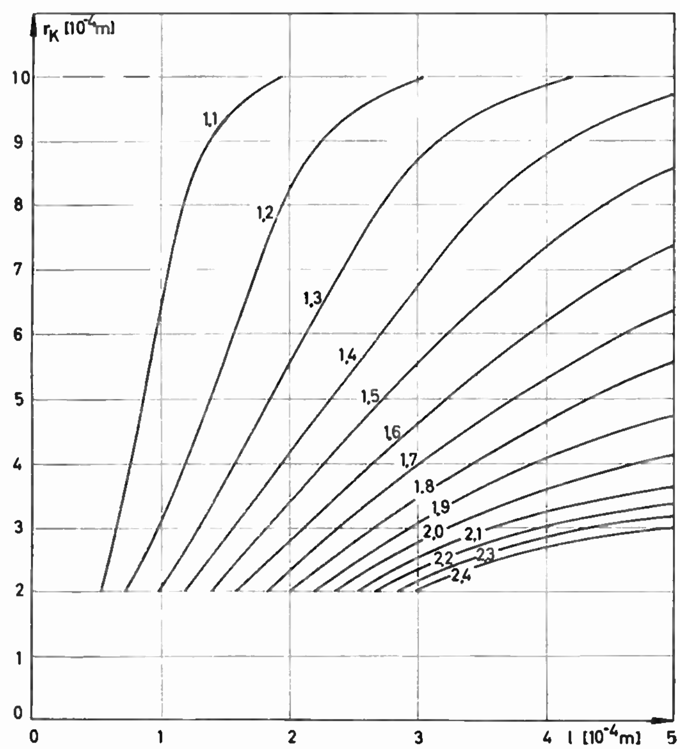
Here, Y_N is the dc input admittance of the impedance network, measured between the said frontier (see Fig. 2 at the depletion layer) and earth. All mesh points of the frontier are connected by a short-circuit lead. The diode output (see Fig. 2 at the right) is similarly represented by mesh points of the impedance network, which are interconnected by a short-circuit lead and are also short circuited to earth. The symbol Y_{Na} represents the ac input admittance of the computer under similar conditions. In order for the dc current factor k_2 to be obtained, the value of Y_N must be multiplied by a known factor [see (4)]. Likewise, in order to obtain the ac current factor k_{2a}, the value of Y_{Na} must be multiplied by another known factor [see (4)].

In the ac case, the complex admittance Y_{Na} must be measured, *i.e.*, its amount $|Y_{Na}|$ and its phase angle φ_0. This was carried out with an impedance bridge to within about 1 per cent. Several cases were considered in the computer measurements, the collector radius (at the right of Fig. 2) being constant and equal to 10^{-3} m, the emitter radius r_k, however, being variable between 2 and 10 times 10^{-4} m. The thickness of the n-type semiconductor was likewise variable between 0.5 and 5 times 10^{-4} m. The surface recombination velocity s_{rec} was assumed to be zero and 25 m/sec, respectively.

The results of the computer measurements are shown in Figs. 3–6. Each curve of Fig. 3 and Fig. 4 corresponds to a definite constant value of k_2 in Fig. 3 and to a con-

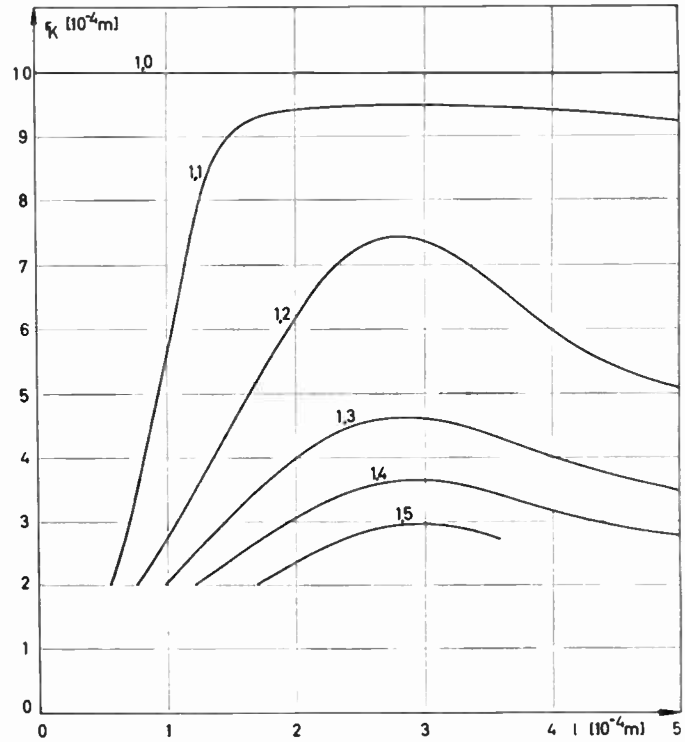


(a)

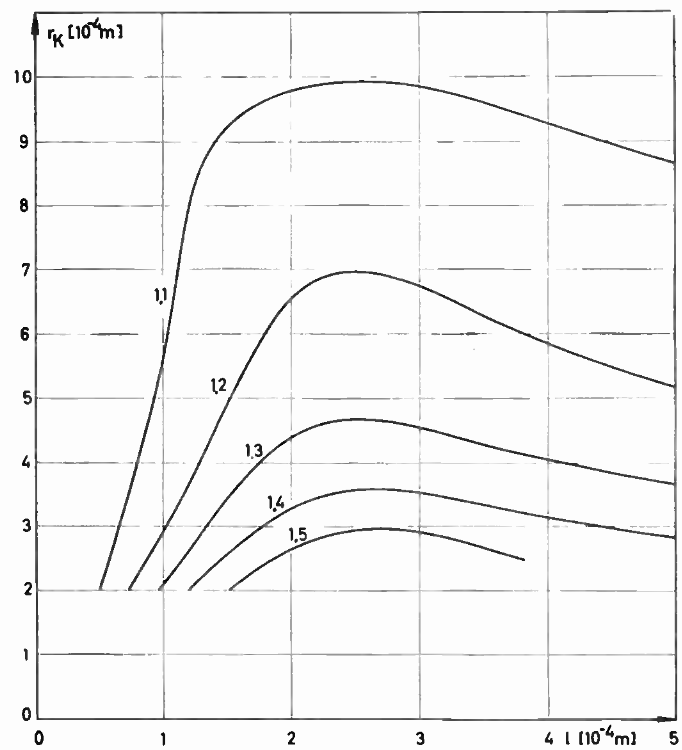


(b)

Fig. 3—Graphical representation of the dc current factor k_2 (k_2 = parameter) at variable width l of the n material in a p-n-junction diode and variable cross sectional radius r_K of the depletion layer. The collector radius r_A is constant and equal to 10^{-3} m. The surface recombination velocity s_{rec} was assumed to be zero in the case (a) and 25 m/s in the case (b).



(a)



(b)

Fig. 4—Graphical representation of the ac current factor k_{2a} (k_{2a} = parameter) at variable width l of the n material in a p-n-junction diode and variable cross sectional radius r_K of the depletion layer. The collector radius r_A is constant and equal to 10^{-3} m. The surface recombination velocity s_{rec} was assumed to be zero in the case (a) and 25 m/s in the case (b).

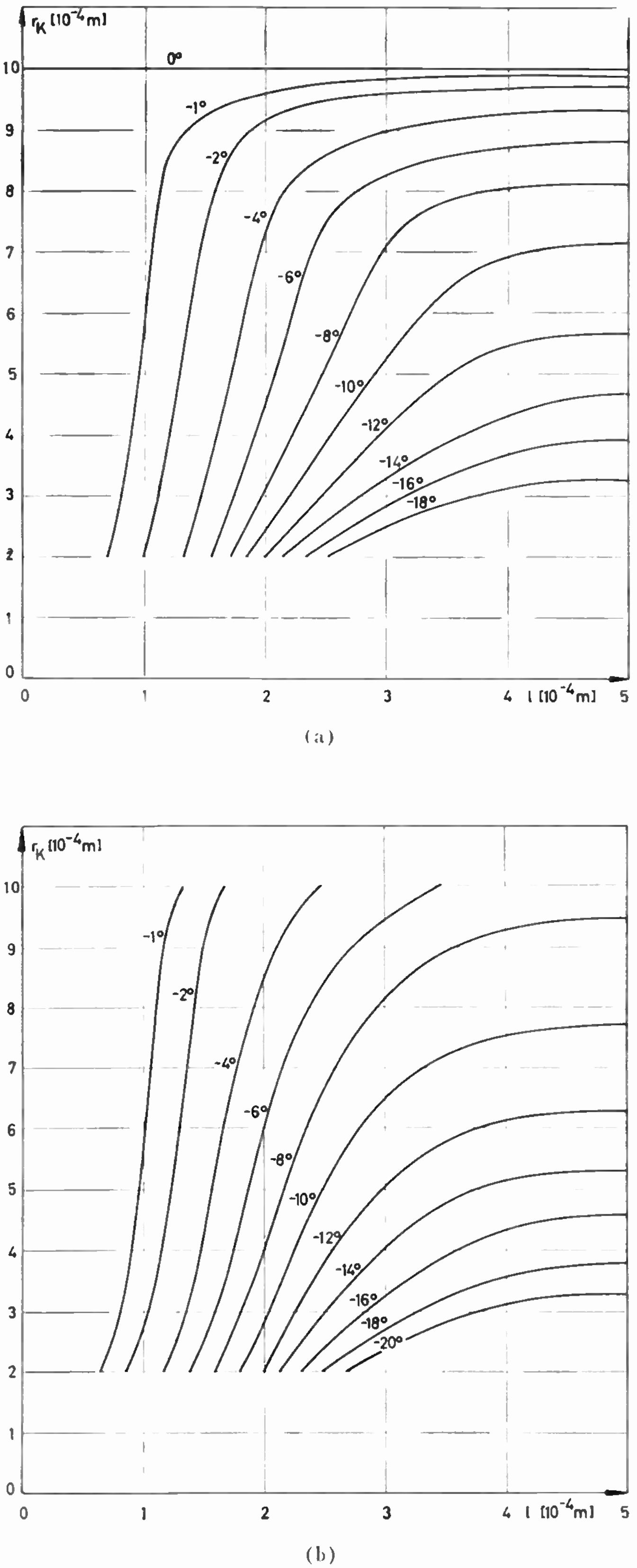


Fig. 5—Graphical representation of the phase angle φ_0 (φ_0 = parameter) at variable cross sectional radius r_K of the depletion layer. The collector radius r_A is constant and equal to 10^{-3} m. The surface recombination velocity s_{rec} was assumed to be zero in the case (a) and 25 m/s in the case (b).

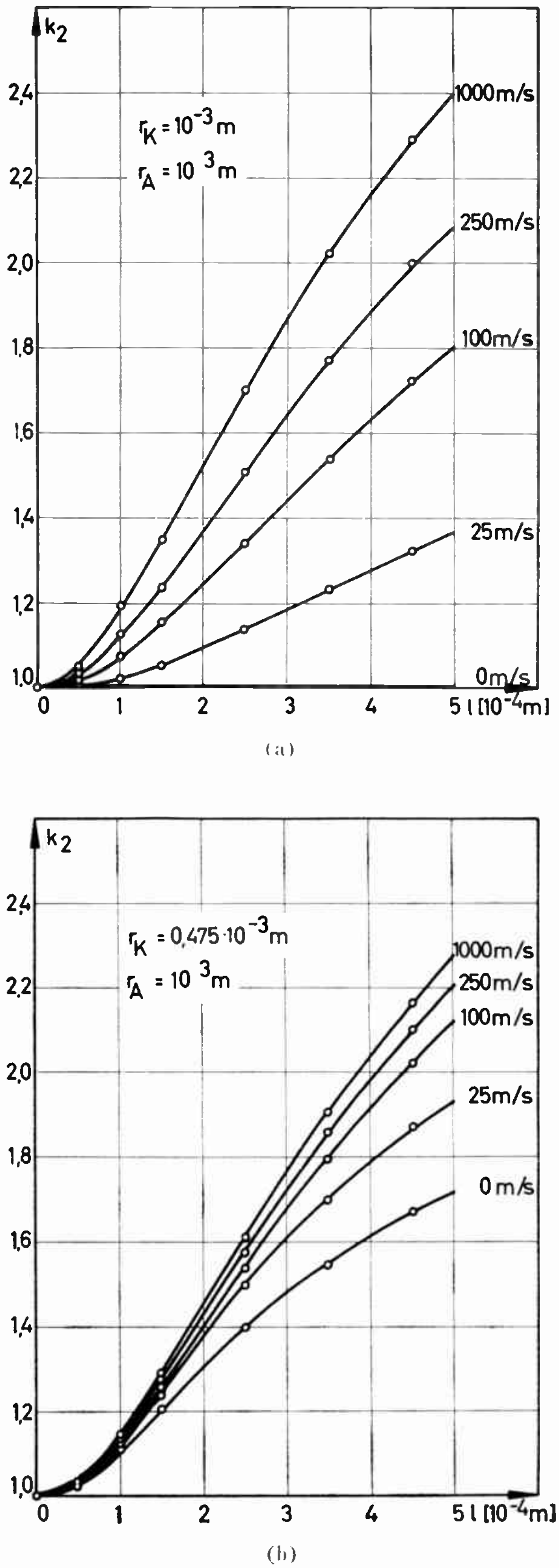


Fig. 6—Variation of the current factor k_2 at dc as dependent on the width l of the n material in a p-n-junction diode at variable surface recombination velocity s_{rec} (s_{rec} = parameter). In the case (a) we have $r_K = 10^{-3}$ m and in the case (b) we have $r_K = 0.475\ 10^{-3}$ m while r_A is constant and equal to 10^{-3} m.

stant value of k_{2a} in Fig. 4. In Fig. 5 the phase angle φ_0 is constant for each curve. The values are indicated at the curves. It may be shown that all curves of Figs. 3–5 start from the zero point at the horizontal axis. In Figs. 6(a) and 6(b), the values k_2 are shown as dependent on diode axial length l at different values of s_{rec} and at two values of r_K (emitter radius).

It is obvious that k_2 and k_{2a} cannot be smaller than unity. It is interesting and new that their values may in some cases exceed 2. Furthermore, the k_2 curves differ markedly from the k_{2a} curves; *i.e.*, the dc case differs significantly from the ac case.

III. Verification of Computer Results by Measurements of Actual Semiconductor Diodes

Three batches of germanium p-n-junction diodes were manufactured by Telefunken G.m.b.H. of Ulm, Germany, according to specifications given by the authors. A cross section of these diodes is shown in Fig. 7. The length l of the n-type germanium was given three values: 125, 275 and 525 microns, respectively, with the three batches. Eleven diodes were measured of the first batch ($l = 125\mu$). Six diodes of the second batch ($l = 275\ \mu$), and five diodes of the third batch ($525\ \mu$) were measured.

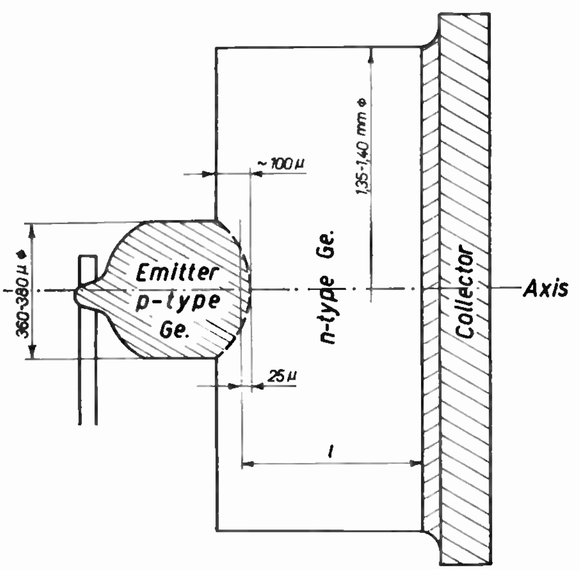


Fig. 7—Cross section of actual p-n-junction diodes manufactured by Telefunken G.m.b.H., Ulm, Germany. The length l of the n-type germanium was given three values: $l = 125\ \mu$, $275\ \mu$ and $525\ \mu$.

The physical constants of all these diodes are approximately as follows: $\tau_p = 40$ to $50\ \mu$sec, $D_p = 5.0 \cdot 10^{-3}$ m²/sec, $s_{rec} = 25$ m/sec. The measurements were carried out at a constant temperature of 273°K.

In the dc case, we have

$$I = I_s\left[\exp\left(\frac{eU_0}{kT}\right) - 1\right],$$

where I_s is the saturation current. In all the dc measurements, U_0 was given the value of 150 mv. Then, the currents I were measured for the three batches of diodes, mentioned above:

$$l = 125\ \mu;$$

$$I = 89.1, 83.2, 77.1, 83.4, 76.2, 75.0, 83.9$$

$$80.1, 72.7, 72.0, 73.5\ \mu A$$

Mean value: 78.8 μA.

$$l = 275\ \mu;$$

$$I = 59.5, 59.5, 62.0, 63.6, 70.3, 77.0\ \mu A.$$

Mean value: 65.4 μA.

$$l = 525\ \mu;$$

$$I = 65.5,\ 59.5,\ 71.5,\ 71.8,\ 75.6\ \mu A.$$

Mean value: 68.8 μA.

These mean values must be proportional to the corresponding k_2 values of Fig. 3(b), multiplied by cth (l/L_p), according to (4). We may hence compare the k_2-values of Fig. 3(b) with the above mean current values, if these are divided by cth (l/L_p).

In comparing the computer measurements with those of the actual junction diodes, it must be borne in mind that the computer model corresponds to a collector radius $r_A = 1000\ \mu$, whereas the actual junction diodes have collector radii between 675 and 700 μ, with a mean value of 688 μ. Hence, the values of emitter radius r_K and the length l corresponding to the actual junction diodes were multiplied by $1000/688 = 1.46$. The values thus obtained were used in order to derive the k_2 and k_{2a} values from the computer measurements. In the actual junction diodes the emitter radii are between 180 and 190 μ, the mean value being 185 μ.

In the computer curves of Figs. 3(b) and 4(b), the diode lengths are hence: $1.46 \times 125\ \mu = 182\ \mu$, $1.46 \times 275\ \mu = 400\ \mu$, and $1.46 \times 525\ \mu = 764\ \mu$ respectively. The value of r_K to be inserted into the computer curves is $1.46 \times 185\ \mu = 269\ \mu$. With these values and $s_{rec} = 25$ m/sec. the k_2 values of the computer curves may be found from Fig. 3(b), partly by extrapolation. The values of I, multiplied by cth (l/L_p) of the actual junction diodes are proportional to k_2. The factor of proportionality was adjusted thus, that for a length l of 400 μ the exact computer value of k_2 was obtained. Table I shows the comparison.

TABLE I

Actual length of junction diode μ	Corrected length of junction diode μ	k_2 from Fig. 3(b)	k_2' derived from current I of junction diode
125	182	1.56	1.52
275	400	2.40	2.40
525	764	3.50	3.46

The k_2 value 3.50 was obtained by extrapolation from Fig. 3(b) and may be about ±5 per cent in error. Even if this is accounted for, the above correspondence between computer measurements is striking.

A similar comparison between computer values and measurements of the actual junction diodes was also carried out for ac. The frequency of the computer measurements and of the diode measurements was 30 kc. In the diode measurements, $U_0 = 220$ mv and $U_1 = 28$ mv. In (4), the admittance $|Y_{Xa}|$ of the computer multiplied by several factors, of which only the factor

$$\mathrm{cth}\left(\frac{l}{L_p}\sqrt{1+j\omega\tau_p}\right)$$

is different for the different groups of diodes. With the actual junction diodes, the factor k_{2a} is equal to a common constant of all diodes multiplied by the diode admittance $|Y|$ and multiplied by the factor

$$\mathrm{cth}\left(\frac{l}{L_p}\sqrt{1+j\omega\tau_p}\right)$$

which varies with the different groups of diodes. The admittance $|Y|$ were obtained under similar measuring conditions to the admittances $|Y_{Xa}|$ with the computer model (see Section II). The measured admittance values of the three batches of diodes were

$$l = 125\ \mu;$$

$$|Y| = 1.63,\ 1.54,\ 1.49,\ 1.45,\ 1.64,\ 1.57,\ 1.49,\ 1.52,\ 1.57,\ 1.70\cdot 10^{-2}\ \text{mho}.$$

Mean value: $1.58\cdot 10^{-2}$ mho

$$l = 275\ \mu;$$

$$|Y| = 1.30,\ 1.45,\ 1.39,\ 1.35,\ 1.44,\ 1.60,\ 1.33\cdot 10^{-2}\ \text{mho}.$$

Mean value: $1.41\cdot 10^{-2}$ mho.

$$l = 525\ \mu;$$

$$|Y| = 1.41,\ 1.47,\ 1.47\cdot 10^{-2}\ \text{mho}.$$

Mean value $1.45\cdot 10^{-2}$ mho.

For these three batches of diodes, the mean values of $|Y|$ each multiplied by the corresponding factor

$$\mathrm{cth}\left(\frac{l}{L_p}\sqrt{1+j\omega\tau_p}\right)$$

yield the values

Length l	k_{2a} (Computer)	k_{2a}' (Diode)
125 μ	1.47	1.50
275 μ	1.46	1.46
525 μ	1.35	1.41

The second column of the table contains the k_{2a} values obtained from the computer. The values in the last column have been adjusted, so as to be equal to the second column for the length $l = 275\ \mu$. It is seen that the coincidence between the two columns is satisfactory, the largest deviation being about 4 per cent.

IV. Computation of Transfer Efficiency of an Intrinsic p-n-p Transistor of Rotational Symmetry

Transfer efficiency β_{FB} is defined as the ratio of collector current to the hole part of emitter current at short circuit conditions of the collector electrode in a grounded base connection. This transfer efficiency has different values at dc and at ac. In the case of a one dimensional intrinsic transistor we obtain

$$\beta_{FB} = \frac{-eD_p\left(\dfrac{dp}{dz}\right)_{z=l}}{-eD_p\left(\dfrac{dp}{dz}\right)_{z=0}} = \frac{1}{\mathrm{ch}\left(\dfrac{l}{L_p}\right)}. \tag{5a}$$

At ac we obtain in the one-dimensional case

$$\beta_{FB_a} = \frac{-eD_p\left(\dfrac{\partial p_a}{\partial z}\right)_{z=l}}{-eD_p\left(\dfrac{\partial p_a}{\partial z}\right)_{z=0}} = \frac{1}{\mathrm{ch}\left(\dfrac{l}{L_p}\sqrt{1+j\omega\tau_p}\right)}. \tag{5b}$$

The corresponding transfer efficiencies for an intrinsic transistor with a cross section which is of rotational symmetry may be determined by our computer. No other method of calculation has yet been successfully applied to this case. The collector electrode is short circuited and the ratio of collector current to emitter current is measured. The emitter current of the computer consists of holes only. This ratio at dc yields β_{FB} and at ac yields β_{FBa}. The results, for different dimensions and surface recombination velocities, are shown in Figs. 8 and 9. The one dimensional case is included in Fig. 8(a) (at dc) and in Fig. 9(a) (at ac) in the curves, marked $s_{rec} = 0$ (surface recombination velocity zero). These curves yield values which coincide within 2 per cent with those obtained from (5a) and (5b) respectively. For the other curves of Figs. 8 and 9, no comparison with values obtained from solutions of the diffusion equations is possible, no such solutions being known.

From Figs. 8 and 9, it is seen that the transfer efficiencies decrease at increasing base width of the transistor (l in the computer) and at increasing surface recombination s_{rec}. If the emitter electrode diameter is only half of the collector electrode diameter, transfer efficiencies are greater at large values of surface recombination velocity s_{rec} than if these diameters are equal.

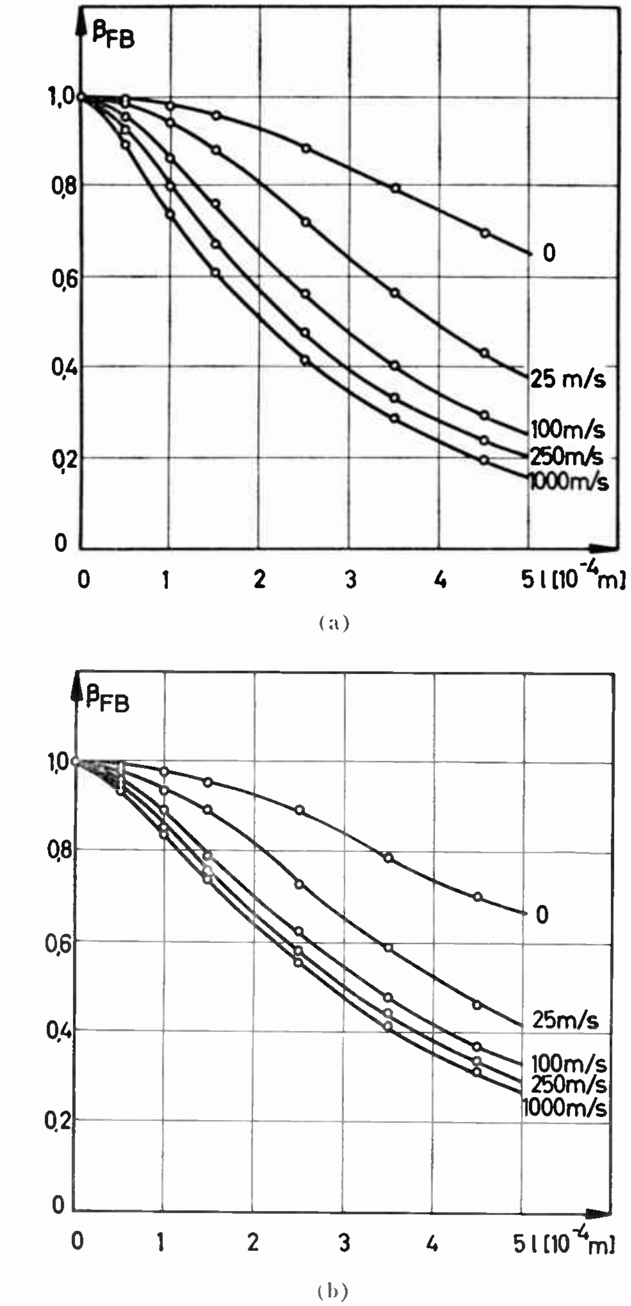


Fig. 8—Graphical representation of the transfer efficiency β_{FB} at dc in an intrinsic p-n-p transistor where l means the thickness of the n material (base layer). In the case (a) we have $r_K = 10^{-3}$ m and in the case (b) we have $r_K = 0.475\ 10^{-3}$ m while r_A is constant and equal to 10^{-3} m.

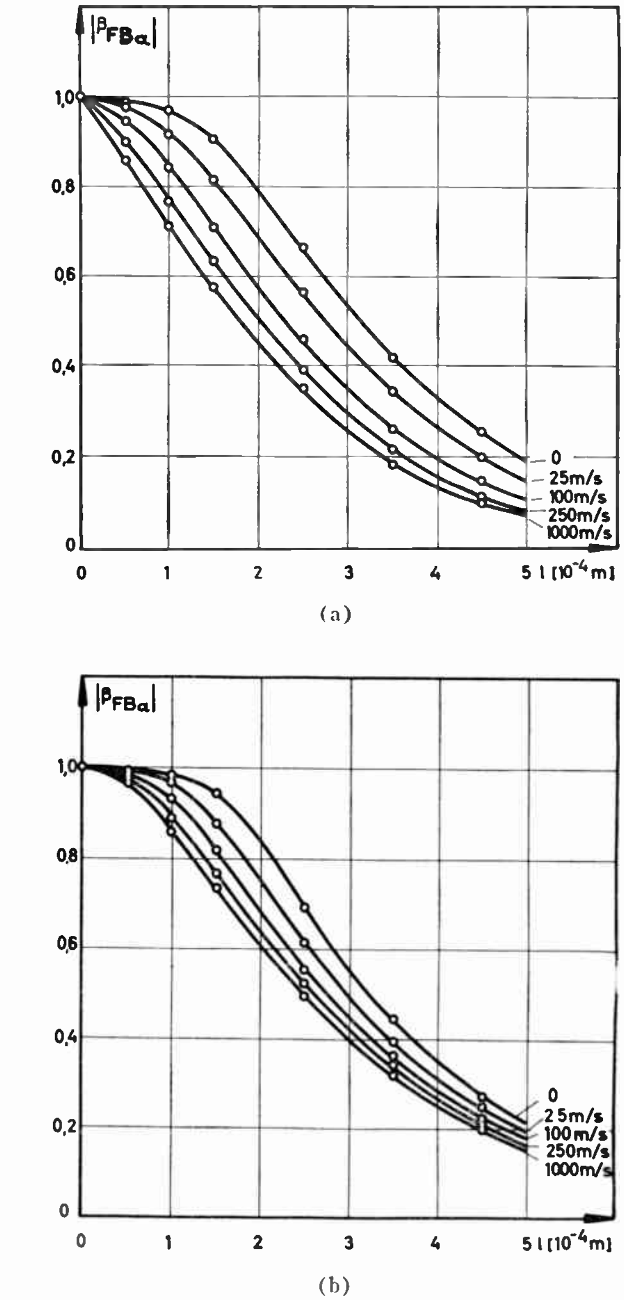


Fig. 9—Graphical representation of the amount of transfer efficiency $|\beta_{FBa}|$ at ac in an intrinsic p-n-p transistor where l means the thickness of the n material (base layer). In the case (a) we have $r_K = 10^{-3}$ m and in the case (b) we have $r_K = 0.475\ 10^{-3}$ m while r_A is constant and equal to 10^{-3} m.

V. List of Symbols

A = cross-sectional area of depletion layer.
D_p = diffusion length of holes.
e = amount of electronic charge ($1.60 \cdot 10^{-19}$ Coul.).
I = diode current.
i_N = network current.
I_s = saturation current.
k = Boltzmann's constant ($1.38 \cdot 10^{-23}$ Joule/degree Kelvin).
k_0 = ratio of network voltage to hole density.
k_1 = ratio of diffusion to network current.
k_2 = dc current factor.
$k_{2a} \exp(j\varphi_0)$ = ac current factor.
l = thickness of the n conductor.
L_p = diffusion length of holes.
n = electron density.
p = hole density.
p_a = ac hole density.
p_n = hole density at thermodynamic equilibrium.
R_0 = network resistance.
r_A = cross-sectional radius of collector.
r_K = cross-sectional radius of depletion layer.

s = dc current density in the one-dimensional case.
s_a = ac current density.
$\overline{s_R}$ = mean value of dc current density in the rotational symmetrical case.
$\overline{s_{Ra}}$ = mean value of ac current density in the rational symmetrical case.
s_{rec} = surface recombination velocity.
t = time.
T = temperature.
U_K = voltage across diode.
U_0 = dc voltage across diode.
U_1 = ac voltage across diode.
Y = admittance of diode.
Y_N = dc input admittance of the impedance network.
Y_{Na} = ac input admittance of the impedance network.
β_{FB} = forward dc transfer efficiency in a grounded base transistor.
β_{FBa} = forward ac transfer efficiency in a grounded base transistor.
τ_p = lifetime of holes in n material.
φ_0 = phase angle between mean value of ac current in the rotational symmetrical case and the one-dimensional case.
ω = angular frequency.

ACKNOWLEDGMENT

The junction diodes, used for measurements, were kindly furnished by Dr. W. Engbert of Telefunken Company, Ulm, Germany. The authors wish to express our thanks for this kind cooperation.

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CORRECTION

I. Ladany, author of "DC Characteristics of a Junction Diode," which appeared on page 589 of the April, 1959 issue of PROCEEDINGS, has requested that the following corrections be made to his paper.

Eq. (8) should read

$$E = \frac{J_\phi + qD_p\nabla p}{q\mu_p p}.$$

Eq. (9) should read

$$E = -\frac{D_p\nabla p}{\mu_\phi(p + N_d)}.$$

In the next line, replace ∇_p with ∇p.

Eq. (10) should read

$$\nabla p = -\frac{J_p}{qD_p}\left(\frac{p + N_d}{2p + N_d}\right).$$

The limits on the integral in (13) should be

$$\int_{p_0 e^{\alpha V_1}}^{p_0}.$$

Eq. (15) should read

$$e^{\alpha V_1} = -\frac{N_d}{2p_0}\Big\{\text{etc.}$$

Eq. (16) should read

$$p_0 = -\frac{N_d}{2}\left\{1 - \left[1 + \left(\frac{2n_i}{N_d}\right)^2\right]^{1/2}\right\}.$$

In the line after (17), replace $A = 2n_i/N_d$ with $a = 2n_i/N_d$.

Discussion of "A History of Some Foundations of Modern Radio-Electronic Technology"

Critique by Lloyd Espenschied*

In tracing their own inventing over the years, Hammond and Purington[1] recall interestingly the evolution of our modern art from the time of spark wireless telegraphy as of about 1910. But they give the impression that their inventing played a more foundational role than was the case. With the possible exception of Fritz Lowenstein's realization of the amplifier out of de Forest's grid audion detector, which appears to have been conceived before he became associated with Hammond, it may be said that the roots of our modern technology trace back generally to sources other than the Hammond Laboratory. Actually, Hammond's work was conducted in secret, as the authors aver, while speculative patents, filed prolifically, were long held in the Patent Office. Meanwhile the advance rolled on, with its literature accumulating, generally oblivious of the Hammond group. Hence claims made in the paper to "firsts" and to the establishment of "principles" are in need of amendment, as discussed below. The writer regrets having to turn critic, for he welcomed the rendering of the Hammond story. As it is, the paper will contribute more to technical history by calling for additional evidence. We follow the sections of the paper beginning with Section II.

The Radiodynamic Torpedo (Section II)

Hammond is best known for his long pursuit of the problem of directing torpedoes by radio control. As still earlier pioneers in this field, mention is made in the paper of a pair of British inventors and of N. Tesla. But singularly omitted from mention is the one who received from the U. S. Patent Office the underlying claims, one of which reads:

> The combination with a source of electrical waves or disturbances of a moving vessel or vehicle and mechanism thereon for steering or operating the same, and controlling apparatus adapted to be actuated by the influence of the said waves or disturbances at a distance from the source, as set forth.

This is Patent No. 660,155, one of a pair, means and method patents, granted to Lieutenant (later Admiral) Bradley A. Fiske, October 23, 1900. Fiske had filed only a few months after Tesla in 1898 and was able to swear back of Tesla and obtain the primary claims. In 1912 Hammond wrote from Washington, D. C., a letter[2] inquiring about this patent, saying "I am very much interested in Patent No. 660,155 . . . ." and going on to observe: "This is the first record of any kind which completely covers the art of wireless control of mechanisms." Thus, one wonders why Fiske is not cited as a pioneer, or *the* pioneer, in the present paper. In the summer of 1915 Hammond was selling to a Congress worried by the European war then raging, his system of wireless control of torpedoes. Fiske became concerned lest his patents be infringed and issued a public statement explaining how he came to make the invention, and observing that, "Mr. John Hays Hammond, Jr. has done splendid work based on my original patent."[3] In his autobiography of 1919 Fiske wrote, ". . . I do not know of Mr. Hammond ever giving me credit for having suggested the plan originally, or of his disclaiming the credit given him for it in many accounts of his achievements."[4]

Of course it was quite impossible for any of the early inventors to get very far on this problem, so crude was early radio; and even after the revolutionary high-vacuum tube came along, to reach an underwater vessel with control signals was most difficult. Two World Wars have now occurred with no military use of the radio-controlled torpedo; it is just as well, except for the futile expenditure of technical effort and public money.

Among the "principles" claimed to have been developed in the 1910–1914 period is that of the automatic stabilization of the course of a torpedo by means of the gyro.[5] Yet one reads in the *Encyclopaedia Britannica* of 1910–1911, eleventh edition, under the article on "Torpedo," of gyrocontrol trials in 1896 which "demonstrated the feasibility of accurately and automatically steering a torpedo in a direct line by this means," and of the Whitehead firm having "produced the apparatus which is fitted to every torpedo made."

Under the subheading "Automatic Course Stabilization" it is said, ". . . in 1912, the Sperry Gyroscope Company and the U. S. Navy were developing a precise and reliable motor-driven gyrocompass with remote repeaters." This development is understood to have been undertaken by Sperry alone, the Navy adopting the system upon its appearance in 1911. The paper continues: "The Hammond Laboratory engineered the modification of one of these devices so that the repeater controlled not a compass indicator, but the operation of a steering engine." This modification was rather obvious since the idea of steering a ship automatically from a magnetic compass was old. This connecting of a Sperry gyro to a steering gear instead of to a compass indicator is referred to as the "automatic pilot principle," as if the automatic pilot originated with Hammond.

This section of the paper, under the previously mentioned subheading "Automatic Course Stabilization," is made more confusing by the inclusion in it of a reference to three Hammond patents,[6] as if these patents cover course stabilization and the automatic pilot. They do not; they apply only to the application thereto of radio control. In fact, the arrangement shown for combining gyro and steering engine is unserviceably crude—the control is intermittent rather than continuous, through declutching, and there is no means for restoring the rudder to normal after a turn, *i.e.*, repeat back, so necessary for an operable automatic pilot. Thus, the three patents added nothing to automatic course stabilization *per se* and their inclusion at this point in the text is misleading.

The man who pioneered and produced the automatic pilot was Elmer A. Sperry. His early progress is described in a comprehensive paper of 1913. "Perhaps the most interesting of all the apparatus which we have developed is the aeroplane stabilizer, . . . " the section on this subject begins. It was applied to a Curtiss Hydroplane, and is pictured in Figs. 28 and 29.[7]

Under the subheading "Security of Radio Control,"[8] Tesla is credited with having "proposed a security system based upon the coincidental transmission on two channels: a forerunner of the 'and' principle of modern computers." But ahead of Tesla, apparently, were the two British inventors, Wilson and Evans, mentioned earlier in the paper. Their U. S. Patent No. 663,400 of 1898–1900, shows coincidental transmission over two channels. Two short-wave Hertzian dipoles disposed mutually perpendicular with a similar pair for receiving, provide two channels by polarization. The receiving control electromagnet is made dependent for its operation upon the receipt of both channels, whereby it appears Tesla was limited to two different frequencies.

A so-called Hammond system which uses in a conjoint manner the familiar marking and spacing waves of the Poulsen arc transmitter is described. Later in the paper this is made the basis of claims to FM. But de Forest, then of the Federal Telegraph Company, who installed Hammond's arc equipment in 1913, already had devised what he called ". . . a diplex system of telegraphy, using a high-speed chopper." It "automatically changed the wavelength of the transmitter from A to B a great number of times per second, so that both the A and B operators could dispatch their messages. . . ."[9] Hence the Hammond claim to ". . . the first example of security systems using both time and frequency diversity" appears to have been an adaptation to a slow "security" operation of de Forest's diplex.

* Received by the IRE, December 2, 1957.

[1] Proc. IRE, vol. 45, pp. 1191–1208; September, 1957.

[2] Letter of January 11, 1912, written to the Western Electric Co., 463 West St., New York, N. Y.

[3] *The New York Times*, p. 11, col. 3; July 24, 1915.

[4] B. A. Fiske, "From Midshipman to Rear-Admiral," The Century Co., New York, N. Y., p. 232; 1919.

[5] Footnote 1, p. 1192.

[6] Footnote 1, p. 1192, reference 7.

[7] E. A. Sperry, "Engineering applications of the gyroscope," *J. Franklin Inst.*, vol. 175, pp. 447–482; May, 1913.

[8] Footnote 1, p. 1193.

[9] L. de Forest, "Father of Radio: The Autobiography of Lee de Forest," Wilcox and Follett Co., Chicago, Ill., p. 277; 1950.

Toward the end of Section II of the paper,[10] there is quoted a claim from a Hammond patent of 1932–1936[11] which is said to be "the statement of the Proximity Fuse principle." This is erroneous, for the claim is for a *torpedo*, meaning in water, from which the energy radiated must be compressional or sound waves, the only kind disclosed in the patent, whereas the proximity fuse employs radio waves.

The Triode Tube (Section III)

The technological revolution that has resulted from the electron tube requires that reports of its onset be rendered correctly. The report given in the paper is from the standpoint of the Hammond group, and while this is a welcome contribution, it is by nature limited and one-sided, and contains some errors. Hence, additional evidence is offered as follows.

The second paragraph quotes our old friend Robert Marriott as having written that de Forest's Audion "was used to some extent as early as 1906 . . . ". The quotation is correct, but the assumption that it referred to the triode rather than the diode, as of 1906, is in error. It was the two-element tube that was publicly christened that year with the name Audion in de Forest's AIEE paper of October 20, 1906, devoted to the diode and entitled simply, "The Audion." Incidentally, Pupin amusingly said of the name: "It is a mongrel. It is a Latin word with a Greek ending!" And he expressed dissatisfaction with de Forest's inability to explain its *modus operandi*.[12] In his recent autobiography, de Forest refers to his detector tube as "a carbon filament surrounded by a platinum plate,"[13] which he used in receiving at 42 Broadway, New York, in 1906.

The grid triode appears to have been invented toward the very end of that year or the beginning of 1907. It was filed upon January 29, 1907 and issued as a patent February 18, 1907, No. 879,532, entitled "Improvement in Oscillation Detectors." (There was no time lost in those days!) Just ahead of it de Forest had devised another form of triode, one in which the control electrode was a second plate located on the side of the filament opposite to the anode. He sought to make of it a telephone amplifier; but being unable to obtain amplification, contented himself with filing a speculative patent application October 25, 1906. It was issued January 15, 1907, as No. 841,387, entitled "Device for Amplifying Feeble Electrical Currents." Interestingly enough, the patent drawing shows, connected in series with the input control electrode, a condenser such as became familiar in the grid audion detector. But whereas in the grid tube it enhanced the detector action, it could only do harm in the amplifier, causing it to block. This probably was one reason for the failure of the two-plate amplifier, the other being the lesser electrostatic control of the filament-anode electron path. So, it fell to the grid form of tube to lead the way.

The earliest public disclosure known to have been made of the grid audion detector was in a lecture de Forest gave on March 14, 1907, on "The Wireless Transmission of Intelligence," before the Brooklyn Institute of Arts and Sciences. It was about that time that the grid audion began to come into some use, "but apparently in very small numbers," as Marriott said. Among those small users, to the writer's knowledge, were a few amateurs who learned where the bulbs could be procured, managed to save up the required $5, and then suffered the filament burn-out soon to follow! The writer, with two other amateurs, Austen M. Curtis and Francis A. Hart, had attended de Forest's Brooklyn lecture and there made the acquaintance of "Doc" and his young assistant John V. L. Hogan. The grid audion detector continued for some years solely as a detector without getting very far, as the authors of the paper indicate, for it was relatively expensive and only moderately more sensitive than the simple crystal-contact detector.

Hammond and Purington are to be congratulated for having at long last pulled aside the curtain of secrecy and revealed Fritz Lowenstein's contribution in making de Forest's grid audion detector into an amplifier and also an oscillator. But they do not tell the whole story and give the impression of Hammond's role having been more than it was. It was not actually "as a consultant of the Hammond Laboratory," that "Lowenstein on May 11, 1911, undertook in New York the development of the three element 'ion controller' . . . ". The date is that of a letter-contract[14] which defined the Hammond-Lowenstein relations, addressed by Hammond to Lowenstein and signed by both, which runs as follows:

> May 11, 1911.
>
> Frederick Lowenstein, Esq.,
> 115 Nassau Street,
> New York City.
>
> Dear Mr. Lowenstein:
> I am writing you a letter to confirm the points taken up in our conversation this morning at the Hotel Belmont, with Mr. John Hays Hammond and Mr. George Clark. These points, as I recollect, are as follows:
> You are to work and develop two separate inventions—one, my own automatic wireless selective system, and the other, your controller.
> On account of the development of my wireless selective system, I agree to advance to you the sum of $1,000.00 for your personal supervision, advice and services in the designing and construction of my invention. . . .
> In the second proposition, to develop your controller, I shall advance you the sum of $1,000.00 for the purpose of constructing such apparatus as may be necessary for experimentation, with a view to perfecting your controller. . . .

(The term "controller," or "ion controller," was Fritz's alias for the name "Audion" which de Forest had bestowed upon Fleming's tube upon adding to it the "B" battery.) The rest of the agreement gives Hammond an option to take a 50 per cent interest in the controller invention (which is understood not to have been exercised).

From the above it is evident that Lowenstein was a consultant to Hammond for the latter's selective radio system for torpedo control, but not in respect to his own ion controller. Hammond's relation to Lowenstein in respect to the ion controller was that of financial backer, or business partner. This is borne out by the fact that in applying for a patent Lowenstein used his own patent attorney; and when he later sold the patent to the American Telephone and Telegraph Company the payment, of the famous $150,000, was to him. Fritz had tried the grid audion as an amplifier before becoming associated with Hammond. This is indicated in his letter to Hammond of September 19, 1911, quoted in part in the paper, wherein Lowenstein speaks of his "efforts on reproducing the telephone tests of last winter . . . ,"[15] the winter of 1909–1910. What these tests were, fortunately, has been told by Fritz himself in another document, in an affidavit submitted to the Patent Examiner in the course of prosecuting his patent application—that which became the famous negative grid, or "C" battery, Patent No. 1,231,764 of 1912–1917. The affidavit, sworn to on September 8, 1915, is to be found in the Patent Office record of that patent. It describes how interest was first aroused by an experience in 1906 wherein Lowenstein observed the sensitivity to electrostatic influence of a Cooper-Hewitt mercury arc rectifier. It then goes on to say:

> In the winter of 1909–1910 I did some work as a consulting expert to the President of the Radio Telephone Company of Newark, N. J., and in their shop and at that time I constructed and assembled an operative telephone system of which the receiving end was arranged substantially as shown in the drawings of my above application, except that the grid 18 was simply connected to one side of the filament 16, *i.e.*, at the same potential as the filament.

Accompanying the affidavit were Exhibits A, B, and C. The latter, a reproduction of the patent drawing, shows a grid Audion provided with audio input and output transformers connected as an amplifier in the receiving leg of a standard telephone subscriber instrument.

The affidavit continues:

> In 1911 I did further work on the invention in the way of perfecting details thereof.

(*i.e.*, when backed by Hammond). The affidavit then reveals how Fritz came to substitute for the grid condenser of the audion detector the negative "C" battery which made of it a class A amplifier, which negative grid condition proved to be the invention of the patent.

> The apparatus employed in the 1911 tests were constructed with the aid of Benjamin T. Miessner (now at Purdue, Indiana), and George H. Scherff, of New York City.
> In a talk in 1911 with Louis Engelhorn at the Chicago end of the line, and Harris Hammond and myself at the New York end, we employed a condenser in the grid circuit and found that the talk over the line was poor, the current through the controller tending to choke or stop. I found that this could be remedied readily by touching the grid binding post repeatedly to discharge the grid and open up the talk. This gave me the idea of connecting the grid to a point ultranegative in potential relative to the filament. Such a connection was actually made in the latter part of 1911 and was found to be a very substantial improvement.

These, then, were the steps by which Lowenstein converted de Forest's grid audion detector to the magic amplifier it became:

[10] Footnote 1, p. 1197.
[11] Footnote 1, p. 1197, reference 21.
[12] L. de Forest, "The audion," *Proc. AIEE*, vol. 25, pp. 719–747; October 20, 1906. Discussion, pp. 863–873.
[13] de Forest, *op. cit.*, p. 210.
[14] This letter-contract has kindly been made available to the writer by the authors. It was part of a report by E. S. Purington, "Early History of Selective Receiving Systems with Special Reference to . . . Cases of RCA vs Splitdorf Co., and RCA vs George H. Walker Co.," p. 22; February 19, 1938.
[15] Footnote 1, p. 1198.

1) Initial interest in the possibility of some form of vacuum tube amplifier, aroused through contact with Peter Cooper-Hewitt's mercury arc rectifier tube (a form of which Hewitt himself was developing as an amplifier).
2) Familiarity with the Audion grid detector, gained while working as a consultant for de Forest's old Radio Telephone Company, soon defaulted, 1909–1910, which led Fritz to try to make of the audion an audio-frequency amplifier. He seems to have had some success, but evidently kept it to himself at the time; his obligation to the Company is not known.
3) Renewal of the audion amplifier tests in the fall of 1911 with Hammond's backing, resulting in amplified reception of long-distance telephone calls and in the attainment of a true Class A amplifier by means of the negative grid discovery.

Therefore, it is seen that Lowenstein was on the trail of the audion amplifier before becoming associated with Hammond and carrying out the "further work on the invention in the way of perfecting the details thereof." Under these circumstances the paper is misleading in implying that he started the development on May 11, 1911 and in the capacity of consultant to the Hammond Laboratory.

The Oscillating Audion

It is well established, as the authors say, that Lowenstein had the audion amplifier working as an oscillator toward the end of 1911, and for both audio and radio frequencies. The strange thing is that it was not followed up, the more so in view of Hammond's high hopes for it as expressed in his letter to Beach Thompson of January 25, 1912, given below. And it is singular that to Hammond, "The exact nature of the ionic devices and the manner of operation are not known." One wonders too that the evidence of Lowenstein's having the audion as an oscillation generator during the winter of 1911–1912 was not presented to the courts in the long de Forest-Armstrong litigation over the oscillating tube. Such evidence would have demonstrated the natural tendency of an amplifier to oscillate and hence how little invention there was in the oscillating audion *per se* once it had become an amplifier. As matters were allowed to go, it fell to others to give to the world the oscillating audion in useful form, most notably, perhaps in terms of early date and application of control of frequency, to Alexander Meissner of Telefunken[16] using a gaseous tube of the von Lieben Triode.

An Important Tip

In view of its probable importance for de Forest, then working at Palo Alto, Calif., for the Federal Telegraph Company, the letter which Hammond wrote the President of that company, and from which the Hammong-Purington paper quotes in part, deserves to appear in full:

January 25, 1911.
(Typist's error, should be 1912)

Beach Thompson, Esq.,
President, Federal Telegraph Co.,
Merchants Exchange Building,
San Francisco, California.

Dear Sir:

I have recently been informed about the wireless work you have been doing on the Pacific Coast, and I am much interested in the results that you have obtained. If you have any descriptive matter I would appreciate your sending it to me.

I have been experimenting with a new form of apparatus designed to produce undamped and high frequency oscillations. Our method is far more reliable and simpler than the Poulsen arc method, or the high frequency alternator system as used by Fessenden and others. We are in process of developing this apparatus, and when it has reached a practical point, I would be very glad to send you more complete information regarding it.

In the experiments we have found that our method is highly suitable for wireless telephony, as there is absolutely no sound produced whatsoever as in the arc or hf alternator. I believe that telegraphy by means of undamped high frequency oscillations is the logical future of this art, for the reasons which you have already discovered and proven: that far better tuning may be obtained where there is no decrement to the wave train, and that there is less absorption of energy in long distance transmission, and also the important fact that low voltages are used at the transmitting station. However, the chief weakness in all systems of this kind is essentially in the means of producing the continuous high frequency oscillations.

I am quite familiar with the art in Europe and during my recent trip to Germany found that most of the companies had abandoned the arc method of oscillation production. It is for this reason that I believe there is quite a future in the development of the work which we are carrying on.

Hoping that you will be kind enough to send me any data which you may be willing to disclose, believe me

Yours very truly,

JHHJr/T

On the same day Hammond wrote to an acquaintance of Thompson, and since this letter, too, is evidence of a stimulative tip given the Pacific Coast people, it is presented:

January 25, 1912

Major F. R. Burnham
Llankershim Building
Los Angeles, California.

Dear Major:

I want to thank you very much for your letter in regard to wireless development in California. It certainly is very interesting to me as the reports I have had of this enterprise have been very meagre. I will herewith write Beach Thompson, informing him that the iron controller which you saw in my experimental laboratory here has recently shown some remarkable results in experiments in which we produced high frequency undamped oscillations with far better results than have ever been obtained by the Poulsen method.

I hope some day that I will be able to introduce a wireless system in Sonora. After the Yaqui has developed enough they will need a station there, and we could furnish a good one.

Very sincerely yours,

JHHJr/T

(The term "iron controller" is a stenographic slip for "ion controller.")

Photo copy of each of the above letters (from Hammond's file carbon copy) were kindly given the present writer some years back by Hammond and Purington. Appreciating the probable bearing of them upon de Forest's making the audion into an amplifier at Palo Alto in the summer of 1912, a copy of the letters was sent to him, with the question of whether they may not account for Beach Thompson having asked him to undertake the further development of his audion. Dr. de Forest's response was: "I do not recall that Beach Thompson ever mentioned his correspondence with Hammond, but very likely that is what induced Thompson to urge me to develop the audion as an amplifier."[17] Whereby we see that Lowenstein, having been indebted to de Forest in the first place for the vital grid tube, repaid his obligation through Hammond's letter, perhaps unwittingly!

Bell System Side of the Picture

An insight into what was happening within the Bell System at the time of the Hammond-Lowenstein approach may be welcome: the American Telephone and Telegraph Company and associated Western Electric Company already had started tackling the telephone repeater problem from the standpoint of a vacuum tube element of some kind. From the University of Chicago laboratory of Prof. Robert A. Millikan a young graduate student, Dr. H. D. Arnold, had been recruited. He started to work at the beginning of 1911 and took up the kind of vacuum tube that seemed likely to carry the load of a telephone repeater, the mercury-arc tube similar to that of Peter Cooper-Hewitt. He had succeeded in getting amplification when our present story opens.

On January 27, 1912, Hammond and Lowenstein took to the office of transmission engineer F. B. Jewett a sealed box and demonstrated it as a prospective telephone repeater. Further contact was made in June, and there were additional telephone conversations, but not until January 12, 1913, while Lowenstein was in Europe, was it fully disclosed what was in the "Black Box."

Meanwhile de Forest himself had succeeded in making his audion into an amplifier on the Pacific Coast, in the summer of 1912, and through John Stone-Stone approached the American Telephone and Telegraph Company. De Forest came east and on October 30 and 31 demonstrated the audion as an audio amplifier to Dr. Jewett and E. H. Colpitts in the laboratory of the Western Electric Company, at 463 West St., New York. He left the device, and the following day, November 1, Colpitts called in Arnold and showed it to him. Arnold later testified frankly as to his impression: " . . . when I went into the room and saw this thing and saw how it worked I was very much astonished and somewhat chagrined because I had overlooked the wonderful possibilities of that third electrode operation, the grid operation in the audion. . . . I knew about the de Forest audion in print, but I was wrong in my impression of what the de Forest audion might do because I had not realized what the grid would do in such a device."[18]

Thereupon Arnold turned from his own mercury arc tube to embrace the audion. As demonstrated by de Forest it was a remarkably sensitive but weak and unreliable amplifier. It would amplify only at low speech levels, about 30 db down; at normal levels it would block and produce noise, for the audion still had in it the grid condenser of its radio detecting days! This trouble was soon overcome and Arnold then addressed himself to the improvement of the tube itself. He had recognized from the beginning

[16] A. Meissner, "The development of tube transmitters by the Telefunken Company," PROC. IRE, vol. 10, pp. 3–23; February, 1922.

[17] Letter to Dr. L. de Forest to L. Espenschied, September 15, 1953.

[18] Arnold-Langmuir high-vacuum tube litigation. Testimony of 1926 in the District Court of the U. S., District of Delaware, No. 589 and 598. In Equity. General Electric Co., Plaintiff, vs The de Forest Radio Co., Defendant. Vol. I of six volumes, pp. 554–556.

that the "blue haze" was due to gas and sought better evacuation. The best of the tubes de Forest had left with Western were repeatedly pushed to the highest plate voltages they would stand to clean up the gas in them. Other tubes were obtained from the manufacturer, McCandless; the electrical characteristics were measured; and by the end of the year Arnold and his assistants had a fair mastery of the newcomer. There was ordered from Germany the latest type of vacuum pump (Gaede, molecular); and upon its arrival in April, 1913, Arnold started making his own tubes of high vacuum and greatly improved design.

Thus, the Western Electric Company engineers were already familiar with the audion when in January, 1913 Lowenstein's "black box" was opened to them. They had expected to find something new, and were disappointed when there appeared only "the ordinary audion" as they reported. The negative grid was noted and Colpitts in his report mentioned that the question of its potential was to be studied and the company should be free to employ any polarization. This point must have seemed minor to his boss Jewett, for he lost sight of it. An examination of the patent papers submitted by Hammond revealed that the three claims stood rejected by the Patent Examiner on two de Forest patents. The work of the Austro-German von Lieben group was then coming to attention, on which the Lowenstein claims were soon to be rejected by the Patent Office. Altogether, the American Telephone and Telegraph Company's patent attorney Lockwood reported, "I do not see that the Lowenstein people really have anything to sell," and since they were unable to demonstrate to the contrary, the case was dropped. But Lowenstein's patent attorney, M. C. Massie, was persistent and resourceful. A few years later he managed to get allowed claims on the negative grid. Imagine the surprise of the telephone patent people when the patent was issued in 1917 with claims that read on the grid polarization that the engineers had found to be necessary and that was then in use! So, the patent was bought, quietly, for the very considerable sum of $150,000. Massie told the writer afterward that the company originally could have had the patent for $20,000 and he had finally boosted the price to $200,000, to come down to the amount agreed upon!

Thereby hangs a supplementary tale too good not to note. The writer had known Lowenstein as a fellow IREer; had been at the American Telephone and Telegraph Company's headquarters on radio matters during all this time, and had been aware of company contacts with Fritz, but without knowing the subject matter. Imagine then, his surprise upon meeting Lowenstein one lunch time in 1918 in the German restaurant beneath the Woolworth Building, and being shown a check to Fritz's order from the American Telephone and Telegraph Company for the $150,000! To the question of what was it all about, he cheerfully explained, and then observed in his quizzical way, "And to think, for just a little dry battery!" The appropriate rejoiner would have been, "But in the right place!" Fritz was so pleased with his accomplishment that he carried the exhibit, a photostat, around with him to show his friends, as an Indian would a scalp! Had the writer not known Fritz personally he would have been unaware of the telephone company's payment, for so unheralded had it been at headquarters! Incidentally, Fritz's insertion of the audion amplifier in the receiving branch of his office telephone in 1911 constituted the first application of that revolutionary device to the Bell System, sub rosa as it was!

The von Lieben Group

Candor requires that we recall at this point something of another group of inventors whose work with the cathode-ray tube was contemporary with that in the U. S. and had been directed from the beginning at the problem of the telephone repeater. The Austrian Robert von Lieben patented in Germany in 1906[19] a telephone repeater comprising a thermionic three-element beam-deflection tube intended to be of high vacuum. It was scientifically sound but not successful. The inventor, with two associates, Reisz and Strauss, fell back upon the gaseous, ionic, type of thermionic tube beginning about 1909. In 1910 several forms were patented in Germany.[20] One of these used an intervening grid for control and was applied for in the U. S. on January 30, 1911, and issued September 17, 1912, No. 1,038,901. It was this patent as well as earlier de Forest audion patents, that was cited against Lowenstein. Thereby he was prevented from covering the audion as an amplifier broadly, his invention reducing to the negative grid feature.

While it seems clear that Lowenstein started on his amplifier quest from de Forest's audion in 1909–1910, his more successful renewed pursuit of it in 1911 may well have been stimulated by some knowledge of the von Lieben work having seeped across the Atlantic. He is known to have followed the German-language literature, and Hammond's visit to Germany in the summer of 1911 had made Hammond "quite familiar with the art in Europe" as he wrote Thompson. Since Lowenstein did not get the blocking condenser out of the grid circuit until late in 1911 (by which time the von Lieben group already were in the U. S. Patent Office), he cannot be credited with having been the first to have arrived at the grid thermionic amplifier even in the U. S. But he *was* the first with the type of device that was to win out. The initial promise of the von Lieben mercury-vapor ionic grid tube proved to be chimerical, for the device was erratic and noisy and did not lead to the final answer, the high-vacuum tube. Fortunately for de Forest and Lowenstein, the audion did! Actually, Lowenstein and Hammond, by coming into the picture when they did, alerting de Forest, and leading on to the two large electrical companies, performed a major service.

High-Vacuum Tube

Singularly enough, as the paper indicates, both of these big companies, the American Telephone and Telegraph Company (and its associates of the Bell system), and the General Electric Company, learned of the audion about the same time, late October and early November, 1912. Each went immediately about the improvement of the mysterious little tube, unknown to the other. Although up to that point the mode of operation had been a mystery, and the tube was flimsy, erratic (hardly any two were quite alike), and unable to carry a material load, within a year or so each company had mastered the device and was producing a tube characterized by regularity and reproducibility, whereby it became the revolutionary electronic tool it did. The contribution of these companies was, then, a major one. It followed from the application of the scientific knowledge of the time to the temperamental little tube, first in diagnosing its troubles, and then in applying the remedies: high vacuum and higher voltage, better filament emission and longer life, and circuits properly designed to go with it. No wonder these two great companies locked horns over the great high-vacuum improvement when the General Electric Company sought and obtained a patent on it on behalf of Langmuir. Arnold more modestly had regarded his high-vacuum tube as an application of scientific knowledge rather than invention. The contest continued many years, with the Supreme Court finally sustaining Arnold's view. He was given credit for having anticipated Langmuir in November, 1912, and de Forest was recognized to have gone part way toward the high-vacuum tube even earlier without knowing the physics of it. The Langmuir patent was invalidated on the basis that, ". . . the relationship of the degree of vacuum within the tube, to ionization, and hence to the stability and effectiveness of the discharge passing from cathode to anode, was known to the art when Langmuir began his experiments." The reference cited is a paper by Lilienfield.[21] It is pointed to as having ". . . made a complete and explicit disclosure of the essentials . . .". The decision went on to recognize that, "Lilienfeld also deduced from meter readings and stated the 3/2 power relation of current to voltage, as Langmuir later stated it in his patent. From this the conclusion is inescapable that Lilienfeld knew and stated, in terms which could be understood by those skilled in the art, that in a high vacuum the current produced is under control, stable, and reproducible; and, as he employed high voltages, that higher power levels of the discharge may be obtained . . .".[22]

To return more directly to the Hammond-Purington paper: it must have been early November, 1912, rather than "in late Oc-

[19] German Patent No. 179,807; patented March 4, 1906; issued November 19, 1906.

[20] German Patent No. 236,716; patented September 4, 1910; issued July 11, 1911. Corresponding U. S. Patent No. 1,059,763; applied for January 30, 1911; issued April 22, 1913.

German Patent No. 249,142; patented December 20, 1910; issued July 12, 1912. Corresponding U. S. Patent No. 1,038,901; applied for January 30, 1911; issued September 17, 1912.

[21] J. E. Lilienfeld, "The conduction of electricity in extreme vacuum," *Ann. Phys.*, vol. 32, no. 9, pp. 673–738; 1910.

[22] (664) de Forest Radio Co., Petitioner, vs General Electric Co. (283 U. S. 664–686). Published in Book 75, "Cases Argued and Decided in the Supreme Court of the United States, October Term, 1930," The Lawyers Cooperative Publishing Co., Rochester, N. Y., vol. 131, pp. 1339–1349.

Also reported by W. R. Ballard, "The high vacuum tube comes before the Supreme Court," *Bell. Labs. Record*, vol. 9, pp. 513–516; July, 1931.

tober," that de Forest was in Gloucester, for he had gone first to the Bell people and was with them the last two days of October.[23]

Fig. 8 of the paper shows a concentric form of vacuum tube structure (with axial filament surrounded by cylindrical grid and plate), a design attributed to G. W. Pierce. As if it were new, Hammond referred it to the General Electric Company as "the proper triode design." But such form of tube was not new; it was to be seen, along with measured E-I characteristics, in a German Scientific paper.[24]

Modern Intermediate Frequency Circuitry (Section IV)

This section of the paper recounts some interesting early inventing by Hammond, but by not fully revealing other contemporary developments of intermediate frequency circuitry, gives the impression of that technique having originated largely in Hammond. Actually, it started before him and developed without knowledge of his activity, secret as it was.

In a patent interference Hammond was awarded a claim which he interprets as giving him "the broad subject matter" of intermediate-frequency circuitry.[25] The claim is quoted as reading:

> A carrier wave transmission system comprising means for receiving and detecting the energy of a modulated wave, means for selecting a component of said detected energy, and means for detecting said selected component.

Thus, the claim is for tandem detection-selection-detection. The paper asserts, "The entire principle of IF selectivity is expressed by the words 'selecting a component' regardless of whether the unselected components were to be utilized otherwise as in multiplex reception, or were to be discarded as in simplex telephonic reception."

What is disclosed in Hammond Patent No. 1,491,772 (1912–1924), of which the above is claim 46? The modulated "carrier transmission wave" is simply that of an intermittent spark discharge of definite group frequency. That group frequency is recovered in the output of a detector, is selected and then detected again, down to the signal frequency. The receiving selectivity is enhanced by tuning to the spark frequency as well as to the radio frequency.

Now this addition of selecting the spark frequency after detection and then rectifying it to obtain the signal frequency, was not new as of 1912. In 1909–1911, the Telefunken singing-spark, quenched-gap, system of wireless telegraphy which contained it was well known. The receiver comprised: a radio-frequency tuner, a detector, an amplifier sharply tuned to the spark frequency of about 500 cycles, a second detector, and a telegraph signal recorder. (The tone-frequency signals, instead of being rectified, could be read in headphones or on a loudspeaker.) The amplifier element was electromechanical, a stretched steel wire tuned to the spark frequency drove a microphone, and three such elements in tandem made up the amplifier, highly selective as it was. The system was described fully in the German technical press of 1910–1911 and in *The Electrician of London*, page 249, November 24, 1911. Since Hammond was familiar with German developments of the time, he must have known about it. A transmitting and receiving set was imported into the U. S. by the Telefunken Company of America and exhibited and offered for sale at their quarters in the little tower of the Trinity Building at 111 Broadway, New York. The present writer was employed by this American subsidiary of the Telefunken Company in 1909–1910 and helped set up and operate the apparatus. Among the U. S. military people to whom it was shown was the then Lieutenant G. C. Sweet, U.S.N. What became of this particular set of apparatus is not known, but other quenched spark sets were sold to the Army and the Navy, and altogether the system became well-known at the time. Hence in allowing to Hammond, as of 1924 on the basis of a 1912 application, this claim which reads on the Telefunken system, the Patent Office must have overlooked prior art. In view of it Hammond cannot be credited with originating the double-detection technique, starting, as it did, with audio-frequency IF.

Of Alexanderson's 1912 tuned-radio-frequency receiving circuit shown in Fig. 12, it is said that *if* the tubes of the first stage were all detectors (they being in parallel), "then the system would be of the intermediate frequency type . . .".[26] But there was then no basis for calling them detectors; the "if" is a pure supposition of the authors, made in the light of later developments. Hence the claim that Hammond's October, 1912, conference with Alexanderson had "disclosed the ultimate selective receiver with double detection . . ." (meaning superheterodyne) impresses one as unjustified.

While on this matter of Hammond's 1912 approach to the double-detection, IF, technique, it is to be observed that the twice-tandem-interrupted kind of spark transmitter of Hammond Patents No. 1,491,773 and No. 1,491,774, is shown in the little book by Miessner where one reads: "Fig. 82 illustrates a type of transmitter-receiver unit suggested by the writer in 1911."[27]

Another claim[28] for a "first" must be corrected:

> The intermediate-frequency principle was first applied outside the laboratory to the solution of a World War I communication problem of high military importance.

The apparatus is said to have been

> . . . constructed by the Hammond Laboratories in 1917–18.

and

> Delivered to the U. S. Army at Tours, France, . . . on October 10, 1918.

Actually, the IF principle had been applied "outside the laboratory" a year earlier here in the U. S. This was the short wave multiplex radio telephone system developed by the Western Electric Company, more directly by R. A. Heising, in 1916 and installed on two U. S. battleships. It is described briefly in the Craft-Colpitts paper on "Radio Telephony" given before the AIEE on February 21, 1919. Not only was this ahead of Hammond's military application, but the technique was superior, enabling a plurality of carrier channels of intermediate frequency taken from the wire carrier art, to be conveyed over a radio carrier.

Fig. 14 of the paper shows the double-detection circuit of the 1917–1918 Hammond Chaffee system, which was covered by two patents.[29] Only one of these patents shows feedback on the first stage, namely the Chaffee one, and it specifically says it is to increase sensitivity and that the feedback should *not* be allowed to oscillate. Yet the authors in afterthought say it was "usable" for heterodyne reception, and assert: "Structurally, therefore, the receiver was of the most general 'superheterodyne' variety, since both detectors could be, and during adjustment often were, of an oscillatory nature." Of course both detectors were "of an oscillatory nature," detection of oscillations being their function, and they would even tend to self-oscillation. But the first detector was not used in the self-oscillating condition; the Chaffee patent enjoined against it. Thus, one sees that the authors use "weasel words" to give the impression they had a true superheterodyne in 1918, whereas they did not.

The first to arrive at the true superheterodyne and apply for a patent on it, was one Lucian Levy of Paris. He invented it as a highly selective antistatic receiver in 1917. Soon thereafter E. H. Armstrong learned of Levy's receiver while in Paris with the AEF, the present writer has learned from Levy. Armstrong's acquaintance with the principle may have started in New York with the Western Electric Company before he went abroad, but his full appreciation of it probably stemmed from Levy more than from Hammond as the present authors suggest, although he probably knew also of the Hammond-Chaffee system. Upon returning to the U. S. after the war, Armstrong sought to exploit the superheterodyne as a means of amplifying and receiving the higher frequencies, and managed to be allowed a patent. But he had nothing new in principle and finally lost to others, mostly to Levy, all the patent claims. In 1919 Levy had offered his American rights for sale to the French house of the Western Electric Company, namely La Material Telephonique. In this way knowledge of his work came to the attention of Bell System engineers. They themselves had evolved the principle of frequency step-up-and-down, but so gradually and generally that they had not appreciated all its inventive features. But upon learning what Levy was patenting they did appreciate it as a high-gain, highly-selective, stable receiver, and purchased his American patent application. Prosecution of it in the Patent Office resulted in Levy securing the definitive patent in the U. S. to the superheterodyne as a means of amplifying and selecting, prevailing over several contestants, including Armstrong and Hammond. (Patent No. 1,734,038 of 1918–1929.) The writer in recent years asked Levy about his invention

[23] Footnote 1, p. 1198.
[24] O. von Baeyer, "Ueber langsame Kathodenstrahlen," (Concerning low-velocity cathode rays), *Phys. Z.*, vol. 10, pp. 168–176; March, 1909.
[25] Footnote 1, p. 1203.
[26] Footnote 1, p. 1201.
[27] B. F. Miessner, "Radio Dynamics," D. Van Nostrand Co., Inc., New York, N. Y.; 1916. See ch. 17, "A means of obtaining selectivity," p. 145.
[28] Footnote 1, p. 1201.
[29] Footnote 1, p. 1201, reference 50.

in relation to Hammond. In a letter of June 14, 1955, he replied: "Hammond's works were entirely unknown in France in 1917."

In the year 1920 the present writer, with the American Telephone and Telegraph Company, became aware that there was developing a considerable technique in the stepping of channels up and down in the frequency scale through modulation and demodulation in conjunction with amplification and frequency selection. He drafted two reports calling attention to the situation. They showed the existence of some fifteen patent applications internal to the Bell System, and some six inventions outside the Bell System, those of A. V. T. Day, Hammond, Meissner (the German), Levy, Alexanderson, and Armstrong. Since the Bell System engineers were early in leading into this technique, a listing of their contributions will illustrate the evolution that occurred quite independently of Hammond:

1) Homodyne reception, *i.e.*, zero-beat heterodyne, invented by B. W. Kendall in 1915, Patent No. 1,330,471 (1915–1920). Used in the transoceanic radio telephone tests of that year, including the present writer's reception at Pearl Harbor, T. H.
2) Single-sideband transmission, wherein the carrier is resupplied at the receiving end by a zero-beat oscillator, invented by J. R. Carson in 1915, in conjunction with the same radio telephone development. Patent No. 1,449,382 (1915–1923).
3) A "static" neutralizing system receiving on two frequencies, beating one down, the other up, to a common frequency, for balancing. Espenschied Patent No. 1,309,400 (1916–1919).
4) Multiplex radio telephony wherein three intermediate carrier channels are modulated en-bloc upon a radio carrier, then demodulated, selected, and finally detected in individual channels. Installed on two U. S. battleships in 1916–1917, R. A. Heising Patent No. 1,633,100 (1916–1927). This same year B. W. Kendall pointed out that Heising's multiplex system could operate with carrier suppressed, if resupplied at the receiving end by a zero-beat oscillator (recalled to the writer by Heising in 1957).
5) World War I intervened.
6) By the end of 1919, Carl R. Englund was using at Elberon, N. J., in the ship-to-shore radio telephone development, a three-channel superheterodyne receiver, according to the remembrance of Harald T. Friis and the writer.
7) Superheterodyne receivers were used on both the ship and shore ends of the development, 1919–1923. System described in paper by Nichols and Espenschied.[30]
8) Double modulation and double detection were used in the radio telephone link established in 1920 between Catalina Island and the mainland, California. Carried a superimposed telegraph channel. (PROCEEDINGS OF THE IRE, December, 1921, and *Bell Telephone Quarterly*, October, 1923.)
9) About this same time, 1920, a superheterodyne receiver was employed to realize sharp selectivity and high gain, in a radio printing telegraph demonstration between New York, N. Y. and Cliffwood, N. J., for the delegates to a preliminary international conference on electric communications.[31]
10) 1922–1923, a superheterodyne field strength measuring set, made portable by means of loop and "peanut" tubes requiring modest battery, used in ship-to-shore development, later in broadcasting.[32] Forerunner of the 4-A broadcasting receiving set referred to below.
11) 1922–1923, appearance of the first commercial superheterodyne developed for broadcast reception, the Western Electric 4-A. A copy of an engineering report on its development was given to Dr. Alfred N. Goldsmith of the Radio Corporation of America in October, 1922, and a model of the receiver itself was given to him a few months later for test purposes. This receiver was so superior to anything else available at the time that "... literally gave Elmer E. Bucher and others responsible for RCA sales the jitters," said Gleason L. Archer in his book, "Big Business and Radio" (1939, page 92).

Thus it is apparent that IF technology came into being and went into service quite independently of Hammond's more secret efforts in the field.

FREQUENCY MODULATION AND RELATED SYSTEMS (SECTION V)

As the paper indicates, the idea of FM was of long standing, without meeting much success until practiced at the higher frequencies with a correspondingly wide frequency swing.

A Hammond patent is cited[33] which undertook to transmit radio telegraphy by the familiar frequency-shift keying of the Poulsen arc while simultaneously modulating both frequencies for telephony. But the arrangement was so crude as to be substantially inoperative: telephone reception would suffer key clicks; telegraph reception in the earphones would experience interference from the telephone channel. Such a "paper patent" can hardly be said to have "established that two independent communications could be sent on the same band ..." (in fact de Forest had already done that). Certainly the disclosure of the patent was not what founded the modern practice of transmitting two chrominance signals in color television.

The presentation of the Chaffee transmitter of Fig. 16 as part of a "noise-reduction system" is misleading since the noise it undertook to reduce was that arising in the transmitter itself, not that of the transmitting medium.

It is appropriate for the authors to recall something of how wide-swing FM arose from the advance of radio to the higher frequencies, where the "natural atmospheric disturbances were of lessened importance." As tube transmitters were pushed to these higher frequencies (of the order of 50 mc), the modulation of a radio telephone transmitter tended naturally to swing the frequency. Appearing initially as a fault, the making of a virtue of this effect was a natural second thought. Chaffee of the Bell Telephone Laboratories thus sought to utilize it and devised a receiver for an FM system. The printed announcement of a paper[34] he was to give was followed immediately by a press release by Prof. E. H. Armstrong of his own invention of FM, now so well-known. In the public demonstrations made by Armstrong he compared the high-frequency FM channel with an ordinary broadcast channel, giving the impression that all the noise improvement was due to his FM system, whereas perhaps half had been bestowed by Nature!

Altogether, the attainment of the higher frequencies by means of the vacuum tube was the primary force in bringing about modern FM, the inventors being those who were on the stage at the time seeking the new. No less is the honor owed to those who, through "inspiration and perspiration," really gave it to the world.

LLOYD ESPENSCHIED
Retired Consultant
99 82nd Road
Kew Gardens 15, N. Y

[30] H. W. Nichols and L. Espenschied, "Radio extension of the telephone system to ships at sea," PROC. IRE, vol. 11, pp. 193–242; June, 1923.

[31] R. A. Heising, *J. Franklin Inst.*, vol. 193, pp. 97–101; January, 1922.

[32] R. Brown, C. R. Englund, and H. T. Friis, "Radio transmission measurements," PROC. IRE, vol. 11, pp. 115–152; April, 1923.

[33] Footnote 1, p. 1200, reference 43.

[34] J. G. Chaffee, "The detection of frequency modulated waves," presented at Washington, D. C., April, 1935. Published in PROC. IRE, vol. 23, pp. 517–540; May, 1935.
———, "Application of negative feedback to frequency modulated systems," PROC. IRE, vol. 27, pp. 317–331; May, 1939.

Rebuttal by John Hays Hammond, Jr. and E. S. Purington*

On February 3, 1958, we received the uncorrected nine-page galley sheets of a paper "Critique of the Hammond-Purington Paper Entitled 'A History of Some Foundations of Modern Radio-Electronic Technology'," by Lloyd Espenschied, a Fellow of the IRE. We appreciate the courtesy of the Editorial Board of the PROCEEDINGS in permitting us to publish this Rebuttal in the same issue in which the Critique also appears.

This Critique contains both material relating to and not relating to that con-

* Received by the IRE, March 31, 1958.

tained in our previous paper. We do not choose to comment upon the extraneous material except when it appears to have an indirect bearing upon our own material. Our critic has long been known to us as a member of the IRE History Committee. As such he has been given, in the past, much material from our files in which he had expressed interest by correspondence. We regret that we did not have the opportunity of commenting upon his present paper before it was presented for publication, and that we are now compelled to take up valuable space to clear up matters that could have been attended to by a continuation of our personal correspondence. In presenting our paper, we expected comments in addition to those kindly furnished by the reviewers of our first submitted draft, and we would have welcomed constructive criticism of our effort to establish radio-electronic history upon a more correct basis.

For purposes of later identification, we will number the items upon which we wish to comment consecutively in the order of appearance in the Critique. We will then give for each item a quotation from the Critique, to assist a reader in locating the material to which we are responding. Thereupon will follow our rebuttal or comment for each case. At the end of our rebuttal or comments on all items, we will list such references to published or unpublished material as may seem appropriate, with reference numbers for each documentation corresponding to the items to which they pertain. Moreover, we will be glad to supply copies of listed unpublished material to the Editor of the PROCEEDINGS, and to the Chairman of the IRE History Committee. The items upon which we wish to comment follow.

I. INTRODUCTION

1) *Critique.* " . . . the roots of our modern technology trace back generally to sources other than the Hammond Laboratory."

Comment. Many of the roots that nourished the work of the Hammond group and its contemporaries were recorded in our paper: the pioneering work of Wilson and Evans, Tesla, Shoemaker, in basic radiodynamics; of Edison, Fleming, De Forest in basic electronics; of Tesla and Fessenden leading to the development of basic intermediate frequency circuitry; and the initial thinking of Gueroult, Taylor, Helmholtz, Ehret in the field of frequency modulation. The later work of the Hammond group and of its contemporaries in these four fields has been set forth on the basis of our best knowledge.

2. *Critique.* "Hammond's work was conducted in secret, as the authors aver, ..."

Rebuttal. Hammond's early work for radiodynamics was conducted in secret in the same sense that the Manhattan Project in atomic energy was conducted in secret. It was not kept from the military, for example, but parts of the work were kept from Congress by the military. All that we averred was that "publication of technical and historical information was highly limited by governmental and self-imposed restrictions." The governmental restrictions were mainly by act of Congress requiring certain patent applications to be placed in the secret archives of the Patent Office, as set forth in the paper. Self-imposed restrictions were for the purpose of observing the proprieties and keeping faith with officials who expressed their convictions very freeely in defense matters. In this rebuttal, we are lifting a self-imposed restriction in one matter, because we know the officials quoted would so desire it under the present circumstances.

Radio-electronic work mainly of defense interest has always been either developed in close cooperation with the military, or in the initial stages has been brought to the attention of the proper Governmental authorities. Radio-electronic work mainly of commercial interest has had a proper outlet to the industry through conferences, demonstrations, and patent arrangements. For example, consider the early work leading up to intermediate frequency circuitry. In February, 1912, even before the filing of a patent application, information of military and commercial value was given both to B. F. Miessner, temporarily of the Lowenstein-Hammond laboratory, and to Dr. L. W. Austin of the Navy Department. Thus in a Miessner letter[1]: "Mr. Lowenstein had kindly loaned me your letters to him concerning the new selective system. Contrary to my usual custom, I believe the invention a good one. It, at least, is a very good way of dodging the Tesla and Fessenden patents. . . ." The manner of referring to the Tesla and Fessenden contributions, of course, was not the way that Hammond had put the matter, but serves to identify[2] the material under discussion. Also, in a later letter[3]: "I am glad to learn what Dr. Austin thinks about your new selective system and the probability of its being applicaable for commercial as well as tel-automatic work. Kindly accept my thanks for the letter. I believe that by being in touch as closely as possible with all phases of the work I can tackle my part of it in the proper way."

Thus there was no improper holding back of information, either from a top level Government expert, or from a technician of a subsidiary laboratory.

II. THE RADIODYNAMIC TORPEDO

3) *Critique.* " . . . but singularly omitted from mention is the one who received from the Patent Office the underlying claims, ..."

Rebuttal. The claim cited is not an underlying claim because of the clause "as set forth," without which the claim would probably not have been allowed, since it would then have read upon the wire-controlled torpedo covered by patents long since expired. A very similar claim, also with the restriction "as set forth," is in the Tesla patent, our previous footnote 2, as follows:

"5. The combination with a source of electrical waves or disturbances of a moving vessel or vehicle, and mechanism thereon for propelling, steering or operating the same, and controlling apparatus adapted to be actuated by the influence of said waves or disturbances at a distance from the source, as set forth." Neither the cited Fiske nor Tesla claims were infringed by Hammond in his 1914 *Natalia* installation, because of the "as set forth" clauses. Wilson and Evans, with a December 29, 1897, British filing date, very probably had an effective date of invention prior to both Tesla and Fiske. The precise reason that Fiske was not named by us as a pioneer is because of our understanding that from a chonological standpoint Wilson and Evans preceded him, and from a practical standpoint of experimental demonstration, Wilson and Evans and Tesla built radiodynamic models controlled from a distance. In the latter years of their effectiveness, the Fiske patents were owned[4] by the Western Electric Company. Presumably because of the response to the cited Hammond letter, it was regretfully entered into the record[5]: "I was informed that the patent was never developed into an operating machine." Fiske's works in Navy fields such as range-finders and torpedo-planes were much admired by Hammond, but not even personal friendship would be a reason for naming him as a pioneer in the field of radiodynamics.

4) *Critique.* "In the summer of 1915, Hammond was selling to a Congress worried by the European War then raging, his system of wireless control of torpedoes."

Rebuttal. This is a gross exaggeration. In March of 1915, the approach to Congress was not by Hammond but by Secretary of War Garrison[6] who urged that the Hammond inventions be given favorable consideration. But the Congressional leaders considered it then too late to enter the matter on the agenda for that session. General Weaver, Chief of Coast Artillery, then advised Hammond that if he so desired, he could now take his inventions to a better market abroad (as Hiram Maxim had done), without deserving any censure at home. Mr. Hammond elected to await the next session of Congress, and the Fortifications Committee of the House held early hearings from January 24 to February 10, 1916. Both the Army and the Navy experts reported very favorably in a hundred pages of printed record, our previous footnote 14.

Even in the summer of 1916, Congress was not worried about the war, as long as it raged only in Europe. Its concern was that sooner or later there would be a winner of that war, and that foreign battleships might eventually appear off our coasts to bombard our cities. The Hammond invention was of great interest because it promised a method of sending a powerful guided missile out to sea under precision control from an aeroplane, much farther than was possible with the ballistic missiles from coast defense guns. Senator Townsend of Michigan[7] expressed it concisely: "Now it seems to me that in these uncertain times, in these times when we are preparing for defensive war, the United States Government cannot

[1] B. F. Miessner to Hammond, February 6, 1912.
[2] Disclosure of Hammond, leading to U. S. Patents 1,522,882 and 1,491,772.
[3] B. F. Meissner to Hammond, February 12, 1912.
[4] Congressional Record, 64th Congress, 1st Session, p. 10875, col. 1; June 13, 1916.
[5] J. H. Hammond, Jr., "Telautomatics," vol. 2, p. 35; 1910–1912. Copy available in Navy Dept. Library.
[6] Congressional Record, 64th Congress, 1st Session, p. 11785, col. 1; June 30, 1916.
[7] *Ibid.*, p. 11667, col. 2; June 28, 1916.

afford to neglect an opportunity of this kind." The leisurely debate appears in the thirty-three pages of the Congressional Record in which the matter is covered, over the period June 13 to June 30, 1916. Members of Congress were quite unanimously willing to make the initial appropriation of $30,000 by which a Board of six Army and Navy officers was expected to find whether the guided missile principles developed by Hammond would be of benefit to the country. The debate in the Senate was mainly whether the Board should be required to report back to a later session of Congress, or should immediately be provided with conditional funds so that it would be able to proceed without delay if it considered the inventions worth adopting into service.

One of those who may have looked beyond the immediate horizon was Senator Stone of Missouri, who spoke[8] as follows:

"What is claimed for it? What will it do? In a word, this is what it is claimed it will do, namely, that through the operation of electrical energies controlled by the devices of this invention, an explosive body may be directed in its course until it comes in contact with a given body, stationary or movable, against which it is directed."

Are we not now, forty-two years later, seeking a device to do just that, with the body against which it may be directed not a battleship twenty-five miles offshore, but a ballistic missile from across the entire ocean?

The lack of real concern for the war raging in Europe, even as late as ten months before our entry into it, is shown by the discussion[9] as to where a plane was to be had so that the Board could make the guidance tests of the Hammond torpedo:

Senator Brandegee: "Mr. President, I was going to ask the Senator this question: Inasmuch as the House provision provides for a Board of Army and Navy experts to make whatever tests they think ought to be made, and to report upon this purchase—and the machine cannot be bought unless they should report favorably—and inasmuch as the Government owns an aeroplane for its Army, officers and aviators can operate and make the test without any expense at all, without purchasing an airship, why does not that fulfill the conditions?"

Senator Lane: "I do not think this Government owns an aeroplane which will travel through the air with any safety to the navigator: in fact, most of them have come to earth and caused a loss of life of the helmsman. I think we are practically out of aeroplanes: at least, I am so informed, although we have spent millions of dollars in building them. We ought, however, to build another type, and we should do so promptly."

Senator Brandegee: "I will say to the Senator from Oregon that I saw a very good-looking aeroplane the other day over at the Senate Office Building—it was a Curtiss machine, I believe, or a Wright machine—which the Government, I suppose, can purchase, if it so desires, for a very moderate sum. Aeroplanes, as I am advised, cost from about $7500 to $100,000, according to size and depending on the kind desired."

Senator Lane: "I also saw the aeroplane to which the Senator from Connecticut refers. It was in a good, safe position; it was near the ground, and I climbed aboard of it."

Congress was not too concerned about the war raging in Europe, even as late as the summer of 1916. It was not until after the election of November, 1916, that the country realized that such slogans as "We are too proud to fight" and "He kept us out of war," had contributed to bringing about the worries of Congress in the summer of 1917.

Actually, in the summer of 1915, while waiting for another session of Congress, Hammond was busy[10] with other activities relating to the firing of standard naval torpedoes. In a letter of August 23, 1915, Captain W. S. Sims, Commander, Destroyer Squadron, Atlantic Fleet, advised the Secretary of the Navy: "Mr. John Hays Hammond, Jr., is the inventor of an appliance to facilitate the fire control of torpedoes. He has explained the principle of the device before the Destroyer Squadron Commander and the Squadron Torpedo Officer, Lieutenant Commander J. V. Babcock, and in our opinion the device merits the Department's serious consideration. . . . It is therefore strongly recommended that Mr. Hammond be given every opportunity and facility to prosecute his experiments." Naval Ordnance then financed the work, equipment was built by the Cummings Machine Works in Boston under the supervision of A. D. Trenor of the Hammond group, and the system is a fundamental method of torpedo firing.

5) *Critique.* "Two World Wars have now occurred with no military use of the radio-controlled torpedo; it is just as well except for the futile expenditure of technical effort and public money."

Rebuttal. The first part of this statement is in error, since the meaning of "torpedo" given in the opening sentences of the section is sanctioned by Congressional and military usage. Most certainly, the inventions acquired by the Government cover applications to both water-surface torpedoes and aerial torpedoes, as well as to the underwater torpedo by which the principles of missile guidance were required to be demonstrated. The "Azon" bombs of World War II[11] and the "Glider" bombs[12] used by the U. S. Air Force in wrecking the railroad yards of Cologne are examples of "aerial torpedoes" operating by the Hammond method. The "Stingray" boats used against the Japanese and the drone-boats used against the Germans[13] are examples of radio-controlled water-surface torpedoes. The Hammond work with underwater torpedoes terminated successfully in 1931, and any failure to use them in World War II is not chargeable to Hammond. When the U. S. Congress failed to consider the 1915 recommendations of Secretary of War Garrison, Hammond was released by General Weaver from any moral obligation to deal further with the U. S. Government, and there were offers from abroad to take over the inventions and devices of the Hammond Laboratory. If Hammond had not elected to continue with the Congress in 1916, it is quite likely that water-surface torpedoes would have been used in Europe during World War I. As it was, the Germans actually did sink a British warship by a surface torpedo under wire control from a shore station.[14]

As to the latter part of the quoted sentence, we assume that what our critic meant was that the technical effort and the public money spent were futile. This is also in error. The same principles of guidance that were developed for underwater torpedoes were also applicable to the control of target ships and airborne target drones. This Hammond contribution alone has been considered by experts[15] to have justified the full cost of the Hammond effort to the Government.

6) *Critique.* "Among the principles claimed to have been developed in the 1910–1914 period is that of the automatic stabilization of the course of a torpedo by means of a gyro."

Rebuttal. No such claim was ever made. We stated as background of the modern principles of missile guidance: "In the absence of a control signal, the torpedo should be stabilized as to course by mechanisms within itself." For the purpose of disclaiming automatic stabilization of the course of a torpedo *per se*, and also of disclaiming any part of the development of the motor-driven gyro, we specifically stated: "Course stabilization had been practiced in naval torpedoes by a gyroscope energized only at the start of a run. But in 1912, ..." The "had been," therefore, very clearly refers to a time prior to 1912, which by our footnote 7 was the date of the initial Hammond concept of the use of a gyro in surface boat control work. Our critic may be sure that if gyro stabilization of the course of a torpedo by itself had been a Hammond invention, we would have used the simple past tense *was* instead of the pluperfect "had been"; furthermore, we would have given a patent number as a footnote. Our critic chooses to read *by Hammond* into the text after the word "practiced," where no such insertion was intended or justifiable.

7) *Critique.* "This development is understood to have been undertaken by Sperry alone, the Navy adopting the system upon its appearance in 1911."

Rebuttal. Gyro-compass equipment was installed upon the *Utah*, the *Wyoming* and the *North Dakota* in the latter half of 1912. On the basis of Navy reports up to March 31, 1913, we considered the work done by the Navy, especially by Chief Electrician England of the *Utah* and Ensign H. R. Saunders of the *Wyoming*, of sufficient importance to compel mention of the Navy as a party to even the technical development of the gyro-compass prior to its adaptation by Hammond for radio-control and other work in the winter of 1913–1914. If we had failed to mention the Navy, we would

[8] *Ibid.*, p. 11795, col. 2; June 30, 1916.
[9] *Ibid.*, p. 11789, col. 1; June 30, 1916.

[10] U. S. Patents 1,388,640; 1,431,140; 1,431,141; 1,431,142; and 1,431,143 to J. H. Hammond, Jr.
[11] J. C. Boyce, "New Weapons for Air Warfare," Little, Brown and Co., Boston, Mass., pp. 225–235; 1947.
[12] U. S. Patent 1,818,708 to J. H. Hammond, Jr.
[13] Authors' previous paper, p. 1192.

[14] Letter of Admiral W. S. Sims to Hammond, November 3, 1917.
[15] Letter of Admiral W. V. Pratt, CNO to JAG, January 24, 1931.

possibly have been criticized from another source.

8) *Critique.* "This modification was rather obvious since the idea of steering a ship automatically from a magnetic compass was old."

Rebuttal. We were writing about engineering, not about ideas. Our critic, as in many other instances, fails to cite any reference by which one may judge whether the engineering was along new and useful lines. The idea of an antiballistic missile is also old, but a tremendous amount of inventive research and development is necessary before the idea gets to the engineering stage, for the modern application.

9) *Critique.* ". . . as if the automatic pilot originated with Hammond."

Rebuttal. Many might consider the gyro-equipment of a naval torpedo to exemplify an "automatic pilot." Certainly Chandler, previous footnote 16, has stated, ". . . we must regard the gyroscope as the compass of the torpedo as well as its pilot." But certainly it is the function of a pilot to change the course of a craft during a run, and the devices of naval torpedos did not provide for that functioning. And the Sperry gyro-compass mechanism did not provide for a connection with the steering engine. We did not emphasize the Hammond contribution to the complete automatic pilot system now used in the navigation of surface vessels and aircraft, since we were concerned mainly with radio-electronics. Now that the issue has been raised, we consider that the officials of the Government involved would approve the lifting of a "self-imposed restriction" and the release of a part of the transcript[16] of a high-level conference of 1916 in which the Hammond contribution was discussed.

"Mr. Hammond: I think that there is perhaps another use which might be of considerable value. I have been in touch with Mr. Sperry about the proposition, and that is the gyroscope control of these boats. We have demonstrated, of course, that the steering is entirely automatic under the control of the gyro-compass, and it is very accurate, in fact a great deal more accurate than any quartermaster would be. In very heavy weather off Gloucester in a fifty-foot boat, we found that a deviation of our course on a compass was very small and better than any of the quartermasters could hold the course, and that even with sudden movements of large waves, that the mechanism responded so rapidly as not to allow the boat to be thrown off the course. It seems that that might have a very valuable application in the future, in the steering of ships in line, and in the steering of submarines, and in the steering of transatlantic vessels to maintain a more accurate course. The mechanism is foolproof, so far as we have been able to find it. Colonel Devine has observed its action for a number of miles.

"Mr. Sherley: What do you gentlemen say as to the value of that, assuming that control is such as has been indicated?

"Rear Admiral Fiske: That is a very old subject, that question of automatic steering of ships, and there are all sorts of opinions upon it. Some people think it is highly desirable to have ships steered automatically, and I am of that impression myself, provided, of course, that the apparatus is of such character that the helmsman, if he was there, in case of danger of collision, could assume control. If there was danger of collision, for instance, you would not want the ship to be steered automatically if a ship were coming across the bow.

"Mr. Sherley: Admiral, what have you to say about that?

"Rear Admiral Benson: I agree with Admiral Fiske, that if you always have a man there. There is always this element of danger if you have something running automatically you are very apt to trust too much to it and take too much risk, but the principle, of course, is absolutely correct. If you have something that is automatic, that can be changed instantly, of course it is very desirable. The only possible danger could be the fact you trust too much to it and leave it.

"General Weaver: I would like to ask Mr. Hammond if it can't be changed instantly.

"Mr. Hammond: It can be changed instantly, because in our work we have a man on the lookout so as to avoid running down targets when we are attempting to hit them, and also to get around the marine laws, because I do not know that it is permissible for vessels to travel without a crew *ad libitum* around harbors. We have found in cases of such emergency that by the mere pressure of a key the gyro-control is disconnected entirely and the control comes under the hand of the man himself. That is, instantaneous.

. . . .

"Captain Bullard: In this system that you speak of, which is your system, you only impress the energy on it when you want to change course?

"Mr. Hammond: Exactly.

"Captain Bullard: Otherwise you are set on a given course and the gyro holds it there.

"Mr. Hammond: Exactly.

. . . .

"Lieutenant Decker: I might be able to say something there. I have seen considerable service in the Navy and am quite familiar with ordinary steering apparatus on board a battleship. From what I know of the Hammond system and the system used by the Navy, I see no reason to believe that this system could not be applied to the ordinary steering apparatus as installed upon the ordinary battleship, and so arranged that the control could pass from the gyroscope or automatic steering device over to either an electrical device that is controlled by hand, or to the ordinary system as at present installed. In other words, the connection between the gyroscope control and the ordinary control is so flexible that it could be changed instantly, with the mere throwing of a switch or depression of a key."

Five years later, Admiral Bullard was in charge of the installation on the target ship *Iowa*, which to our best knowledge was the first capital ship to be fitted with automatic pilot equipment. The gyroscope was supplied by Sperry, and the linkage to the rudder was by the Hammond Laboratory and the General Electric Company. The Navy report[17] states that the equipment "will keep vessel on a steady course and is reliable. The instrument will prove of great value in any experiment or operation where vessel must maintain a steady course. It is an expert Helmsman."

The Hammond contribution, therefore, was not in the development of automatic course stabilization *per se*, nor was it in the development of a continuously-running gyroscope. His contribution was to the mechanisms for shifting the course that was stabilized by the gyro, and for switching the system from automatic to manual steering. This point of view is substantiated by some of the patent claims in the three co-filed and co-issued patents cited. Thus typically broad claim 54 of patent 1,418,788 is pertinent:

"The combination with a movable body, of means automatically operative to stabilize said body with respect to a given axis, and means to modify the automatic operation of said stabilizing means and to rotate said body selectively in either direction about said axis."

10) *Critique.* ". . . as if these patents cover course stabilization and the automatic pilot. They do not: they apply only to the application thereto of radio control."

Rebuttal. This statement is in error. There are more than ten claims such as that just quoted in which radiant energy is not mentioned. The patents cover change of the stabilized course by manual as well as by remote radiant energy control.

11) *Critique.* ". . . the arrangement shown. . . is unserviceably crude."

Rebuttal. The word "unserviceably" is highly improper. The patents were cited to support the antecedent statement and to provide a readily-available reference to the basic ideas of this new art. The reference was in a footnote because the techniques were mainly not radio-electronic. As the patents were co-issued, our critic should have at least examined the patents 1,418,789 and 1,418,791 with arrangements more in accordance with our description of the *Natalia* installation, using patented and unpatented improvements. Anyone familiar with the history of inventions knows that practically all important basic patents show devices which years later would be considered crude. The Telephone Company itself developed from the capital value inherent in the crude device shown in the Bell patent application of February 14, 1876. But the fact that such devices are considered to be "new and useful" by the experts of the Patent Office is the best nontechnical evidence that they were not "unserviceably" crude. Note that our critic is not only challenging the conclusions of our own Patent Office, but also the experts who permitted the issuance of the corresponding European patents: French 474,-906; British 16,328; German 348,277. It should be noted that this was a period in which even Army and Navy officers retained personal rights even in defense-connected inventions. Actually, these Hammond ap-

[16] "Informal Conference on the Hammond Radiodynamic System for the Control of Torpedoes," Washington, D. C., pp. 15–20; February 9, 1916.

[17] "Special Radio Report, Radio Design Problem No. 70, Radio and Sonodynamic Control," U.S.S. OHIO, p. 8; September 20, 1920.

plications were filed when, as a civilian, he was working informally with the U. S. Government represented at Gloucester by Captain F. J. Behr, and they were drawn with a view of conserving the personal rights of the inventor without disclosing the engineering details necessary for the construction of a militarily-acceptable installation.

The degree of engineering perfection obtained by the end of the year 1914 is covered by the following[18] public record:

"Lieutenant Decker: . . . On this trip we met Mr. Hammond in Boston. That was the latter part of November, 1914. The *Natalia*, the boat on which the installation was at that time, was lying at the wharf at the Navy Yard at Boston. . . . From the Navy Yard we got under way, went out of Boston Harbor under the control steering apparatus, and after we rounded the headlands outside of the harbor, we set a course for Gloucester. The boat was allowed to steer itself in order to test out the reliability of the steering mechanism. The steering mechanimsm performed as nearly perfect as anything could perform. There was not the slightest hitch, and at all times it functioned properly and kept the boat on the predetermined course. I should estimate we ran a distance of something like 15 miles from the entrance to Boston Harbor up to a lighthouse off The Graves, and during that time the boat was not touched, as I remember it, and we did not miss that lighthouse in this 15-mile run more than about a quarter of a mile. . . . As we entered the harbor of Gloucester, the observer on shore saw us. We set the boat for shore control, and the operator on shore made us perform various fancy curves over the harbor and steer around a few spar buoys. During the whole of the test, while I was on the boat, I did not see a single thing go wrong that would in any way have disabled the boat or taken it away from the control of the operator."

12) *Critique.* "The man who pioneered, and produced, the automatic pilot, was Elmer A. Sperry."

Rebuttal. This is in error. Any device which can conceivably be termed an "automatic pilot" must be one which has to do with steering, since without the word "automatic," a pilot is[19] "the steersman of a ship; that one of a ship's crew who has charge of the helm and the ship's course." Our critic has cited the "aeroplane stabilizer" developed in 1913–1914 to support his statement that Sperry pioneered the "automatic pilot." But he failed to observe that the Sperry device did not stabilize the course of the aeroplane, but functioned solely to maintain the plane at a suitable small angle with the horizon while the course of the plane was steered by a human pilot.

By attempting to set up Sperry as the pioneer, our critic now concedes that Obry and Whitehead who developed the course stabilizer for naval torpedoes did not thereby pioneer the "automatic pilot," presumably because the device was neither continuously running nor alterable during the run. Likewise, he concedes that whoever conceived the idea of using a magnetic compass as a course stabilizer was not a pioneer, presumably due to lack of reduction to practice or to lack of practical utility. With these two possibilities thus excluded, the pioneer in the field of automatic pilot systems was not Sperry; it was Hammond. The device of the *Natalia* developed in 1913–1914 provided for automatic course stabilization between the controls of the stabilized course, and provided for changing the course manually as well as by remote radiant energy signals. It was the true prototype of the automatic pilot system commercialized ten years later by the Sperry Gyroscope Company for the steering of large ships, technically known as the "Gyro-pilot," but more generally termed the "Metal-Mike."

Sperry's unquestionably basic contribution to many fields of application[20] was his development of the motor-driven gyroscope. In 1911, he successfully demonstrated his first experimental gyro-compass on the U.S.S. *Delaware*, leading to the development of an improved gyro-compass used by all Allied Navies in World War I. Also in 1911, he was working upon the problem of reducing the roll of ships, and cooperated with the Navy in producing equipment used experimentally on the U.S.S. *Worden*. As of 1927, the largest gyroscope installed for ship stabilization purposes was on the Japanese airplane carrier *Hosho*, reducing the roll by a factor of about eight. In 1913–1914, with his son Lawrence and the Curtiss Company, he developed a special gyroscope system for reduction of roll and for the automatic balancing of aeroplanes. With demonstrations in the summer of 1914 in France, the younger Sperry and "The Sperry Gyroscopic Stabilizer"[21] won the $10,000 prize in the "Concours pour la Securite en Aeroplanes." These three Sperry developments of the Gyro-Compass, the Gyroscopic Ship Stabilizer, and the Gyroscopic Aeroplane Stabilizer, were of utmost importance. But they were not automatic pilots, for the simple reason that in all three cases, the course of the craft was held and was changed only by a human pilot.

Hammond's first business contact with the Sperry Gyroscope Company was in July, 1913, in the same month that Hammond filed applications for the three patents basic to all automatic course stabilization, either by radiant energy or with manual control of the stabilized course. Sperry supplied special gyro-compass type equipment specified in a contract with Hammond dated September 8, 1913; this equipment was successfully installed by Hammond and his staff in the *Natalia* and operated in an automatic pilot system over a distance run on March 25, 1914. After this pioneer work, followed by the U.S.S. *Iowa* installation of 1921, the Sperry "Gyro-pilot" began to be manufactured and sold in considerable numbers. In 1925, after an installation on the *Leviathan* and a patent interference with Hammond, the attorneys of the Sperry Gyroscope Company could cite no prior Sperry art that would dominate the gyro-pilot claims of Hammond in the then-issued patents resulting from the 1913 filings. Presumably since these patents were in effect exclusively optioned to the U. S. Government, the matter was not pressed and the Government received the still-active gyropilot rights as an added value in its purchase agreement of 1932.

Except for supplying gyro equipment to Hammond in 1913, Sperry was not seriously involved in automatic piloting until about August 16, 1915, when he filed for a patent 1,446,276 on an electrically-sustained azimuth gyroscope. About a hundred of these devices were sold to the U. S. Navy for naval torpedo work. Sperry's difficulties with the automatic stabilizer for aeroplanes were such that he did not give early serious thought to the "aerial torpedo" or the "radio-controlled aeroplane" requiring course stabilization. Disturbed by the problems of radio reception in a plane, in which field he had not been concerned, Sperry bought rights under a Dubilier application of July 10, 1916, later patent 1,383,177. As the Chairman of the Committee on Aeronautics of the Naval Consulting Board, Sperry on April 11, 1917[22] reported upon "the whole Aerial Torpedo proposition," and within a few days was ordered to proceed to construct Aerial Torpedoes capable of carrying 1000 pounds of explosives. By act of Congress, two classes of torpedoes were to be developed, the completely automatic type and the wireless-controlled type. About a hundred test shots were made before the Armistice, at which time quantity production had been started.

Despite this intense effort sustained throughout the twenty months of World War I, the information concerning the lessons learned did not, apparently, penetrate down to Army and Navy workers in radio control of aeroplanes in the early post-war period. Early work by the Air Service of the Army was carried out in 1920 under the direct supervision of Lieutenant R. E. Vaughan, and literally started from the ground up. Hammond contributed to this development by supplying[23] security-type equipment "with a view of using it in connection with the control of airplanes by radio, the noninterferable characteristics being especially valuable for this work." Tests proved the selectivity and secrecy features of the equipment, and that control up to 40 miles was possible from a ground transmitter. It is noteworthy that in their early post-war work, both the Army and the Navy were hopeful of successful aeroplane control either with no special automatic stabilizer, or without the use of a gyroscope.

In this post-war period, patent interferences developed in the field of aeroplane control by radio, mainly between the Sperry application 207,786 filed December 18, 1917 to cover his wartime aerial torpedo work, and two Hammond patents pending based upon applications of 1914 and 1915 resulting in patents 1,568,972 and 1,568,974 issued subsequent to the interference. As a result, Hammond received the priority, and since his Government obligations were solely in

[18] Our previous paper, footnote 6.
[19] The Century Dictionary, The Century Co., New York, N. Y.; 1914.
[20] "E. A. Sperry, John Fritz Medalist for 1927," ASME Meeting, New York, N. Y.; December 7, 1926.
[21] L. B. Sperry, "The Sperry Gyroscopic Stabilizer," *Flying*, pp. 197–220; August, 1914.
[22] Deposition of E. A. Sperry, U. S. Patent Office Interferences 47,032 and 47,883, p. 44; May 3, 1923.
[23] Contracts 245 and 358, Engrg. Div., Air Service, McCook Field, Dayton, Ohio.

waterborne carriers of explosives, Hammond was able to grant Sperry an exclusive license for developing the airborne field under claims dominating the change of course of an aeroplane under radiant energy control, such as the following from patent 1,568,974:

"29. In combination, a self-stabilizing, self-steering aircraft, and means comprising a radiant energy transmission system for causing said aircraft to make any desired turn in azimuth at any point in its flight."

"30. In combination, a self-stabilizing, self-steering aircraft, means comprising a radiant energy transmission system for causing said aircraft to make any desired turn in azimuth at any point in its flight, and means for automatically banking the craft while turning."

As a result of the Hammond-Sperry agreement of 1925, all later radiant energy airborne-guided missile-type equipment produced by Sperry involving a change of a stabilized course, carried Hammond patent numbers. Pre-World War II radio-guided aeroplanes came to be known as Hammond-Sperry Drones.

Additionally, it is to be noted that automatic pilot systems commercially developed for passenger planes with a pilot in attendance are covered by the basic patents of Hammond filed in 1913, such as that cited in the discussion of Section II–9 above.

Therefore we submit that Hammond, and not Sperry, was the true pioneer in the development of the "automatic pilot" for waterborne craft, and that patentwise, the invention is dominated by patents that went exclusively to the U. S. Government in 1932; that Hammond's gyro-pilot ideas applied in the waterborne field were also applicable in the airborne field, regardless of whether there was a human pilot also available, or whether the craft was radio-controlled; that in the explicit field of radiant energy control of airborne torpedoes in which he was not obligated to the Government, Hammond granted the Sperry Company exclusive rights which it acknowledged and exercised in pre-World War II aerial torpedoes and the like. Hammond pioneered the "automatic pilot" and both Hammond and Sperry produced it. Sperry[24] did not hesitate to regard Hammond as a personal friend as well as an associate in this development.

13) *Critique.* "But ahead of Tesla, apparently, were the two British inventors, Wilson and Evans. . . . The receiving control electromagnet is made dependent for its operation upon the receipt of both channels. . . . "

Rebuttal. The cited patent in Fig. 7 shows such a control magnet. However, there does not seem to be any patent claim specific to this manner of practicing the invention, nor to any security advantages. The Wilson and Evans arrangement would be subject to interference on a "channel" intermediate between the horizontal and vertical polarization "channels": the Tesla system would not so readily be forced by any single channel, since in the Tesla system, the word "channel" is in the usual frequency sense. It is the Tesla system that is the background of the Hammond single-shot FM system of security used on the *Natalia*.

14) *Critique.* "de Forest had devised . . . a diplex system of telegraphy. . . . "

Rebuttal. Such a system is irrelevant to the discussion. Hammond's system was simplex, depending upon the reception of both ends of the transmitted spectrum to establish a single control signal. The de Forest system, apparently, was diplex, with one end of the spectrum for conveying one message and the other end for another independent message. There was no cooperative action between two channels to produce a single signal as in the Hammond security system and as in modern FM reception.

15) *Critique.* " . . . the claim is for a *torpedo*, meaning in water. . . . "

Rebuttal. The words "as set forth" do not appear in the claim, therefore the claim is not limited by the drawings and specifications, but only by the allowable breadth of the words "torpedo" and "energy." We consider the word "torpedo" applies to aerial torpedoes, and that "energy" applies to electromagnetic energy. Note that contrary to the statement of our critic, the word "fuse" is absent from our paper. While the "proximity fuse" principle may apply to airborne devices, the "proximity" principle applies to both air- and waterborne devices. Neither we nor presumably our critic are in a position to know whether or not the proximity principle, in fact, has been applied in both media.

III. The Triode Tube

16) *Critique.* "The quotation is correct, but the assumption that it referred to the triode rather than the diode, as of 1906, is in error."

Rebuttal. On the contrary, our assumption that it applied to the triode both in 1906 and also later is not in error. The Marriott reference in its entirety on this point is:

"5. Audion. This form of detector was used to some extent as early as 1906, but apparently in small numbers until about 1912 when the amateurs became active in its use, and within the last year or more it has been used to some extent by the Government."

Now our critic later states, "It was about that time that the grid audion began to come into some use, 'but apparently in very small numbers' as Marriott said." Since Marriott referred to "this form of detector," and since the form of detector as of 1907 was the grid audion form, as our critic admits, it follows that Marriott was referring to the detector of 1906 as a triode. Our critic does not openly say that the Marriott statement was in error, but does say that we were in error in interpreting it. Our critic, in his first printed discussion of the Marriott paper in the Proceedings,[25] failed to challenge the Marriott statement in this matter, as he must now do.

17) *Critique.* "The grid triode appears to have been invented toward the end of that year or the beginning of 1907."

Rebuttal. The patent application covering the invention of the grid triode[26] was signed by de Forest on December 21, 1906. Therefore, since de Forest undoubtedly used the grid triode in 1906, it appears that the Marriott statement as to dates is just as correct as our interpretation of it.

18) *Critique.* "It . . . issued as a patent February 18, 1907. . . . (There was no time lost in those days!)"

Rebuttal. The grid triode patent referred to issue February 18, 1908. The parenthetical statement shows that the error could not have been due to any typographical error of the printer.

19) *Critique.* "Among those small users, to the writer's knowledge, were a few amateurs. . . . The writer, with two other amateurs . . . attended de Forest's Brooklyn lecture. . . . "

Comment. Our critic, according to "Who's Who in Engineering," joined the engineering staff of the Telephone Company in 1910. There was, therefore, at least one member of the Telephone group who knew of the de Forest triode two years before the Telephone Company, as our critic later states, learned about it in 1912.

20) *Critique.* "The grid audion detector continued for some years solely as a detector."

Comment. The critic here lapses into the early usage of "Audion detector" as meaning a triode tube regardless of the use to which it was put in circuitry.

21) *Critique.* "But they do not tell the whole story and give the impression of Hammond's role having been more than it was."

Rebuttal. There was no need of telling the whole story. The prior attempts at developing the triode were recorded in the first and last parts of the paragraph, showing that Lowenstein still, in 1912, had a lot more to develop. Hammond's immediate interest was the procurement of a better type of "relay-operating rectifier-detector," and in this branch of the work, Lowenstein may have been a true consultant. At least, he never changed the words "Consulting Engineer" on his letterhead[27] when writing about any phase of the ion controller project. Hammond's role quickly became that of an unsecured creditor, but with paper rights in exploiting the triode developments. On the basis of personal friendship, a settlement of financial matters was made with Lowenstein about three months before his sale of the grid bias patent to the Telephone Company, at such a figure that the entire amount received from the sale went directly to Lowenstein as a clear profit. The Hammond role was not exaggerated in the paper.

22) *Critique.* "The term 'controller' or 'ion controller' was Fritz's alias for the name 'Audion' which de Forest had bestowed upon Fleming's tube upon adding to it the 'B' battery."

Rebuttal. This is greatly in error. The Lowenstein "ion controller" was a triode, and not a Fleming diode. Our critic may have confusedly substituted a battery ex-

[24] Letter of Sperry to G. A. E. Lundell, October 21, 1926.

[25] L. Espenschied, "Discussion," Proc. IRE, vol. 5, p. 196; June, 1917.

[26] U. S. Patent 879,532 to L. de Forest.

[27] Letter of Lowenstein to Hammond, November 13, 1911.

ternal to the Fleming valve, where he may have meant a grid within the valve.

23) *Critique.* "The strange thing is that it was not followed up."

Rebuttal. Our conjectures as to why the work was dropped by Lowenstein are stated in the paper. Lowenstein's research was made available to the General Electric Company,[28] with more facilites and competence than Hammond-Lowenstein for carrying out "a systematic investigation of the influence of the vacuum." Dr. G. W. Pierce once recalled to us, "We were not at all sure there was any great future for the triode tube, but felt that any future lay in the development of the hard tube." The hard tube was a General Electric development.

24) *Critique.* "And it is singular that to Hammond, 'The exact nature of the ionic devices and the manner of operation are not known'."

Comment. The technical details may be in the Lowenstein files concerning which our inquiries have been futile. There is a suggestion of a method of regeneration in the book, critic's reference 27, in the figure of a detector circuit for which unusual sensitivity was claimed. (See Fig. 77, p. 133.)

25) *Critique.* "One wonders . . . that the evidence of . . . an oscillation generator . . . was not presented to the courts."

Rebuttal. There is a statement of the oscillatory use of triodes in the book, critic's reference 27. (See p. 179.)

26) *Critique.* " . . . the letter which Hammond wrote the President of that company. . . . "

Comment. This action proves, if nothing more, the Hammond objective of getting radio-electronics on a foundation free from foreign domination. It also discredits the critic's view that all of Hammond's work was conducted in secret.

27) *Critique.* " . . . a copy of the letters was sent to him. . . . "

Comment. We wish to make it clear that the letters were not given to our critic for that purpose.

28) *Critique.* "Arnold later testified . . . 'I was very much astonished and somewhat chagrined. . . . ' "

Comment. Arnold's chagrin must have been matched later by that of our critic who had known of the superiority of the de Forest detector during two years of engineering service with the Telephone Company.

29) *Critique.* "There was ordered from Germany the latest type of vacuum pump."

Comment. This is a point of important historical significance, since it shows the Telephone Company work was much later than that of the General Electric Company in the production of high vacuum tubes.

30) *Critique.* " . . . the company originally could have had the patent for $20,000."

Comment. The Lowenstein patent was used by the Telephone Company in many infringement suits, and presumably made a profit even at the final figure.

31) *Critique.* "Actually, Lowenstein and Hammond, by coming into the picture when they did . . . performed a major service."

Comment. Mr. Hammond, by foreign travel and contacts, realized that the growth of the "Telefunken" and other foreign companies, and their penetration into the United States, could be checked only by interesting large electrical companies in getting into the communication and equipment field. Appearing before the board of the General Electric Company for the purpose of interesting it in an American radio company, he was unable to make the progress later achieved by the efforts of the Navy Department. It is gratifying to learn that his indirect approach to the General Electric Company, the Telephone Company, and the Federal Company is getting to be accepted as a major service.

32) *Critique.* "It must have been in early November, 1912, that de Forest was in Gloucester."

Rebuttal. The premise that de Forest first went to the Bell people is in error. Our documentation is a letter[29] and a telegram[30] which places de Forest in Gloucester in late October, as stated in the paper. The visit was not in connection with the amplifier work of either de Forest or Lowenstein, and the information was casually given.

33) *Critique.* "As if it were new, Hammond referred it to the General Electric Company as 'the proper triode design'."

Rebuttal. The clause "as if it were new" is unjustified. It was an engineering design, expressing the personal views of the members of the Hammond group. This design was not at first followed by the General Electric Company for the reasons stated. Nor was it followed by the Telephone Company in their first wartime type E and J tubes. The original pencil sketches, of which our critic presumably has a copy,[31] carry two dated signatures. These are the signatures of attorneys of the General Electric Company, and not of anyone of the Hammond group. In fact, the only direct indication of the Hammond source of the design is that the material would be recognized as being in the handwriting of Dr. G. W. Pierce, then a consultant to the Hammond Laboratory. Not even he bothered to sign the sketch.

IV. Modern Intermediate Frequency Circuitry

34) *Critique.* "In a patent interference, Hammond was awarded a claim which he interprets as giving him 'the broad subject matter of intermediate frequency circuitry'."

Rebuttal. Our statement rather was, "the broad subject matter in controversy was awarded Hammond." We have very clearly stated our belief that Hammond contributed the selective features of intermediate frequency circuitry, while it was Alexanderson who, due to his development of the TRF amplification idea, contributed the sensitivity features. The withdrawal of the Levy patent application from the interference should not be overlooked.

35) *Critique.* "In 1909–1911, the Telefunken. . . . "

Rebuttal. The Telefunken equipment was not pertinent to the decision in the interference, because what our critic calls a second detector was not a detector. In fact, our critic later states parenthetically: "The tone-frequency signals, instead of being rectified, could be read in headphones or on a loudspeaker." Apparently, our critic's subconscious mind compelled him to use the correct word "rectified" as applied to a tonal frequency signal, and it was only his conscious mind that has attempted to call that rectifier also a detector. As our critic and former advisor of attorneys of the Telephone Company well knows, subconsciously at least, the question of when a rectifier is also a detector was thoroughly discussed in the interference. Our critic gives no patent number for the Telefunken equipment. Its operation in the respect here discussed is unquestionably subservient patentwise to earlier references cited by the Telephone Company in the interference. A brief review of that case appears to be in order.

The patent claim that has been cited originated with Heising or his attorneys. It soon developed that, if patentable, Hammond would win the claim over Heising and over Levy. Levy withdrew. Thereupon, in accordance with usual legal procedures, the Telephone Company, in order to try for a partial victory, attempted to prove its own written claim was actually unpatentable over art prior to both Heising and Hammond. Cited, for example[32] were Fessenden 727,326; Fessenden 752,894; Blondel 824,682; and Scheller (German) 208,836. The Scheller patent was cited especially to upset the claim as to "selecting a component" and the others more especially to upset the claim as to "detecting said selected component." These efforts of the Telephone Company were unsuccessful, and would also have been unsuccessful if the well-known Telefunken equipment had also been cited. If there was anything established technically by the interference, it was that the word "detector" relates to a device involved in the finding of something hidden, and therefore a device acting upon audio-frequency currents already being indicated by headphones or a loudspeaker cannot be termed a detector. If therefore follows that the circuit preceding a second detector must contain currents of a frequency or frequencies above the audible frequency range. Our concept of intermediate-frequency circuits was clearly stated at the beginning of the section, and it is in accordance with present accepted usage.

36) *Critique.* " . . . the Patent Office must have overlooked prior art."

Rebuttal. The Patent Office considered carefully, at the very top level, all the art which the experts of the Telephone Company presented. If the Telefunken equipment had been pertinent, and it was not, any failure to uncover it would not be chargeable to the Patent Office, since this was an important interference and not a usual search case involving Patent Office personnel only. The final legal decision was by three Examiners-in-Chief, paper 67, January 27, 1922. Because of the importance of the matter, the papers then went before William A. Kirman, First Assistant Commissioner, so that both parties would have the informal views of even a higher official of the Patent

[28] Letter of Alexanderson to Hammond, October 21, 1912.

[29] Hammond to Lowenstein, October 24, 1912.

[30] Hammond to Lowenstein, October 23, 1912.

[31] Enclosures in letter of W. G. Gartner of the General Electric Company to Hammond, July 1, 1913.

[32] Brief for Heising, U. S. Patent Office Interference 43,858, Paper No. 64; October 21, 1912.

Office. On January 29, 1923, he wrote: "There is no appeal from this holding of the Examiners-in-Chief that Count 4 is patentable. . . . This patent to Scheller does not, in my judgement, disclose any 'means for selecting a component of said selected energy'. . . . The decision of the Examiners-in-Chief is affirmed." And on April 13, 1923, Mr. Kirman further wrote: "As to the allowance of the claim to the party Hammond, it may be allowed in any application in which it can be properly made." The case is finished, the Telephone Company bought rights under the Hammond patents, the patent numbers were put upon equipments, the patents have expired. It is rather late for a consultant of the Telephone Company to now suggest that the Patent Office overlooked such a well-known "Telefunken" art and that the Telephone Company bought patents that were invalid.

37) *Critique.* "Hammond cannot be credited with originating the double-detection technique, starting, as it did, with audio-frequency IF."

Rebuttal. Readers will be amazed to find someone who as of the present date will refer to an *audio-frequency intermediate frequency.*

38) *Critique.* "Of Alexanderson's 1912 tuned-radio-frequency circuit shown in Fig. 12. . . . "

Rebuttal. The caption of this figure does not justify this interpretation. The circuit shown originated with Alexanderson after his invention of TRF in Gloucester, and came to Hammond in a letter of October 21, 1912.[33] The word "later" in our paper should not be overlooked. For the purpose of showing the relations between the TRF system of Alexanderson and the IF system of Hammond, we were entitled to hypothesize the devices commonly termed "Audion detectors" or simply "detectors," either as actual detectors or as actual amplifiers.

39) *Critique.* "One reads: 'Fig. 82 illustrates a type of transmitter-receiver unit suggested by the writer in 1911."

Rebuttal. The implication of this statement is most readily refuted. Fig. 82 of the book cited corresponds almost exactly with two figures of U. S. Patent 1,491,773. That Hammond was the inventor of the patented material shown in this patent is evidenced by his signature at the end of the specifications. If our critic does not consider this sufficient, the fact that Miessner was not the inventor of the patented material is evidenced by his signature as a witness. Our critic presumably has failed to note an alleged relation between the receiver system of the figure cited and the TRF circuitry of Alexanderson, available in a reference[34] cited in our previous paper.

40) *Critique.* "Another claim for a 'first' must be corrected."

Rebuttal. In preparing our paper, we endeavored to leave out any reference to the word "Hammond" wherever we considered that there would be no misinterpretation as to the meaning. Most readers, we believe, made a mental insertion of the words *by Hammond* after "was first applied" in the cited passage.

41) *Critique.* "The apparatus is said to have been. . . . "

Rebuttal. Some may construe this method of expression to convey doubts on the part of the critic that the alleged equipment ever existed, or, if it existed, that it ever went to France. Our records show that on November 2, 1918, a conference was held at the Department of Development and Inspection,[35] Signal Corps, A.E.F., in France, to discuss this Hammond equipment with Messrs. Chaffee and Buswell of the Hammond Staff. Present were General Russel, Colonel Carty, Captain Armstrong, and Lieutenant Fahys. From published biographical material, it is most certain that Colonel Carty ("who during the whole interview was polite and cordial") was then also the chief engineer of the Telephone Company, and one with whom Hammond had had dealings in 1912 in the Lowenstein amplifier matter. Our critic undoubtedly can check with the records of the Telephone Company to settle his doubts in this matter.

42) *Critique.* " . . . the short wave multiplex radio telephone system developed by the Western Electric Company. . . . "

Comment. This is presumably the equipment upon which the party Heising properly wrote the patent claim which has been cited. If so, this equipment was later proven to be subservient to the prior Hammond art. The time lag between installation and publication is noted, but without surprise. Contemporary equipment upon which publication was also delayed included the Hammond IF system that had been previously used in control work, and that upon which construction was started in 1917 and which was delivered in Europe in 1918. The comparison between the merits of the Heising naval equipment and the Hammond military equipment is out of order. The Telephone Company work was in communication between ships; the Hammond work was in finding a military solution to such problems as avoiding the necessity of requiring U. S. Infantry to advance into its own Artillery barrage.

43) *Critique.* "Of course both detectors were 'of an oscillatory nature,' detection of oscillations being their function."

Rebuttal. This statement is improperly based upon a confused concept of an oscillatory detector. Crystal diode detectors were detectors of oscillations but they were not therefore oscillatory detectors. Although the first detector of the equipment under discussion was advisely used in the non-oscillatory condition for the proper reception from the corresponding military transmitter, it did often oscillate during the course of making adjustments. Under these conditions, the circuit most certainly had the structure of a CW "superheterodyne" regardless of whether or not there was any CW to be received. We fail to understand why our critic is disturbed by our remark upon this point, since the CW "superheterodyne" is an Alexanderson invention in whatever respects it is not a Hammond invention, as now to be recorded.

44) *Critique.* "The first to arrive at the true superheterodyne and apply for a patent upon it, was one Lucian Levy of Paris."

Rebuttal. This is in error. The French patent to Levy was filed August 4, 1917, and his U. S. Patent 1,734,038 was filed August 12, 1918, with an additional figure. This U. S. patent later received claims 1, 2, 3, 6, 7, 8, and 9 from the well-known Armstrong patent. These claims relate to the amplification of readily amplifiable intermediate frequencies in the telephonic-type superheterodyne. At the same time, and in the same manner, Armstrong claims 4 and 5 went to Alexanderson U. S. Patent 1,508,151, and claim 4 remained there as claim 20. This relates to the amplification of readily amplifiable intermediate frequencies in the CW-type superheterodyne. This U. S. patent was a divisional patent based upon disclosures that are also in a prior Alexanderson patent 1,465,961, filed April 19, 1916, over a year before the date of Levy's filing in France. This is the effective filing date of both the Alexanderson patents, and it takes precedence over Levy as to the general concept of amplifying a readily amplifiable IF in a superheterodyne. Before both Alexanderson and Levy, Hammond in 1912 had filed upon his U. S. 1,491,774 which specifically covers the telephonic-type superheterodyne structure, but of course without the addition of tuned amplification of superaudible frequencies which was a later Alexanderson contribution, also of 1912.

45) *Critique.* "Hammond's works were entirely unknown in France in 1917."

Rebuttal. This statement is in error. Hammond at that time had several patents pending in the French Patent Office. Radiodynamic patents 474,906 and 475,888 had been published in 1915; intermediate-frequency patent 519,811 was entered into the French Office on December 12, 1917, based upon U. S. Patent 1,491,775, filed September 28, 1916.

46) *Critique.* "Thus it is apparent that IF technology came into being and went into service quite independently of Hammond's more secret efforts in the field."

Rebuttal. This statement is not justifiable by the facts. That Hammond's efforts were not secret even from the pioneer in the commercialization of the superheterodyne, and that they had a very definite technological bearing is evident from the following incident. On November 8, 1918, Captain E. H. Armstrong requested the presence of Dr. E. L. Chaffee to explain a technical point as to the proper design of intermediate-frequency transformers. The desired information was freely given.

V. Frequency Modulation and Related Systems

47) *Critique.* "A Hammond patent is cited. . . ."

Rebuttal. Our critic apparently cannot think back to the period of the filing date of this patent, when receivers were very broadband, CW signal rates were twenty words per minute, and key shifts were 300 cycles. The patent certainly could be practiced without serious cross-signalling, just as can its modern television counterpart. There are other patents in the Hammond group, and other pertinent Hammond activities, that make the tie between the early patent and

[33] Alexanderson to Hammond. October 21, 1912.
[34] Authors' previous paper, footnote 32, p. 1198; see p. 1366 of the reference.

[35] Compare this address with that of p. 5 of the reference in the authors' previous paper, footnote 62.

modern color TV technique more close than would appear from our paper.

48) *Critique.* "The presentation of the Chaffee transmitter of Fig. 16 as a part of a 'noise reduction system' is misleading. . . . "

Rebuttal. The system undertook to reduce all noises coming into the receiver insofar as they produced equal effects in circuits tuned to the two ends of the spectrum. Hum reduction was cited as an obvious example. The system also reduces disturbances that are not so equally present in both ends of the spectrum. Our critic should check the nature of the spectrum shown with that of a sine-wave FM signal with a modulation index of 2.4 to see that the Chaffee system had the noise reduction merits of modern FM systems, except as to amplitude limiting.

49) *Critique.* "Appearing initially as a fault, the making of a virtue of this effect was a natural second thought."

Rebuttal. Knowledge of this effect even in the early broadcast brand of frequencies was the basis of the 1921 work of our junior author in developing the first all-electronic type of FM transmitter, with performance as indicated in Fig. 15.

50) *Critique.* "Chaffee of the Bell Telephone Laboratories thus sought to utilize it and devised a receiver for an FM system."

Comment. By the critic's footnote reference to the papers of J. G. Chaffee, this work was published in 1935. By that time, nearly all the really basic work in the field of FM had been completed and the important patents of the workers cited in our paper had come to issue. It is further believed that the Telephone Company was among those which had rights under the patents and the patent applications covering the Hammond Laboratory work in this field. If it had so desired, the Telephone Company could have made a good commercial beginning in the FM field on the basis of the Hammond patents alone.

General Comments

Our critic's attempts to downgrade the judgments of contemporary experts as to the value of Hammond's work are not the first. At least three prominent pioneers in radio-electronics advised the Government against taking its initial step in radio guidance of missiles. The acting Secretary of War,[4] in rebuttal advised Congress, in effect, that: Hammond's work was far advanced over the prior art, with a hundred patent applications and a thousand allowed claims; the patents of Rear Admiral Fiske which some claimed to be basic would expire in about sixteen months, so that there was little danger of trouble from the Western Electric owners; Hammond was holding the Government free from liability in case of infringement suits; and the actual risk was but $30,000 since the Government would not be obligated in case of an adverse report by the newly-to-be-created Board. The recommendations of the previous Army and Navy officers had been so favorable that Congress had no hesitation in proceeding. The value of the inventions, developments and patents had been affirmed by the legal departments of the Army and the Navy, and is most generally accepted by officers and others who have examined the facts. As to the field of intermediate-frequency circuitry, chief patent attorney George E. Folk of the Telephone Company conferred with Hammond even before the termination of the Heising interference, and acknowledged that Hammond was going to be the winner. Therefore he took steps to secure rights under the Hammond inventions and patents for the Telephone Company and further advised the Radio Corporation he thought it should do likewise. This action, coupled with the direct findings of the Radio Corporation and similar advices from others including E. H. Armstrong, resulted in the Hammond work becoming immediately available to the industry without any further litigation.

Conclusion

In conclusion, we deeply regret the necessity of having had to point out the many errors of simple facts on the part of our critic. We regret the necessity of having had to correct so many misinterpretations of our statements in the paper, on points that should not have disturbed him and others. We regret the necessity of having had to go into more personal matters, to show how the various developments appeared to others at the times at which they were being made. By failing to comment upon other matters than the fifty items here discussed, we do not wish to be considered negligent, if there are errors of fact in the extraneous material presented in the critique.

John Hays Hammond, Jr.
E. S. Purington
Hammond Research Corp.
Gloucester, Mass.

Replication of Rebuttal by Mr. Espenschied*

Of the fifty points of rebuttal, only a few need further attention:

3) Of the twin Fiske patents of 1900 on the wireless control of torpedoes, Nos. 660,155 and 660,156, Hammond in 1912 called one of them "the first" of its kind, when seeking information about it, as already noted. Subsequently he ignored Fiske, and now the authors reject him. In claiming Hammond did not infringe Fiske, there is quoted a claim, obviously broad, with the contention that the words with which it ends, "as set forth," limit its scope in some unspecified manner. This refusal to recognize a predecessor lends significance to certain additional information known to the writer which is now presented in justice to Admiral Fiske.

When in 1915–1916 Hammond was selling to a war-worried Uncle Sam his own project of a wireless-controlled torpedo—which never succeeded—Admiral Fiske became concerned lest his patents be infringed and wrote the company to which he had assigned them, the Western Electric Company. That company then had the matter studied by a college-professor patent attorney, P. I. Wold. On March 9, 1916, he reported to Western's Patent Counsel, D. C. Tanner, saying "we have made a study of the Fiske Patents and certain patents to Hammond, who is a possible infringing party." He concluded:

> . . . that the claims of the Fiske patents are valid, are not subsidiary to other patents and are basic in scope. His arrangement is operative. Hammond has a patent for a somewhat similar device but his claims are limited to his specific apparatus and are furthermore subsidiary to the Fiske claims. We are also unable to see, at present, how Hammond or others can make any devices of this nature without infringing Fiske's patents. Fiske's patents, however, expire in about 19 months and any steps, such as infringement warnings, should therefore be taken in the near future. In view of Court decisions on Government contracts it may be that Hammond or others will ignore these patents, in which case our only recourse will be to the Court of Claims.

On March 10, 1916, Tanner sent to his superior, Mr. Swope, a copy of the Wold report. He recommended against taking "any such action against the Government, . . . in view of our relations." He suggested the company offer to turn back to Admiral Fiske his two patents, "upon repayment by him of what they have cost us." They are not known to have been repossessed by Fiske. Probably he realized only too well the trials and expense of a lengthy law suit. He is not known to have gained anything from his imaginative and patriotic endeavor; and now we see him denied even recognition. Such is hardly objective technical history!

6) Three statements in the original paper gave the impression that Hammond claimed to have developed, in 1910–1914, the principles of automatic course stabilization by means of the gyro:

> Many developments by the Hammond Laboratory established basic principles used in modern airborne guided missiles, including the stabilization principle. . . . (Summary of paper.)
>
> Preliminary work by Hammond . . . resulted in the development of the automatic course stabilization principle . . . (p. 1192, column 1, lines 27–30).
>
> Automatic Course Stabilization. . . . The first navigational application of this automatic pilot principle was to the third boat, the *Natalia* of Fig. 2; the system was first put into long period operation on March 25, 1914. . . . (P. 1192, column 2, lines 34–38.)

Surely these claims give that impression. But all is well, now that the authors deny such intention.

12) The insistence of the authors that Hammond, and not Sperry, was the true pioneer in the development of the "automatic pilot," goes quite against the writer's impression, and doubtless of many others, but he will not further press the point, leaving it to others better qualified to judge. Perhaps the Sperry people will have something to say. Here is a good example of the many versions one can have of a certain technological origin, depending upon the viewpoint.

15) Hammond U. S. Patent 2,060,198 describes a torpedo in water and electromechanical means for communicating by compression waves with the object to be detected. In the absence of the disclosure of another medium and corresponding communication means, the term torpedo in the claims is to be construed in its ordinary meaning of a waterborne body. Hence, the attempt at making the claim read on the radio-controlled proximity fuse by calling the action the "Proximity" principle, is misleading.

* Received by the IRE, November 3, 1958.

16), 17), and 18) Robert Marriott's statement concerning the early use of the Audion, namely that:

This form of detector was used to some extent as early as 1906.

is quite correct if it is understood what "this form of detector" was as of 1906. It was the diode, or two-element form, with "B" battery added by de Forest, as described in his 1906 AIEE paper entitled "The Audion." Marriott was well aware of this, the first, form of Audion, and was not limiting himself to the subsequent form, the grid triode, which later "stole the show." It was the authors, Hammond and Purington, who so limited him, by inserting, in connection with his statement, the words "three-electrode form of detector." Thereby Marriott's statement was altered, making the use of the grid Audion appear somewhat earlier than was the case. A small point, but one worth keeping straight historically.

The grid Audion was invented in the winter of 1906–1907. It began to be "used to some extent" not until 1907 to the writer's knowledge. The first public disclosure of it was by de Forest himself in his Brooklyn lecture on "Wireless" of March 14, 1907, already mentioned. The patent for the grid Audion was applied for January 29, 1907. It issued February 18, 1908 (not 1907 as previously stated by the present writer, an error he is glad to have corrected)—No. 879,532.

22) The authors appear not to have understood the comment. Fritz Lowenstein, employing de Forest's grid Audion, but wishing another name for his own purposes, called it an "ion controller"—much as had de Forest re-christened the Fleming Valve, to which he added a plate voltage, calling it "The Audion" as per his 1906 AIEE paper. How interleaving are inventions, but how reticent inventors to admit it!

29) The present writer mentioned several steps taken by Dr. H. D. Arnold in the telephone company's laboratory in attaining higher vacuum and generally improving de Forest's grid Audion, one of them being the importing of the latest kind of vacuum pump, the Gaede molecular, received in April, 1913. From this date, the authors assert " . . . the Telephone Company work was much later than that of the General Electric Company in the production of high-vacuum tubes." Quite the opposite was the conclusion of the Supreme Court of the United States in the famous Arnold-Langmuir litigation over the high-vacuum tube, where the telephone company's priority was recognized in these words:

August 20, 1912, the earliest date claimed for Langmuir was rejected rightly, we think, by the district court, which held that Langmuir was anticipated by Arnold in November, 1912.[1]

The high-vacuum tube, in its use in transmitting as well as receiving, over wires as well as by radio, proved to be second only to the grid Audion itself as a cornerstone of modern radio-electronics.

[1] The decision, handed down May 25, 1931, recognized also that de Forest himself had gone part way to high-vacuum, without fully knowing what he was doing. Published in "Cases Argued and Decided in the Supreme Court of the United States," Book 75, Lawyers Edition, 1931. De Forest Radio Co., Petitioner, vs General Electric Co., No. 660.

Finally, a personal note. The writer's discussion of the Hammond-Purington paper has been offered to balance its excessive claims, in the interest of more correct technical history. The writer is not a spokesman for the telephone company, has not consulted the company, from which he has been retired some years. He has merely used knowledge that came to him in a long active career in the field.

Comments on "Replication of Rebuttal" by Mr. Hammond and Mr. Purington*

At the outset, we wish to thank our critic for making clear to all what we had long since sensed, that his "Critique" of our September, 1957 paper in the PROC. IRE was not reviewed by an executive of the Telephone Company. We are pleased that we must proceed further with only nine of the fifty items which we discussed in our first response. Our final comments follow.

3) *The Fiske Patents:* In our files, the only statement by Admiral Fiske concerning these patents is that made in the high-policy conference of February 9, 1916, (our rebuttal reference 9a, p. 35 of the report): "Rear Admiral Fiske: I suppose you understand that I am not an unbiased witness. I want that understood. I invented this general scheme in 1897 and got a patent on it, and the patent has not expired. It is a basic patent. I want that understood." There was no mention of any reduction to practice either by Fiske or by the Western Electric Company. Our interpretation of the word "as" in the phrase "as set forth" is the dictionary meaning "in the same manner." We feel no patent lawyer would accept a claim with the words "as set forth" at the end of an otherwise basic claim unless by compulsion. We did not quote the Fiske claim, but a very similar Tesla claim. By the words " . . . our only recourse . . . " in the Wold report to Tanner, it would appear that Prof. P. I. Wold was more closely related to the Western Electric Company than our critic would like his readers to believe. Upon continued study of the patents after his inquiry concerning them cited by our critic, Hammond found design errors which experts believed rendered the structures inoperative. Even after the Tanner recommendation not to take action against the Government, the possibility of such action was called to the attention of the Government by another and presumably disinterested engineer, to which Hammond responded by guaranteeing to hold the Government harmless in case of any and all patent difficulties. In a paper covering contributions to technology, it would have been improper to refer to one as a pioneer whose inventive effort was not reduced to practice by himself or by others as his agents. Please note, however, that in rebutting our critic in this matter, we do not wish to create the impression that we do not have great admiration for the extensive pioneering work of Fiske in other fields of naval development.

" . . . which never succeeded . . . " is a complete misstatement of fact. The Hammond torpedo-control system tested by the Navy at the Newport Torpedo Station fulfilled the exacting requirements of the Chief of Naval Ordnance, and the report of these successful demonstrations was given by Admiral Leahy to the Secretary of the Navy. The non-use of the device was due to naval policy and this was formulated upon the problem of mass production for war and difficulties encountered with the new magnetic detonators. The Hammond control system was widely used in the control of naval targets, and the Hammond-Sperry inventions created the target "drones" and finally the basic control of modern missiles.

6) *Developments of 1910–1914:* The sentences which gave our critic trouble will, we believe, not confuse those who take the adjacent sentences and paragraph headings into consideration.

12) *The Automatic Pilot:* Others better qualified to judge will first consider the meaning of the word combination "automatic pilot." Then we are confident that they will not have difficulty in differentiating between the contributions of Sperry and of Hammond to the navigational art.

15) *The Proximity Principle:* Patents are written to be read and understood by experts. The terms used are to be interpreted as broadly as expert usage will permit. As regards this patent, the expert usage had been established for more than a decade. Thus, during the conference of a distinguished group of high officers with Hammond at Fort Monroe, August 23, 1918, the following was officially recorded. (Emphasis is by us.)

"General Squier: You have shown this afternoon the control of the *regular torpedo in water* both by the *radio device* and also by *acoustics*. Why do you limit yourself to water? Why don't you get an *aerial torpedo?*

"Mr. Hammond: I have covered that entirely from the standpoint of patents. The reason that I have not given it more thought is that I am specifically attacking the underwater section of capital ships. . . . I have not made any experiments; I have merely concentrated on this proposition for the Coast Artillery.

"General Squier: Is the *aerial torpedo* a harder problem?

"Mr. Hammond: It is harder to my mind."

Note that in a single paragraph, the Chief Signal Officer of World War I referred to two kinds of torpedoes and to two kinds of energy by which they may be controlled. Clearly the examiner of the proximity patent years later would have required Hammond to insert *underwater* before "torpedo" and *acoustic* before "energy," if he had intended to restrict the scope of the patent to the form set forth. It is inconsistent for our critic to assert that the Fiske claim terminating with the words "as set forth" is basic to the radio-control principle in forms not shown, and yet to deny that the Hammond claim cited in our paper is not basic to the proximity principle generally, with the words "as set forth" absent. In contrast with

* Received by the IRE, May 13, 1959.

the Fiske radio-control patent, the Hammond proximity patent did get reduced to practice, we understand, in both media. It was not generally known, presumably because no money passed in making it available, and there was no threat of a law suit.

16), 17), and 18). *The deForest Audion:* We feel that there is need for a formal syllogism:

> *Major Premise:* In writing the words "Audion. This form of detector . . . ," Marriott had in mind only one form of detector, since he used the word "this."
>
> *Minor Premise:* The form to which he referred as being used in 1912 was unquestionably the triode.
>
> *Conclusion:* The form to which he referred as being used in 1906 was therefore the triode.

Our critic is charging that Marriott was in error or that he did not express himself clearly. Since the patent application was signed in 1906, no one has the right to assume that the inventor did not practice his invention in 1906. The fact that it was not publicly disclosed until 1907 is irrelevant. Strangely, it is we who have to defend an author and his paper, when both were praised by our critic when the paper was first published.

As to the date of the patent, we corrected this formally and not by letter to the critic, because of the criticism, expressed by the exclamation point, that the Patent Office may have failed to consider it for a sufficient period of time. As far as we have observed, one would have to go back to the original Bell patent of 1876 to find an important patent which was issued after only three weeks of consideration.

22) *The Lowenstein Designation "Ion Controller":* We believe we understood the comment as written. An "alias" connotes another name without change of the object named. Lowenstein's device was a triode, and most certainly not an "audion" comprising a diode in series with a B battery, which is the only type of "audion" mentioned in the sentence. We doubt that Lowenstein's reasons for using "ion-controller" rather than "grid audion" were unethical. The word "audion" automatically signifies a detector in the patent and technical literature. Such a terminology was not sufficiently general for one with the superior knowledge that a triode was also useful for amplification and oscillation uses.

29) *Priority of Production of a High-Vacuum Tube.* Our critic finally invokes a 1931 statement of the Supreme Court in its termination of a General Electric suit against the de Forest company for alleged infringement of Langmuir's high vacuum patent 1,558,436 of 1925. This seemingly self-contradicting sentence was lifted from pertinent context, but upon consulting acceptable reports of all three courts involved, we have paraphrased the final decision as follows. "We do not question that Langmuir and Sweetser produced a 250 v, 5 ma high vacuum tube 'in which the current was limited by space charge, substantially independently of positive ionization,' in August, 1912. But by legal precedent we think that his legal date of conception of the patent claims based upon this work cannot be the date that he did it but rather the date that he knew that he had done it. We rule that Langmuir's legal date of conception was in late November of 1912. We further rule that Arnold's legal date of conception was November 14, 1912, despite de Forest's strange willingness to accept November 1. Therefore in November, 1912, Arnold anticipated Langmuir by about a week in conceiving the subject matter of the Langmuir claims. But the Federal Telegraph Company used a 54 volt amplifying triode in August, 1912, and a 67½ volt triode in November, thereby anticipating both Arnold's and Langmuir's legal dates of conception. Partly upon this Federal evidence, the Langmuir patent is hereby and irrevocably declared invalid and therefore not infringed. We feel it highly unnecessary to rule on the prior statement that Arnold's reduction to practice of the Langmuir claims was on April 25, 1913." If our interpretation of the Supreme Court decision is acceptable, our critic must agree that the Telephone Company work was much later than that of the General Electric in the production of high vacuum tubes.

In conclusion, it appears that our critic refused further comment on about eighty per cent of the points listed in our first rebuttal. We trust that he would refuse further comment on at least eighty per cent of the points treated again in this second response to his critiques. We note that his discussions have been offered "in the interest of more correct technical history." We hope the four papers of these discussions have done much to confirm the correctness of the technical history which we set forth in our September, 1957 paper.

Correspondence

Low-Noise Tunnel-Diode Amplifier*

Since Hull first disclosed the dynatron,[1] negative conductance amplifiers have received sporadic attention. As early in 1935, E. W. Herold pointed out the possibility of using negative conductance for amplification.[2] However, lack of convenient negative-conductance elements made such amplifiers unattractive. The purpose of this note is to report some results on a new negative-conductance amplifier using a novel semiconductor device called a tunnel diode[3] which was developed by H. S. Sommers at the RCA Laboratories.

The amplifier circuit is shown in Fig. 1. The tunnel diode D, having a capacitance C_d, is energized by a battery V_0 through a dc load resistance r_0. The resistance r_0 should be smaller than the negative resistance produced so that stable biasing is possible. The biasing point at which the negative conductance is realized is defined by the combined adjustment of the load resistance r_0 and the supply V_0. As shown in Fig. 1, the negative conductance is shunted by an RF tank which determines the amplifier resonant frequency f_1. C_b, a by-pass condenser in the tank, should be made as large a possible to prevent parasitic oscillations in the battery circuit. For stability, the RF load conductance presented by the combination of the generator conductance G_g and load conductance G_L through a tap transformation should be larger than the negative conductance (G) of the diode. Stable amplification can be achieved only when both dc and RF load conditions are fulfilled.

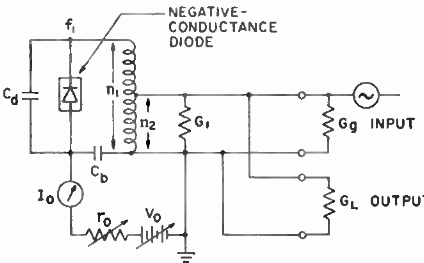


Fig. 1—Amplifier circuit using a negative-conductance diode.

Expressions for power gain (g_p), bandwidth (Δf), and noise factor (F) of Fig. 1 have been calculated; they are

$$g_p = \frac{4G_gG_L}{\left(G_T - \left(\frac{n_1}{n_2}\right)^2 G\right)^2} \tag{1}$$

$$B = \frac{1}{Q_L\sqrt{g_p}}\left[1 + \sqrt{1 + 4Q_L{}^2g_p}\right] - 2$$

$$\cong \frac{1}{Q_L\sqrt{g_p}}\left[1 + \frac{1}{4Q_L\sqrt{g_p}}\right] \quad (\text{if } Q_L\sqrt{g_p} \gg 1) \tag{2}$$

* Received by the IRE, May 1, 1959. This research was sponsored in part by the Electronic Res. Directorate, AF Cambridge Res. Center, under Contract AF19-(604)-4980.

[1] A. W. Hull "Description of the dynatron," PROC. IRE, vol. 6, p. 5; February 1918.

[2] E. W. Herold "Negative resistance and devices for obtaining it," PROC. IRE, vol. 23, p. 1201; October, 1935.

[3] H. S. Sommers, "Tunnel diodes as high frequency devices," PROC. IRE, (this issue, p. 1201).

$$F = 1 + \frac{T}{T_0}\left[\frac{G_1}{G_g} + \frac{G_L}{G_g} + \frac{G_e}{G_g}\right] \quad (3)$$

where

$$G_T = G_1 + G_g + G_L \quad (4)$$

$$Q_L = \frac{\omega_1 C_d}{2\sqrt{G_g G_L}}\left(\frac{n_1}{n_2}\right)^2 \quad (5)$$

$$G_e = \frac{eI_0}{2kT}\left(\frac{n_1}{n_2}\right)^2 \quad (6)$$

$$B = 2\frac{\Delta f}{f_1}. \quad (7)$$

It is assumed that the primary noise in the negative-conductance diode is shot effect. Ge is the equivalent transformed noise conductance due to the shot effect of the diode which has a dc current I_0. T_0 is the reference temperature of the generator conductance, and T is the ambient temperature.

An experimental lumped circuit based on Fig. 1 has been built. The operating frequencies were 80 mc, 66 mc, and 30 mc. They all gave stable gains of about 20 db. Representative experimental results at 30 mc are shown in Table I. The computed gains, bandwidths, and noise factors are based on (1)–(3), using values of $G_g = 0.02$ mhos, $C_d = 40$ $\mu\mu$f, and $(n_1/n_2)^2 = 7.65$. These values were used in the experimental amplifier. It is noted that these computed results agree reasonably well with those measured.

1) Both amplifiers achieve gain through a negative conductance. In the nonlinear-susceptance amplifier, the negative conductance is derived from an RF pump through a nonlinear interaction, while in the negative-conductance amplifier, the negative conductance is directly realized from a dc source.

2) By lowering the idling circuit Q's of a nonlinear-susceptance amplifier, one can increase the bandwidth, as shown by (9). However, the increase of bandwidth in this way is accompanied by an increase in the pumping power. This strong limitation is not present in the negative-conductance amplifier where the loaded Q_L is not subject to any limitation.

3) At first thought, one might well assume that the negative conductance amplifier, because of the shot noise, will generate more noise than the nonlinear-susceptance amplifier. This is not necessarily so according to (3). While the negative-conductance amplifier has noise originating from shot effect, the nonlinear susceptance amplifier has an equivalent noise due to the presence of the idling circuit. In practice, the idling-circuit noise in the latter case can be minimized by choosing a very high idling frequency (ω_2) compared to the signal frequency (ω_1). Analogously, the shot-effect noise in the former case can also be reduced by using a source conductance (G_g) which

TABLE I
EXPERIMENTAL RESULTS

I_0	G	G_L	g_p		$2\Delta f$		N.F.	
Diode current (μa)	Negative conductance (mho)	Load conductance (mho)	Power gain (db)		Bandwidth (mc)		Noise factor (db)	
			measured	computed	measured	computed	measured	computed
250	$-\frac{1}{375}$	$\frac{1}{1000}$	20	23	0.2	0.3	4.5	4.7
300	$-\frac{1}{310}$	$\frac{1}{200}$	40	36	0.19	0.16	6.3	4.5
350	$-\frac{1}{206}$	$\frac{1}{50}$	27	26	0.8	1.05	8.0	6.8

The amplifier has much broader bandwidth at low gains. For instance, typical measured bandwidths at 10 db gain are of the order of 3 or 4 mc. According to (2), the bandwidth varies inversely as the voltage gain at high values of circuit Q's.

It is interesting to note that the negative-conductance amplifier has a striking resemblance to the nonlinear-susceptance amplifier[4] as far as the gain, bandwidth, and noise factor are concerned.

For the nonlinear-susceptance amplifier, it has been shown that:

$$g_p = \frac{4G_g G_L}{(G_T - G)^2} \quad (8)$$

$$B \cong \frac{\omega_2}{\omega_1}\frac{1}{\sqrt{g_p}}\frac{1}{Q_2} \quad (9)$$

$$F = 1 + \frac{T}{T_0}\left(\frac{G_1}{G_g} + \frac{G_L}{G_g} + \frac{\omega_1}{\omega_2}\frac{G}{G_g}\right). \quad (10)$$

By comparing (8)–(10) with (1)–(3), it follows that:

[4] S. Bloom and K. K. N. Chang, "Theory of parametric amplification using nonlinear reactances," *RCA Rev.*, vol. 18; pp. 578–593; December, 1957.

is higher than the equivalent noise conductance (G_e) [see (3)]. However, the negative-conductance amplifier has the advantage of using a dc source and design of a high conductance source seems quite feasible.

The diodes which were used in the experimental amplifier happen to be low negative conductance diodes. According to (3) a low ratio of current to negative conductance will give a low noise factor. It is quite conceivable that a diode can be made with a negative conductance $G = -0.02$ mho, at a diode current $I_0 = 200$ μa, giving a conductance ratio $G_e/G_g = 0.20$, and hence, a noise factor of the order of a few tenths of a db, regardless of the operating frequency.

The present experiments have demonstrated the principle of low-noise amplification using the new tunnel diodes. While the results obtained are for the 30–100 mc range, there seems to be no reason why this principle cannot be used to obtain low-noise amplification in the microwave region. Diodes for this purpose should have the properties of low conduction current and high negative conductance.

It is a pleasure to thank Dr. H. S. Sommers, who developed the negative-conductance diodes, for many enlightening discussions.

K. K. N. Chang
RCA Labs.
Princeton, N. J.

Superregenerative Reactance Amplifier*

The superregenerative amplifier is characterized by the repeated build-up and decay of self-oscillations on or near a signal frequency. In the ordinary regenerative amplifier, the limit of gain is reached when the positive feedback is increased to the point where the amplifying device breaks into oscillation.

By allowing the circuit to oscillate for a fraction of the time, one may extend operation into the region of oscillation, thereby greatly increasing the gain of the ordinary regenerative amplifier.

An L-band variable reactance amplifier with a lower sideband regenerative gain of 17 db and a bandwidth of 3 mc has exhibited a gain of 72 db with a slight increase in bandwidth when operated as a superregenerative amplifier. Signal frequency was 1450 mc. Pumping was done at 10,150 mc. The over-all receiver noise figure was approximately 5 db, as determined by a stable minimum discernible signal level of -104 dbm. The results obtained indicate the feasibility of a single-stage low-noise microwave receiver.

The receiver in all the experiments was essentially composed of a Bell Telephone Laboratories mesa type silicon varactor diode, an MA410 crystal detector, a Tektronix 531 oscilloscope, and a 2K-39 Klystron "pumping" source. A schematic of the circuit is shown in Fig. 1. The varactor diode

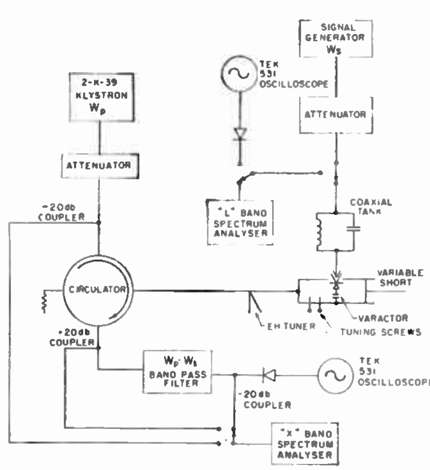


Fig. 1—Schematic of amplifier circuit.

* Received by the IRE, April 24, 1959.

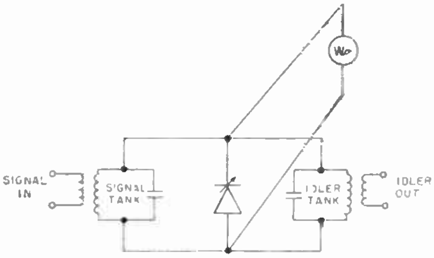
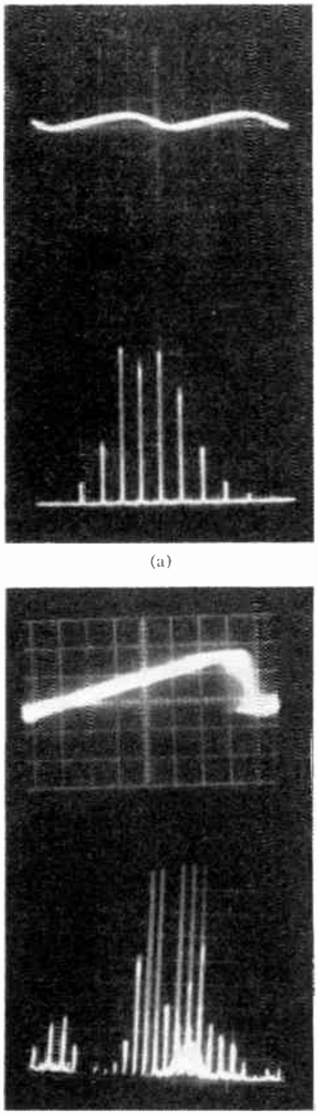


Fig. 2—Equivalent circuit showing the capacitor diode as a part of two resonant tanks.



(a)

(b)

Fig. 3—(a) 2.5 mc self-quench recurrence frequency and its corresponding spectrum. (b) 1.25 mc self-quench recurrence frequency and its corresponding spectrum.

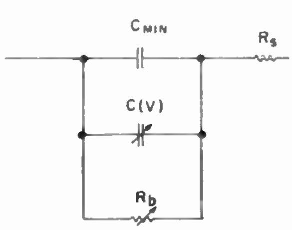


Fig. 4—Equivalent circuit of a varactor diode.

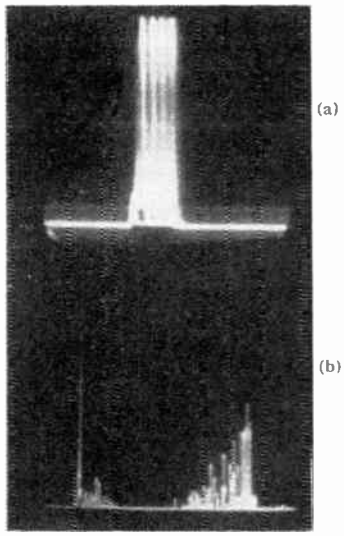


Fig. 5—(a) Relaxation oscillations occurring when a 10 μsec signal pulse is present. (b) Relaxation oscillations in idler tank.

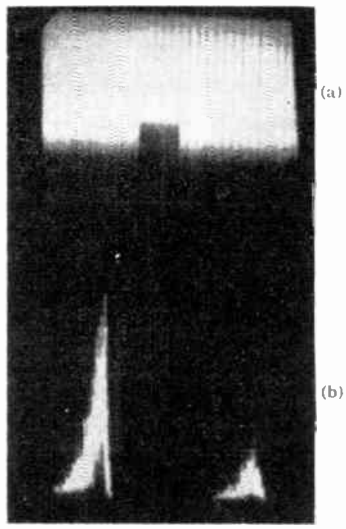


Fig. 6—(a) A −90 dbm signal pulse damping the relaxation oscillations. (b) Relaxation oscillations in idler tank.

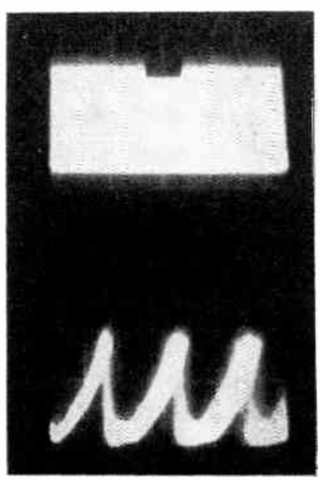


Fig. 7—Quench waveform and corresponding output for a 10 μsec, −90 dbm signal pulse.

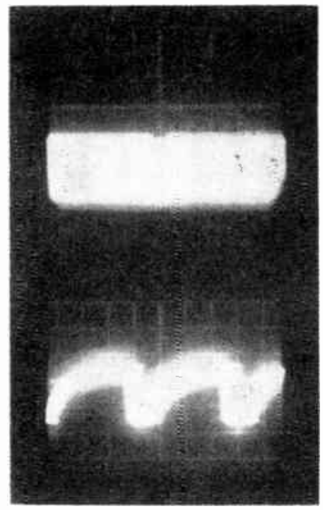


Fig. 8—Quench waveform and corresponding output for a 10 μsec, −104 dbm discernible signal pulse.

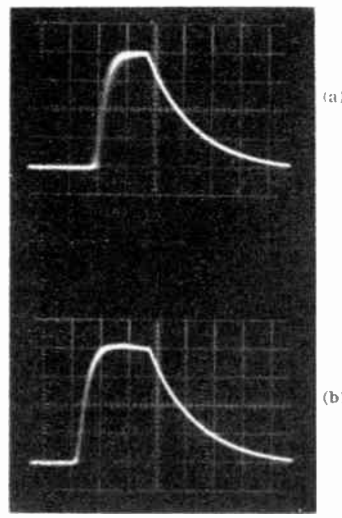


Fig. 9—(a) Oscillation envelope with no signal present. (b) Oscillation envelope with signal present.

was operated without a dc return in order to have a hole storage effect and therefore obtain self-biasing. The diode had a zero bias capacitance of 1.60 $\mu\mu$f and a series resistance of 2.78 ohms. The cutoff frequency of the diode was 81.0 kmc.

An equivalent circuit of the receiver is shown in Fig. 2. The circuit could be made to oscillate simultaneously at the seventh subharmonic frequency, $W_p/7$, and at the resonant frequencies of the signal and idler tanks. The tank oscillations could be varied approximately 100 mc by tank tuning. Under certain conditions, when the resonant oscillating frequency of the signal tank approached the subharmonic frequency, a periodic relaxation of all the oscillations occurred (Fig. 3). The decay time of this self-quench waveform was reasonably constant at approximately 0.10 μsec. This is the RC time constant one would expect from an average varactor diode (Fig. 4).

It appears that the diode is acting in a manner similar to a squegging triode oscillator. The amplitude and period of the quench waveform was inversely proportional to pump power. As the oscillations build up, the self-bias voltage across the diode and consequently the diode capacity, varies. The change in capacity detunes the tank circuits, causing the oscillations to die out.

The oscillations could be made to build up from the decaying oscillations of the preceeding cycle of quench.

An incompletely quenched oscillation resulted in a great reduction of sensitivity. The percentage modulation of the oscillations could be varied by tank tuning. In some instances, the observed spectrum indicated an over modulation effect.

The amplifier was first operated with a signal at the subharmonic frequency injected into the subharmonic tank. A signal pulse length of 10 μsec was used. By either varying pump power or the tuning of the idler tank, the circuit could be placed on the threshold of relaxation oscillations. At this point, the circuit oscillates continuously, with no quenching. The presence of the signal pulse causes the oscillations to increase and to quench (Fig. 5).

At the end of the signal pulse, the tanks return to their normal oscillatory state (no relaxations). This type of amplifier differs from the normal superregenerative amplifier in that the oscillations have a definite threshold value of -85 dbm. The self-quench frequency was dependent upon signal level. Since the superregenerative receiver can only detect modulation frequencies lower than the quench frequency, the threshold values will vary with pulse width. The output of this amplifier was always constant in amplitude and polarity regardless of signal strength. It should be noted that this type of amplifier can be used as a threshold device or as a limiter.

The same amplifier was operated in a self-quenched super-regenerative mode by allowing the relaxation oscillations to occur continuously (Fig. 6). The presence of the 10 μsec signal pulse at the signal tank oscillating frequency caused the relaxations to decrease in amplitude. A minimum discernible signal of -104 dbm was measured at a signal bandwidth of 3 mc, indicating an over-all receiver noise figure of approximately 5 db. The damping effect appeared linear with signal strength through a dynamic range of 70 db. No apparent change in quench frequency was observed for signals less than -30 dbm. However, for signals greater than -30 dbm, a variation in quench frequency seemed to occur. The amount of noise that appears at the peaks of oscillation depends upon the particular quench waveform. Figs. 7 and 8 indicate that a sharp waveform is noisier than the exponential waveform.

The amplifier was operated in an externally quenched superregenerative mode by pulsing the pump for one microsecond pulses at a PRR of 2000 cps. The presence of CW signal frequency in the signal tank coalescing with the free tank oscillations causes the oscillations to build up earlier, as shown in Fig. 9. This type of amplifier seems to have more sensitivity and is definitely easier to tune than the self-quenched amplifier.

Bernard B. Bossard
U. S. Army Signal Res. and Dev. Lab.
Ft. Monmouth, N. J.

Parametric Amplifiers as Superregenerative Detectors*

Parametric amplifiers can be made to operate as superregenerative detectors. We have investigated superregeneration in two cavity-type parametric amplifiers that use semiconductor diodes. The experiments have involved both self-quenched and separately-quenched operation. One of the amplifiers operating separately quenched at L band has a noise figure of 1 db, a stable gain of 56 db and a bandwidth of 2 mc. The quench frequency for this operation is 0.25 mc. The other amplifier, which operates at S band, gives a bandwidth of 12 mc and a stable gain of 35 db when self-quenched at a quench frequency of 56 mc. The gain of both amplifiers can be increased to well over 80 db, but the other characteristics change as a consequence. The purpose here is to describe briefly the details of the superregenerative operation and to explain qualitatively some of the effects that permit self quenching.

Fig. 1 illustrates the L-band amplifier, which amplifies over a small range of frequencies centered at 780 mc. The pump frequency is 10 kmc and an idler frequency is generated at approximately 9.22 kmc. When the parametric diode is operated with no dc return, a curious charge and discharge behavior can be produced which results in a periodic self-quenching of oscillations within the device, and consequently permits superregenerative amplification.

With the proper adjustment of parameters the behavior is typical of self-quenched superregenerative operation. For example, the quench frequency is a linear function of the logarithm of the signal strength. This is illustrated in the plot of Fig. 2. An oscilloscope photograph of the video output showing the envelope of the self-quenching oscillations is presented in Fig. 3. The photograph is a double exposure showing two traces. The trace having the longer period corresponds to no signal at the input. It appears blurred due to the fact that noise at the signal frequency triggers the build-up of the oscillations. When a CW signal greater than the noise is present the resulting pattern is illustrated by the other trace. The oscillations are then triggered by a coherent source which removes the blurring, and the oscillations build up from a higher initial level, which results in an earlier build-up and therefore a higher quench frequency. By operating in this fashion the amplifier will detect pulsed signals as small as 114 db below 1 mw with a stable gain of 85 db and a bandwidth of 1.5 mc.

The amplifier can be operated also with external quenching by modulating the frequency of the klystron pump. In this fashion the pump frequency sweeps through the operating range of the amplifier and oscillations occur in short bursts. In the absence of a signal the oscillations build up from noise and the output is then a train of noise modulated pulses at the quench frequency. The energy varies from pulse to pulse in a random manner. The time allowed for the oscillation build-up is not sufficiently long for the oscillation level to reach saturation so the presence of a signal causes the energy per pulse to increase, resulting in a corresponding increase in the detected output.

* Received by the IRE, April 24, 1959.

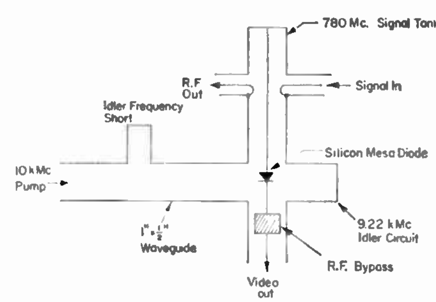


Fig. 1—Diagram of the L-band parametric amplifier.

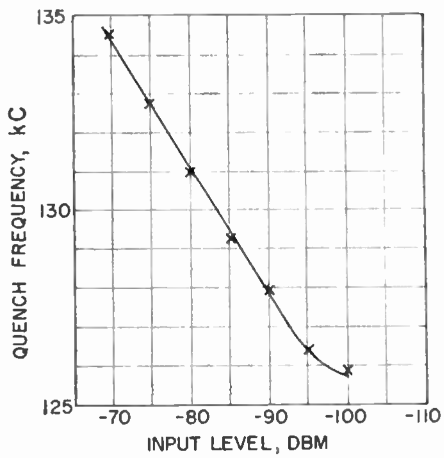


Fig. 2—Quench frequency as a function of input signal level.

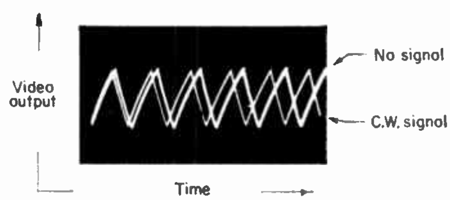


Fig. 3—Photograph of oscilloscope traces showing the video outputs when no signal is present and when a signal is present.

This type of superregenerative operation gives constant gain over a 40-db range of input power level. With a quench frequency of 0.25 mc the noise figure is 1 db and the bandwidth 2 mc. At higher quench frequencies the noise figure deteriorates, increasing to 2 db for a quench frequency of 1 mc. The bandwidth improves with higher quench frequencies increasing to 4 mc for the 1 mc quench frequency. The gain remains at 56 db for both quench frequencies.

The super-regenerative operation gives both higher gain and higher bandwidth than does the conventional type of parametric operation. For example, this amplifier, operated conventionally, can give a gain only as high as about 23 db before instabilities set in. The bandwidth corresponding to that value of gain is only 150 kc. Also, for super-regenerative operation the amplifier appears less sensitive to changes in the input impedance, the pump power and the frequency than for the conventional operation.

The S-band amplifier receives pump energy at 6 kmc and amplifies signals at frequencies around 3 kmc. Fig. 4 shows a diagram of the amplifier. Most of the experiments performed with it have involved self-quenched operation. Much of the behavior observed has been quite different from that usually associated with self-quenched superregenerative operation. For example, the amplifier can be made to indicate the presence of an input signal by either a substantial decrease or a substantial increase in the level of the oscillations. In the oscilloscope photograph of Fig. 5 the video output is shown to decrease in the presence of an input signal. An input signal pulse of one microsecond duration appears at about the seventh cycle of the video pattern. The quench frequency in the absence of a signal was approximately 1 mc. The peak power of the input pulse was 80 db below 1 mw.

Fig. 6 shows just the opposite effect, an increase in the oscillation level with the application of an input signal. The photograph shows the oscilloscope pattern of the video output. Both the time scale of the abscissa and the amplitude scale of the ordinate have been greatly reduced from the conditions of the previous figure. Pulses of input signal are clearly delineated by the very large increase in the level of oscillation. For this operation the peak pulse power for the input signal was 100 db below 1 mw, the pulse width was 5 μsec and the quench frequency was 2 mc.

Other experiments with the S-band amplifier clearly demonstrated that bandwidth can be increased by sacrificing gain. Table I lists the power gain and corresponding bandwidth obtained with the amplifier for self-quenched operation. The changes of gain and bandwidth were brought about by making certain adjustments of amplifier parameters such as pump power level. As the gain decreased and the bandwidth increased there was also a general increase in the quench frequency. The quench frequency corresponding to the last entry in the table was 56 mc.

It is beyond the scope of this letter to describe in detail the processes responsible for the self-quenching operation. However, we will point out qualitatively some of the effects which are present and do play a part in the quenching mechanism. We have observed experimentally that the periodic growing and quenching of the oscillations is synchronized with periodic charging and discharging of the parametric diode. For self-quenched operation the diode is self-biased and has no dc return. The value of the bias changes periodically at the quench frequency. As the bias changes, the level of oscillation changes in synchronism.

For a qualitative explanation of what takes place consider the S-band amplifier which operates in the degenerate mode. A change in the oscillation level can obviously result from a change in the tuning of the tank. When the tank is tuned at precisely one-half the pump frequency the conditions are best for large oscillations. The tuning of the tank can be changed by changing the average capacitance of the parametric diode. Two effects will produce a change in the average diode capacitance: 1) a change in the value of the self-bias (the average capacitance decreases as the self-bias becomes more negative); 2) a change in the level of the voltage swing about the self-bias point (the average capacitance increases as the level of voltage swing increases). These two effects coupled with inertia in the build-up of the oscillations combine to play an important part in the mechanism for self-quenching.

Imagine that at the time the oscillations start to build up the tank is tuned below one-half the pump frequency. As oscillations build up, the total voltage swing across the parametric diode increases. This drives the bias further negative tending to reduce the average capacitance (effect number 1 of the two listed above) and therefore raise the frequency to which the tank is tuned. However, the build-up in the oscillations produces the opposite effect on the average capacitance (effect number 2) and hence tends to slow down this action. Gradually there is an improvement in the tuning and eventually the tuning is correct for maximum oscillations. Because of the inertia, the build-up continues with further reduction in the average capacitance and the condition for optimum tuning is passed. Oscillations eventually start to decrease. As the oscillations decrease the average capacitance continues to go down (effect number 2) resulting in further detuning of the tank. Consequently a snowballing takes place. The decrease in the oscillations brings about further detuning which decreases further the oscillations. This can produce the sudden quenching which is characteristic of much of what we have observed. When the oscillations are quenched the initial bias condition is eventually restored and the cycle begins anew.

Obviously the above explanation is far from complete. It represents a part of the mechanism responsible for part of the observed operation. It seems to fit rather well the operation illustrated by Fig. 3. It certainly does not account for the operation illustrated in Figs. 5 and 6. Our search for the remaining effects is continuing.

We would like to express our gratitude to R. Hanbury-Brown of Jodrell Bank, England, and to B. Bossard and E. Frost of the Signal Corps Laboratory, Ft. Monmouth, N. J. Their influence was significant in this work. Mr. Hanbury-Brown first suggested to us that superregenerative operation of a cavity-type parametric amplifier might result in more stable gain. Messrs. Bossard and Frost communicated freely with us the results of their experiments on self-quenched superregeneration in a parametric amplifier (described in the accompanying letter). Their work greatly stimulated our interest in this matter.

J. J. Younger
A. G. Little
H. Heffner
G. Wade
Stanford University
Stanford, Calif.

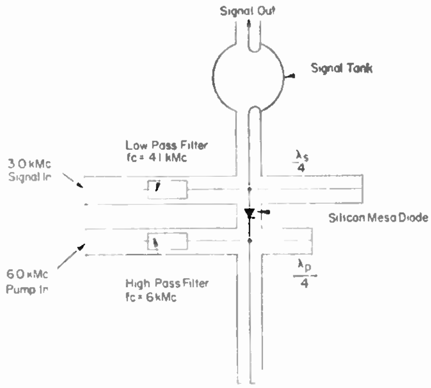


Fig. 4—Diagram of the S-band parametric amplifier.

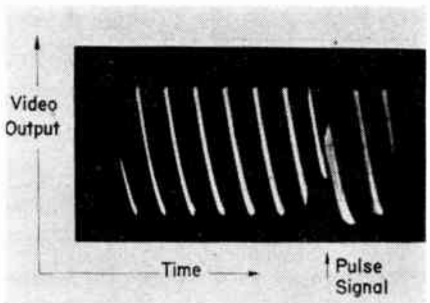


Fig. 5—Photograph of oscilloscope trace showing the video output when a pulsed signal is present.

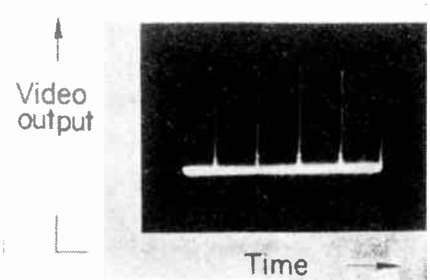


Fig. 6—Photograph of oscilloscope trace showing an increase in the oscillation level due to the presence of a signal pulse.

TABLE I

Gain and Bandwidth Data for S-Band Amplifier under the Conditions of Self-Quenched Operation

Power Gain (db)	Bandwidth (mc) for −3 db points
87	1.2
80	1.8
76	2.2
57	4.0
41	5.2
35	12.0

Comment on a Result of L. Joseph and W. K. Saunders*

In a recent letter,[1] the following observation was made. Let S be an $n \times n$ normalized scattering matrix and suppose that $Q = 1_n - SS^* \geq 0$, *i.e.*, $Q = Q^*$ is the matrix of a non-negative quadratic form. Then, each column of S has Euclidean length less or equal

* Received by the IRE, January 29, 1959.
[1] L. Joseph and W. K. Saunders, "A theorem on lossy n-port junctions," Proc. IRE, vol. 47, p. 102; January, 1959.

to unity ($\tilde{A}$, $\bar{A}$ and A^* denote the transpose, the complex conjugate and the complex conjugate transpose of the matrix A respectively). The Euclidean length $l(\mathbf{a})$ of an n-vector $\mathbf{a}$ is given by

$$l(\mathbf{a}) = \sqrt{\mathbf{a}^*\mathbf{a}}$$

and 1_n is the unit $n\times n$ matrix. The object of the present note is to show that their interesting result is a consequence of a more general theorem whose proof depends on the fact that for two arbitrary $n\times n$ matrices A and B, AB and BA possess the same eigenvalues.[2]

Theorem:
If

$$Q = 1_n - SS^* \geq 0,$$

then

$$Q_1 = 1_n - S^*S \geq 0.$$

Proof: Since SS^* and S^*S have the same eigenvalues, Q and Q_1 have the same eigenvalues. By hypothesis those of Q are all non-negative. Thus $Q_1 \geq 0$, Q.E.D.

Now $Q_1 \geq 0$ if and only if all the coaxial minors of Q_1 are non-negative.[2] In particular the diagonal elements

$$(Q_1)_{rr} \geq 0, \qquad (r = 1, 2, \cdots, n).$$

If $\mathbf{a}_r$ denotes the rth volumn of Q_1, an easy calculation yields

$$(Q_1)_{rr} = 1 - \mathbf{a}_r^*\mathbf{a}_r, \qquad (r = 1, 2, \cdots, n),$$

whence

$$\mathbf{a}_r^*\mathbf{a}_r \leq 1, \qquad (r = 1, 2, \cdots, n).$$

The above theorem admits a very neat network interpretation: $\tilde{S}(p)$, $p=\sigma+j\omega$, is the scattering matrix of a linear passive n-port if and only if $S(p)$ is the scattering matrix of a linear passive n-port.

D. Youla
Microwave Res. Inst.
Polytechnic Inst. of Brooklyn
Brooklyn 1, N. Y.

[2] L. Mirsky, "An Introduction to Linear Algebra," Clarendon Press, Oxford, Eng.; 1958.

A Simple Measurement for Transistor Current Gain in Magnitude and Phase*

This note describes a simple method of determining transistor current gain directly in phase and magnitude from modulus measurements of the common-base current gain α and the common-emitter current gain $\beta=\alpha/(1-\alpha)$. Although the procedure described is almost self-evident, discussions with other workers in the field suggest that it is not common knowledge and may not have been proposed previously.

The method is based on the principle, indicated in Fig. 1, that the position of α on

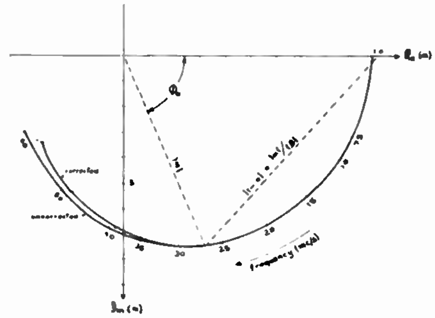


Fig. 1—Example locus for a 2N247 drift transistor measured by this method; the geometric principle is indicated.

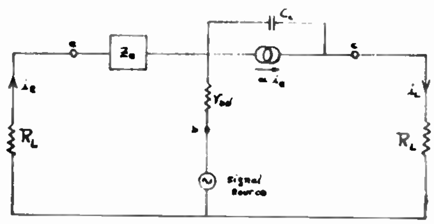


Fig. 2—Measurement circuit.

its frequency locus is known if the magnitudes

$$|\alpha|$$

and

$$|1-\alpha| = \frac{|\alpha|}{|\beta|} \tag{1}$$

are known. Similarly, the position of β on its frequency locus is known if the magnitudes

$$|\beta|$$

and

$$|1+\beta| = \frac{1}{|1-\alpha|} = \frac{|\beta|}{|\alpha|} \tag{2}$$

are known. Thus, the loci of α and β can be constructed by geometry from measurements of $|\alpha|$ and $|\beta|$ with frequency.

A sample locus found by this method is drawn in Fig. 1 for a 2N247 drift transistor, the measurements for which were carried out using VHF circuit techniques developed by Das and Boothroyd.[1] For the $|\alpha|$ measurement, the transistor is excited at the base inducing the two currents of interest to flow at the emitter and collector. (See Fig. 2.) The ratio is taken between voltages developed across similar small loads R_L. With the transistor turned around in the circuit (excitation applied at the emitter), $|\beta|$ is measured in a similar way with suitable rearrangement of the transistor bias supplies (not shown in Fig. 2).

If the two loads were identical impedances, and if current flow in C_c were negligible, the ratio of the two measured voltages would be $|\alpha|$ (or $|\beta|$). Because of the presence of C_c, the measured ratios are over-all terminal current gains $|\alpha_T|$ and $|\beta_T|$; it can be shown that the true values $|\alpha|$ and $|\beta|$ are related to the measured over-all values as follows:[2]

$$|\alpha| = |\alpha_T|\sqrt{\frac{1+(R_L\omega C_c)^2}{1+\dfrac{2nR'\omega C_c+(R'\omega C_c)^2}{|\alpha|^2}}} \tag{3}$$

$$|\beta| = |\beta_T|\sqrt{\frac{1+\dfrac{2nR_L\omega C_c+(R_L\omega C_c)^2}{|1-\alpha|^2}}{1-\dfrac{2nR''\omega C_c-(R''\omega C_c)^2}{|\alpha|^2}}} \tag{4}$$

where

R_L = the load resistance
$R' = R_L+\mathrm{Re}(Z_c)$
$R'' = R_L+r_{bb'}$
$n = \mathrm{Im}(\alpha)$.

In practice, the loads are not identical. To eliminate any error from this source, an extra measurement may be made of; a) $|\alpha_T|$, with the transistor emitter and collector terminals reversed, b) $|\beta_T|$, with the transistor base and collector terminals reversed. If the resulting four measured current gains are designated $|\alpha_T|_1$, $|\alpha_T|_2$, $|\beta_T|_1$, and $|\beta_T|_2$, it can be shown that the true over-all current gains are given by[3]

$$\alpha_T = \sqrt{|\alpha_T|_1\,|\alpha_T|_2}$$
$$\beta_T = \sqrt{|\beta_T|_1\,|\beta_T|_2}. \tag{5}$$

Eqs. (3) and (4) are evaluated from an initial plot for α using the measured over-all values $|\alpha_T|$ and $|\beta_T|$. If the correction is large it may be re-evaluated from a corrected plot for α, and may then again be calculated iteratively until the true α is obtained. For the drift transistor example of Fig. 1, both the uncorrected and corrected loci are shown, for load (R_L) values of 10 ohms. The corrections are seen to be negligible at frequencies below 30 mc; on the other hand, above 50 mc they are large enough to require iterative evaluation—a tedious process.

A limitation to the method is evident for frequencies at which the phase ϕ_α approaches 180°, and where the arcs of radius $|\alpha|$ and $|1-\alpha|$ approach concentricity. Here, the small errors in the uncorrected values for $|\alpha|$ and $|\beta|$ produce a large error in the initial plot for the locus position, and the resulting corrections must be applied iteratively many times before the locus position is completely determined. In spite of this difficulty at certain frequencies, the method has been found extremely useful, permitting the accurate determination of α and β over a wide range of frequencies with the use of very simple apparatus.

D. F. Page
A. R. Boothroyd
Dept. of Electrical Engineering
Imperial College
London, England

* Received by the IRE, February 9, 1959; revised manuscript received, March 6, 1959.

[1] M. B. Das and A. R. Boothroyd, "Measurement of equivalent circuit parameters of transistors at V.H.F.," forthcoming paper in *Proc. IEE*.

[2] It is to be noted that the signal generator impedance does not enter the analysis.

[3] This use of the geometric mean was suggested by Das and Boothroyd, *op. cit.*

Alternative Detection of Cochannel FM Signals*

H. W. Farris' comment on our correspondence[1] concerning his paper[2] contains statements that invite further discussion.

First, it is not true that " . . . any nonlinear device will operate to the detriment of the weaker signal which is sought." What is known is that certain nonlinear operations, when followed by appropriate filtering, will lead to the suppression of the weaker signal; others will actually boost the weaker signal in certain spectral zones and will leave the relative amplitudes of the signals unaffected in others. Moreover, the statement that, "it is well known that nonlinearities will lead to the weak-signal suppression effect" is misleading and irrelevant. What is usually meant by the "weak-signal suppression effect" of nonlinear devices is something that concerns a signal embedded in random noise of certain properties. That this situation is irrelevant here follows from the fact that if the signal is too weak to override the random noise, then neither the AGC system nor the simple feedforward system that is under discussion will be of much use.

Next, the statements about AGC need some clarification. Of course, an appropriate AGC system is in the category of what I termed "suitable means to regulate the level of the signal." However, the type of AGC system that will insure proper performance for slow variations in signal level will not handle as wide a range of important interference conditions as a simple narrow-band limiter would. AGC systems that simulate narrow-band limiters are more advantageous than the quasi-linear variety under certain interference conditions, but they raise the weaker-signal capture threshold. When the AGC system begins to simulate a wideband limiter (as far as the sum of the two incoming signals is concerned), no weaker-signal capture will be achieved.

In operation under field conditions, it is extremely important that the properties of the signal at the output of the feedforward system be independent of the input signal level. Although the use of a narrow-band limiter accomplishes this result at the expense of some loss of weaker-signal capturability, the gain in reliability of capture performance that ensues (in operation under a wide variety of interference conditions) will often more than offset the (maximum of 6 db) loss in the threshold of weaker-signal capture. In any case, an understanding of the principles involved and of the conditions under which the receiver must operate will always be the best guide to the desired choice of signal-level regulator.

Finally, it must be noted that the most general block diagram of a feed forward system does admit nonlinear operations in *both* channels, but it should be clear from the description of the feedforward mechanism that this is not a "requirement." In fact, the simple connection of a linear amplifier in feedforward across a limiter, when it was contrasted with the feedback, connection, suggested "feedforward across the limiter" as a title[3] to this writer.

It is somewhat surprising that a misinterpretation of my writings has led Farris to believe that he had a different and independent development in the "alternative method," especially since the system was fully described in a report[4] (exclusively devoted to the weaker-signal capture problem), addressed to him, which he had received a month and a half before his own date of "independent conception" of the idea. Farris' conclusion that "experience in the laboratory" showed the linearity of one of the paths "to be necessary" attributes the failure of his experiment to the wrong cause.

Elie J. Baghdady
Res. Lab. of Electronics
Mass. Inst. Tech.
Cambridge, Mass.

* Received by the IRE, February 17, 1959.

[1] Proc. IRE, vol. 47, p. 994; May, 1959.

[2] Proc. IRE, vol. 46, pp. 1876–1877; November, 1958.

[3] E. J. Baghdady, "Feedforward Across the Limiter," Quart. Prog. Rep., Res. Lab. of Electronics, M.I.T., Cambridge, Mass., pp. 52–53; October 15, 1957.

[4] E. J. Baghdady, Classified Quart. Prog. Reps. on Contract DA-36-039sc-73195, General Electronic Labs. Inc., Cambridge, Mass.; May, 1957, September, 1957, December, 1957, *et seq.*

A Possible Mechanism for Radiation and Reflection from Ionized Gas Clouds*

The traditional assumption in electromagnetic theory[1] that an initial distribution of charge density in a homogeneous conducting medium decays exponentially, and so may be set equal to zero, no longer holds if gradients of conductivity or dielectric constant are present. In that event, a charge distribution can be maintained in the presence of an electric field as we shall show below, and if the electric field is oscillatory, so also is the charge distribution. As in the case of an antenna, an oscillating charge distribution will radiate, so that we have another possible mechanism for radiation and re-radiation from auroras, meteor trails, and the like.

The concept will first of all be illustrated by a vastly over-simplified analysis, and afterwards directions in which it may be generalized will be discussed.

We shall assume not only the divergence condition and equation of continuity of Maxwell's equation, but also, going further, that a form of Ohm's law holds, so that in some sense,

$$J = \sigma E. \tag{1}$$

We then have, from the equation of continuity,

$$\nabla\sigma \cdot E + \sigma\nabla \cdot E + \frac{\partial\rho}{\partial t} = 0. \tag{2}$$

But from the divergence condition,

$$\nabla \cdot E = \rho/\epsilon - \nabla \ln \epsilon \cdot E. \tag{3}$$

Inserting this result into (2) yields the following differential equation in $\rho(r, t)$:

$$\frac{\partial\rho}{\partial t} + \frac{\sigma}{\epsilon}\rho = \sigma\nabla \ln \frac{\epsilon}{\sigma} \cdot E. \tag{4}$$

If σ and ϵ are functions only of the spatial coordinates r, not of t, and if E varies in time as $e^{i\omega t}$, this equation can be integrated by elementary means to yield the solution

$$\rho(r, t) = \epsilon \frac{\nabla \ln \dfrac{\epsilon}{\sigma} \cdot E(r)}{1 + i\omega \dfrac{\epsilon}{\sigma}} e^{i\omega t}. \tag{5}$$

We note from this expression that the charge density will oscillate in tune with the electric field whenever there exist non-vanishing gradients of conductivity and dielectric constant. The electric vector E in (5) represents the total field, consisting of the impressed as well as the true field [or current, since the above expression can also be written under the assumption (1) as $\rho = (\epsilon/\sigma)\nabla \ln (\epsilon/\sigma) \cdot J(r)e^{i\omega t}/(1 + i\omega\epsilon/\sigma)$].

We now consider some generalizations of the above approach. First suppose that the conducting medium is moving with velocity v. Then (1) must be replaced by

$$J = \sigma(E + v \wedge B). \tag{6}$$

If we now assume as before that σ, μ and ϵ are functions of position only and, in addition, that the velocity v is constant and B varies harmonically,

$$B = B(r)e^{i\omega t},$$

we thus obtain a differential equation similar to (4), which can be integrated to yield

$$\rho(r, t) \sim \frac{\epsilon e^{i\omega t}}{1 + i\omega \dfrac{\epsilon}{\sigma}} \left[\nabla \ln \frac{\epsilon}{\sigma} \cdot E(r) - \nabla \ln \sigma \cdot v \wedge B(r) + \frac{\sigma + i\omega\epsilon}{\mu} v \cdot E(r) \right] \tag{7}$$

which again represents an oscillating charge density.

Let us now consider the case when an ambient magnetic field is present, as in the case of a magneto-ionic medium. In that event, the conductivity $\boldsymbol{\sigma}$ is a tensor, and

* Received by the IRE, March 30, 1959.

[1] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, N. Y., p. 15, p. 222; 1941.

the relation (1) must be written as the dyadic inner product

$$\boldsymbol{J} = \boldsymbol{\sigma} \cdot \boldsymbol{E}. \tag{8}$$

It no longer appears possible to determine any simple differential equation such as (4) for the charge density ρ.

The case when the electron density N_e and collision frequency ν of the ionized medium vary harmonically with time:

$$\omega_e^2 \propto N_e \propto e^{i\Omega t}; \quad \nu \propto PT^{-1/2} \propto e^{i\Omega t}, \tag{9}$$

as in a pressure oscillation of the medium, is essentially different from the situation considered above. Suppose further that $\omega \ll \nu$, where ω is the angular frequency, and that the constitutive parameters ϵ and σ vary only with time. We shall then have

$$\epsilon(r, t) = \tilde{\epsilon} e^{-i\Omega t} \sim 1 - \frac{\omega_e^2}{\nu^2} \tag{10}$$

$$\sigma(r, t) = \tilde{\sigma} \sim \omega_e^2/(\omega\nu). \tag{11}$$

The Faraday and Ampere equations then yield the following wave equation for the magnetic field vector $\boldsymbol{H}$:

$$\nabla^2 \boldsymbol{H} + i\Omega\mu_0\epsilon\left(1 + i\frac{\sigma}{\Omega\epsilon}\right)\frac{\partial \boldsymbol{H}}{\partial t} - \mu_0\epsilon\frac{\partial^2 \boldsymbol{H}}{\partial t^2} = 0. \tag{12}$$

Note that the second term contains a dissipative factor involving the modulating frequency Ω. If the magnetic vector $\boldsymbol{H}$ is also assumed to vary harmonically:

$$\boldsymbol{H}(r, t) = \boldsymbol{H}(r)e^{i\omega t},$$

then the wave equation (12) takes the form

$$\nabla^2 \boldsymbol{H}(r) + k^2\left\{1 - \frac{\Omega}{\omega}\left(1 + i\frac{\sigma}{\Omega\epsilon}\right)\right\}\boldsymbol{H}(r) = 0, \tag{13}$$

where

$$k^2 = \omega^2\mu_0\tilde{\epsilon}e^{-i\Omega t}. \tag{14}$$

The corresponding equations for the electric vector are

$$\nabla^2 \boldsymbol{E} + \Omega^2\mu_0\epsilon \boldsymbol{E} + 2i\Omega\mu_0\epsilon\left(1 + i\frac{1}{2}\frac{\sigma}{\Omega\epsilon}\right)\frac{\partial \boldsymbol{E}}{\partial t} - \mu_0\epsilon\frac{\partial^2 \boldsymbol{E}}{\partial t^2} = 0, \tag{15}$$

and, assuming harmonic time dependence,

$$\nabla^2 \boldsymbol{E}(r) + k^2\left\{1 - \frac{2\omega\Omega}{\omega^2 + \Omega^2} \cdot \left(1 + i\frac{1}{2}\frac{\sigma}{\Omega\epsilon}\right)\right\}\boldsymbol{E}(r) = 0, \tag{16}$$

where

$$k^2 = (\omega^2 + \Omega^2)\mu_0\tilde{\epsilon}e^{-i\Omega t}.$$

The modulating and dissipative ffeects of the pressure oscillation at the frequency Ω are again in evidence.

W. C. Hoffman
Hughes Res. Labs.
Culver City, Calif.

Negative Feedback a Third of a Century Ago*

Apparatus

A rather curious radio receiver called an Infradyne[1] was designed by Sargent and described in 1926. The intermediate frequency was about 3600 kc which is the sum of the incoming signal and local oscillator frequencies. Thus as the signal varies from 550 to 1500 kc the oscillator varied from 3050 to 2100 kc. The image frequency is twice the IF less the signal frequency. Consequently it varies from 6650 to 5700 kc. A single tuned circuit in the mixer grid gives ample image and IF rejection.

Intermediate Amplifier

The heart of this receiver is the three-stage IF strip using UX199 tubes. A circuit diagram is shown in Fig. 1. No neutralization is incorporated. The amplifier is made stable by over-all feedback controlled by condenser C. When C has small capacity a maximum of feedback is secured. Recently I have been fortunate in securing[2] a brand new sample of this IF strip which never had been used.

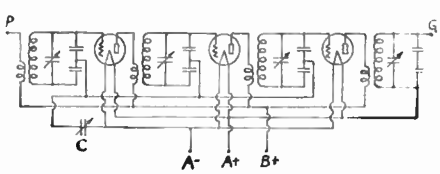


Fig. 1—Circuit diagram of 3600 kc amplifier. Condenser C controls the feedback.

Tests

A setup was made using a 12,000-ohm resistor and a 0.005 mfd condenser in series between signal generator and terminal P. This simulates the plate resistance of a UX199 tube. The output was measured by a vacuum tube voltmeter across terminals G and $A-$. No additional load resistance was used. Several tests were made using different adjustments of C. The response curve of Fig. 2 was secured with C at its smallest value. Fig. 3 shows the response for a bit larger value of C. Fig. 4 shows the response for C set to a value just less than that which will produce oscillation. The gain values are for 0 db of response curves. Bandwidths at 3 and 20 db are shown.

* Received by the IRE, February 21, 1959.
[1] E. M. Sargent, "The infradyne," *Radio*, vol. 8, pp. 11–14, 46; August, 1926.
[2] J. J. Simpson, 85-39 152nd St., Jamaica 32, N. Y., private communication.

Discussion

It is clear that this amplifier incorporates adjustable over-all degeneration to make it stable. As may be expected, the degeneration may be reduced and the gain increased at the expense of bandwidth. It is not clear from the inventor's discussion that he appreciates how the circuit works although he gives instructions on how to produce a response curve similar to Fig. 4. In any case, I nominate Sargent as the inventor of negative feedback. Does anyone else have an earlier

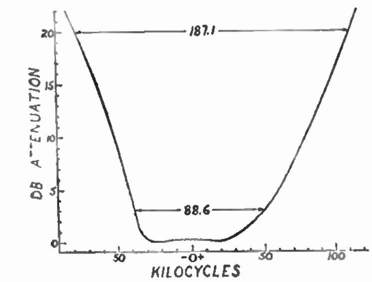


Fig. 2—Response curve with maximum feedback. Over-all gain, 11.1 db.

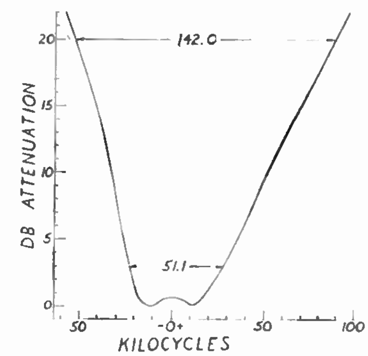


Fig. 3—Response curve with moderate feedback. Over-all gain, 18.1 db.

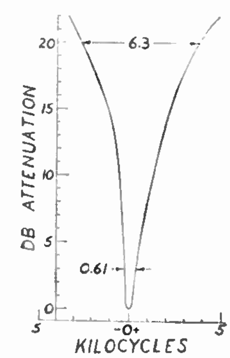


Fig. 4—Response curve with minimum feedback. Over-all gain, 47.2 db.

example? Perchance a reader may have one of these amplifiers. If so, I'd appreciate entering into a correspondence with him.

Grote Reber
National Radio
Astronomy Observatory
Green Bank, W. Va.

WWV Standard Frequency Transmissions*

Since October 9, 1957, the National Bureau of Standards radio stations WWV and WWVH have been maintained as constant as possible with respect to atomic frequency standards maintained and operated by the Boulder Laboratories, National Bureau of Standards. On October 9, 1957, the U.S.A. Frequency Standard was 1.4 parts in 10^9 with respect to the frequency derived from the UT 2 second (provisional value) as determined by the U. S. Naval Observatory. The atomic frequency standards remain constant and are known to be constant to 1 part in 10^9 or better. The broadcast frequency can be further corrected with respect to the U.S.A. Frequency Standard, as indicated in the table: values are given as part 10^{10}. This correction is *not* with respect to the current value of frequency based on UT 2. A minus sign indicates that the broadcast frequency was low.

The WWV and WWVH time signals are synchronized; however, they may gradually depart from UT 2 (mean solar time corrected for polar variation and annual fluctuation in the rotation of the earth). Corrections are determined and published by the U. S. Naval Observatory.

WWV and WWVH time signals are maintained in close agreement with UT 2 by making step adjustments in time of precisely plus or minus 20 msec on Wednesdays at 1900 UT when necessary; no time adjustment was made during this month at WWV and WWVH.

* Received by the IRE, May 22, 1959.

WWV FREQUENCY*

1959 April	Versus NBS** Atomic Standards 30-day moving average Seconds pulses at 15 mc	Versus Atomichron at WWV measuring time 1 hour 2.5 mc	Versus Atomichron at NRL measuring time 56 minutes 2.5 mc
1	−31	−38	−35
2	−31	−38	−34
3	−31	−38	−34
4	−32	−38	
5	−32	−37	
6	−33	−37	−34
7	−34	−37	−33
8	−35	−37	−34
9	−36	−37	−34
10	−36	−37	−33
11	−37	−37	
12	−37	−37	
13	−36	−36	−33
14	−36	−36	−33
15	−36	−37	−33
16	−36	−36	−33
17	−35	−36	−33
18	−35	−36	
19	−34	−36	
20	−33	−36	−32
21	−33	−36	−32
22	−32	−36	−32
23	−32	−36	−33
24	−31	−35	−32
25	−31	−36	
26	−30	−35	
27	−30	−36	−31
28	−30	−35	−31
29	−31	−36	−32
30	−31	−36	−32

* WWVH frequency is synchronized with that of WWV.
** Method of averaging is such that an adjustment of the frequency of the control oscillator appears on the day it is made. No adjustments were made during April.

NATIONAL BUREAU OF STANDARDS
Boulder, Colo.

ERRATA

In the June, 1959 issue of these PROCEEDINGS, p. 1157, all values for March should have been minus.

Contributors

Philip J. Allen (SM'57) was born in Whitinsville, Mass., on December 30, 1919. Following two years of employment with the General Radio Company, Cambridge, Mass., he entered Pennsylvania State University, University Park, and in 1944 received the B.S. degree in physics.

P. J. ALLEN

He has been with the U. S. Naval Research Laboratory, Washington, D. C., since joining that organization in July, 1944. In his various capacities, he has been engaged in developing automatic tracking radar systems, antenna feeds, lobeswitching and monopulse techniques, and related microwave techniques and components. Since 1951, he has headed the New Techniques Section responsible for devising new microwave techniques and components, and also functions as Assistant Branch Head of the Tracking Branch, Radar Division. He holds several patents in the field of microwaves.

Mr. Allen is a member of RESA.

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Malcolm R. Boyd (S'50–A'52–M'57) was born on January 23, 1924, in Boulder, Colo. He received the B.S. degree in electrical engineering from the University of Colorado, Boulder, in 1947. During the period from 1943 to 1946, he served in the U. S. Army Signal Corps.

M. R. BOYD

From 1947 to 1949 Dr. Boyd was employed at the RCA Laboratories in Princeton, N. J., where he was engaged primarily in gaseous electronic studies. In 1949 he entered Stanford University, Stanford, Calif., as a graduate student and research assistant in the Electrical Engineering Department, and received the M.S. and Ph.D. degrees in 1951 and 1953, respectively, both from Stanford. Since 1953 he has been a member of the technical staff at the General Electric Research Laboratory in Schenectady, N. Y.

Dr. Boyd is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

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Kai Chu was born in Nangtong, Kiangsu, China, on June 10, 1932. He received the B.S. degree in general engineering from the University of Portland, Portland, Ore., in 1954, and the M.S. degree in electrical engineering from the University of California, Berkeley, in 1958.

In 1954, he was associated with the RCA BIZMAC Computing System Engineering of the Radio Corporation of America. While in the information-handling and data-processing group, he was engaged in the design projects of high-speed magnetic tape files and multichannel magnetic recording heads, and later was in charge of magnetic head production. He was appointed to a research assistantship in the Department of Electrical Engineering of the University of California, Berkeley, in 1957. He spent a summer in research on high-density magnetic recordings with the International Business Machines Research Laboratory. Since entering the University, he has been conducting computer research in ferromagnetic thin films.


K. Chu

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Armni Frei was born on May 26, 1931 in Zürich, Switzerland. He attended high school at Zürich and the Swiss Federal Institute of Technology from 1952 to 1956. He received the M.S. degree in electrical engineering in 1956. Since 1956, he has been an assistant and research fellow in the Department of Advanced Electrical Engineering, Swiss Federal Institute of Technology, Zürich.


A. Frei

Mr. Frei has published papers on universal impedance curves and on resistance-capacitance networks as analog computers.

❖

Harry J. Gray, Jr. (S'45–A'46–M'55) was born in St. Louis, Mo., on June 24, 1924. He received the B.S. and M.S. degrees in electrical engineering in 1944 and 1947, and the Ph.D. degree in 1953, all from the University of Pennsylvania, Philadelphia.


H. J. Gray, Jr.

After serving in the U. S. Navy as a radio specialist officer, he returned to the University of Pennsylvania in 1946. He worked on both the EDVAC and MSAC computers and then was associated with the development of a digital operational flight trainer (UDOFT). In 1954, he joined the Remington Rand Univac Division of Sperry-Rand Corporation, Philadelphia, Pa., where he worked on magnetic and transistor circuits. He developed the high-speed circuit system of LARC, and when he left in 1957, he held the position of staff engineer. At present, he is associate professor of electrical engineering in the Moore School of the University of Pennsylvania. He has been responsible for the design of a multiple cockpit digital operational flight trainer for the U. S. Navy training Devices Center and a task on a multiple task study for USASRDL, Fort Monmouth, N. J.

Dr. Gray is a member of the Association for Computing Machinery, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

❖

Winston E. Kock (S'45–F'52) was born in Cincinnati, Ohio on December 5, 1909. He received the E.E. degree and the M.S. degree in physics from the University of Cincinnati in 1932 and 1933, respectively, and the Ph.D. degree in physics from the University of Berlin, Germany, in 1934. He was a teaching fellow at the University of Cincinnati from 1934 to 1935, attended the Institute for Advanced Study, Princeton, N. J., from 1935 to 1936, and was a Fellow of the Indian Institute of Science at Bangalore in 1936.


W. E. Kock

In 1936, he became Research Engineer and later Director of Electronic Research at the Baldwin Piano Company.

In 1942, he joined the Radio Research Department of the Bell Telephone Laboratories, where he conducted research on microwave antennas. He was appointed Director of Acoustics Research in 1951, in which capacity he directed the research on the Navy Jezebel project, and in 1956, he became Director of Audio and Video Systems Research.

He joined the Systems Division of Bendix Aviation Corporation as Chief Scientist in late 1956, and became Director and General Manager of the Research Laboratories Division in January, 1958.

In 1938, Dr. Kock received the award of Outstanding Young Electrical Engineer from Eta Kappa Nu, and in 1952 he was awarded the honorary degree of Doctor of Science by the University of Cincinnati. He is a Fellow of the American Physical Society and the Acoustical Society of America, and a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

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Donald A. Linden (S'49–A'51) was born in Vienna, Austria, in 1922. He received the B.E.E. degree from New York City College in 1949, and the S.M. degree in electrical engineering from the Massachusetts Institute of Technology, Cambridge, in 1950. Until 1957, he was with the research division of the Philco Corporation, Philadelphia, Pa., where he was engaged primarily in radar systems analysis and video data processing. He joined the Philco Western Development Laboratories, Palo Alto, Calif., in 1957. At present, he is on a leave of absence at the Stanford Electronics Laboratories, Stanford University, Stanford Calif., as a National Science Foundation fellow.


D. A. Linden

Mr. Linden is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

❖

Alan L. McWhorter (S'51–M'56) was born in Crowley, La., on August 25, 1930. He received the B.S. degree from the University of Illinois, Urbana, in 1951, and the Sc.D. degree from the Massachusetts Institute of Technology, Cambridge, in 1955, both in electrical engineering.


A. L. McWhorter

From 1953 to 1955, he was a research assistant in the M.I.T. Research Laboratory of Electronics and Lincoln Laboratory, Lexington, Mass., studying noise and related surface effects in germanium. In 1955, he joined the staff of the Lincoln Laboratory, at first continuing his semiconductor surface studies. Since then he has been engaged in research on solid-state masers and low temperature electrical properties of germanium.

Dr. McWhorter is a member of the American Physical Society, Sigma Xi, Tau Beta Pi and Eta Kappa Nu.

❖

Robert H. Rediker (A'53–SM'58) was born in Brooklyn, N. Y. on June 7, 1924. He received the B.S. degree in electrical engineering in 1947 and the Ph.D. degree in physics in 1950 from the Massachusetts Institute of Technology, Cambridge, Mass.


R. H. Rediker

During 1950–1951, Dr. Rediker was a research associate in cosmic rays in the Physics Department of M.I.T. In 1951 he became a staff member of M.I.T.'s Lincoln Laboratory in Lexington, Mass., where he worked on transistorized computer circuits.

During the academic year 1952–1953, he was a research associate at the Physics Department of Indiana University, Bloomington, Ind. Since June, 1953 he has been engaged in semiconductor device research at Lincoln Laboratory where he now heads the Applied Solid-State Physics group.

Dr. Rediker is a member of the American Physical Society and Sigma Xi.

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Glen M. Roe was born in Stanley, Wis., on December 17, 1916. He received the Bachelor's degree from St. Olaf College, Northfield, Minn. and the Master's and Ph.D. degrees from the University of Minnesota.

G. M. Roe

He served as a senior physicist in in the U. S. Navy's Bureau of Ships, and as a research associate in the Knolls Atomic Power Laboratory, Schenectady, N. Y., before joining the General Electric Research Laboratory, Schenectady, N. Y., in 1954. A member of the Electron Physics Research Department, he specializes in theoretical physics, including nuclear physics, electromagnetics, and acoustics.

Dr. Roe is a member of the American Physical Society, Society for Industrial and Applied Mathematics, American Mathematical Society, AAAS, and Sigma Xi.

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J. R. Singer (S'55–M'57) was born on October 16, 1921, in Cleveland, Ohio. After four years at sea as a navigator, he received the B.S. degree in mathematics from the University of Illinois, Urbana, and the M.S. and Ph.D. degrees in physics from Northwestern University, Evanston, Ill., and the University of Connecticut, Storrs, in 1953 and 1955, respectively.

J. R. Singer

He has been a member of the engineering staff of Sperry Gyroscope Co. and Boeing Airplane Co. He was a solid-state physicist at the Naval Ordnance Laboratory, White Oak, Md., and chief staff physicist of the National Scientific Laboratories, Inc., Washington, D. C. He is presently an associate professor in the electrical engineering department at the University of California, Berkeley. His interests are in solid-state physics, particularly magnetic materials, masers, and electronic systems.

Dr. Singer is a member of the Physical Society, Sigma Xi, AAAS, Physical Society of Great Britain, and the Optical Society of America.

Henry S. Sommers, Jr., was born on April 21, 1914 in St. Paul, Minn. His undergraduate work was at Stanford University, Stanford Calif., and the University of Minnesota, Minneapolis, Minn., and he received the B.A. in physics from the University of Minnesota in 1936. In 1941 he was granted a Ph.D. in physics from Harvard University, Cambridge, Mass., majoring in nuclear physics. During 1941–1942 he was an instructor in physics at Harvard University and Radcliffe College, Cambridge, Mass. From 1942–1945 he was a staff member at the Radiation Laboratory, M.I.T., Cambridge Mass., where he was a radar systems engineer on Army and Navy antiaircraft fire control radars.

H. S. Sommers, Jr.

After the war he was appointed assistant professor of physics at Rutgers University, New Brunswick, N. J., where he remained until 1949. Here his research was devoted to developing a precision current control for a large electromagnet. In 1949 he became a staff member at the Los Alamos Scientific Laboratories, where he did fundamental studies on the thermodynamic properties of liquid helium and on the scattering of neutrons by liquid helium.

He left there in 1954 for his present position as a senior member of the Technical Staff of the RCA Research Laboratories in Princeton, N. J., where he has been studying the electrical properties of insulators and semiconductors. He is the author of a variety of papers in the fields of solid-state physics, cryogenics, nuclear physics, chemical kinetics, and instrumentation.

Dr. Sommers is a Fellow of the American Physical Society and a member of the Federation of American Scientists.

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Max J. O. Strutt (SM'46–F'56) was born on October 2, 1903 in Surakarta, Java. He studied at the University of Munich, Institute of Technology at Munich, and Institute of Technology at Delft, The Netherlands. From the latter he received the M.S. degree in electrical engineering and the Doctor of Tech.Sc. degree (cum laude) in 1926 and 1927, respectively. He was a research engineer at The Philips Company, Ltd., Eindhoven, The Netherlands, from 1927 to 1948. Since 1948, he has been professor and director of the Department of Advanced Electrical Engineering, Swiss Federal Institute of Technology, Zürich, Switzerland.

M. J. O. Strutt

He holds more than 60 U. S. patents on electron tubes and circuits, especially at VHF and UHF. Among his awards are the Doctor Honoris Causa (1950) conferred by the Institute of Technology, Karlsruhe, Germany and the Carl Friedrich Gauss Medal (1954) of the Society of Sciences, Brunswick, Germany. In 1955 he was elected an honorary member of the Society of Sciences, Brunswick.

Dr. Strutt is a member of the Swiss Society of Electrical Engineers, the German Society of Electrical Engineers, the Swiss Society of Sciences at Berne, Switzerland, the German Physical Society, the Swiss Mathematical Society, and the Zürich Physical Society.

❖

R. D. Tompkins was born on July 27, 1926, in Paterson, N. J. After serving as an electronic technician in the Navy during World War II, he attended Case Institute of Technology, Cleveland, Ohio, where he received the B.S. degree in electrical engineering in 1950.

R. D. Tompkins

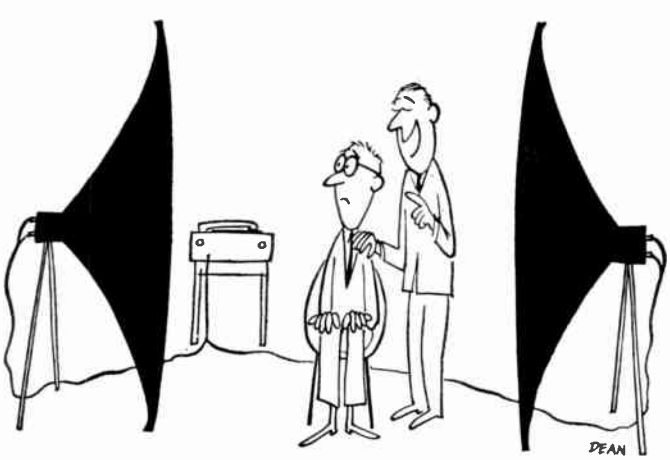
From 1950 to 1952, he was employed by the Radar Division of the Naval Research Laboratory, Washington, D. C. In 1952, he left to accept a position as the electrical engineer for the Bethlehem Chile Iron Mines Company in El Tofo, Chile. Returning to the United States in 1954, he rejoined the staff of the Naval Research Laboratory, where he is currently working on microwave systems and techniques for the Tracking Branch of the Radar Division.

Mr. Tompkins is a member of Eta Kappa Nu.

Scanning the Transactions

Christian Doppler, an Austrian mathematician and physicist, first enunciated the physical law that bears his name in 1842. Yet, during the following 100 years about the only ones who found Doppler's principle of practical importance were astronomers, who used the Doppler shifts of spectral lines of star light to measure the velocities of celestial bodies. Now, rather suddenly, Doppler is becoming a household word to radio engineers, thanks to the high velocities encountered with jet and rocket propelled vehicles and the increased frequency sensitivity of modern radio equipment. The Doppler phenomenon is proving to be a most useful tool in air and space navigation, in satellite tracking, and in studying the electron densities of the upper ionosphere. On the other hand, it also causes undesired frequency deviations in communication links between high-speed space vehicles and ground stations. Moreover, in order to escape from the earth's gravitational field a space vehicle must reach speeds that are a perceptible fraction (about $\frac{1}{4}$ of one per cent) of the velocity of light, and at these and higher speeds relativistic effects are encountered which cause the Doppler effect to become even more pronounced. This leads one to wonder what headaches lie in store for the interplanetary FCC of the future when space craft transmissions begin to wander off their assigned frequencies in accordance with Mr. Doppler's law. If this should ever come to be a problem, however, we have no doubt that radio engineers will provide an ingenious solution, perhaps by designing anti-Doppler circuits which will automatically shift the frequency of the space vehicle's transmitter in accordance with its speed in order to compensate for the Doppler shift. (F. J. Tischer, "Doppler phenomena in space communications," IRE TRANS. ON COMMUNICATIONS SYSTEMS, May, 1959; C. L. Temes, "Relativistic consideration of Doppler shift," IRE TRANS. ON AERONAUTICAL AND NAVIGATIONAL ELECTRONICS, March, 1959.)

The simplification of stereo amplifiers has been very much in the limelight of late. Equipment for handling two-channel stereo with one amplifier instead of the usual two was described last March at the IRE National Convention. Now comes a method for producing three-channel stereo with only two amplifiers. The system uses three microphones, two sound tracks, two amplifiers, and three loudspeakers. The output of the third (middle) microphone is mixed into the two tracks and, with the help of a phase shifting network, is recovered during playback and fed into the third (middle) loudspeaker to produce directional effects that are practically indistinguishable from the real thing. (P. W. Klipsch. "Three channel stereo playback of two tracks derived from three microphones," IRE TRANS. ON AUDIO, March–April, 1959.)



"Now sit right here to get the full effect. Uh . . . you had better give me your glasses."

Reprinted from IRE STUDENT QUARTERLY

Communication satellites, which only a year ago seemed to be a far-off dream, have now become a matter of serious planning for the near future. The National Aeronautics and Space Agency plans to place experimental 100-foot metallized balloons into orbit this year. A recent PROCEEDINGS paper[1] has shown that it would be technically feasible to use these balloons in a polar orbit as passive relays for transoceanic communication. While quite a number of such satellites would have to be placed in orbit to ensure that one would always be within sight of both continents, the modest altitude (3000 miles) holds the power requirements down within the capabilities of the present microwave art. Meanwhile, the further suggestion has been made that ultimately we might use a 22,000 mile equatorial orbit.[2] At this altitude a satellite would circle the earth at just the same rate as the earth rotates, so that the satellite would remain essentially stationary in the sky. This scheme has the advantage of providing coverage to nearly half the earth with only one satellite. However, the problems of launching, positional control, and transmitter power are far more formidable. Such a satellite might be used either as an active repeater or as a passive reflector. The latter case has now been analyzed in some detail. It turns out that the requirements for transmitter power, antenna size, and satellite diameter are indeed formidable, but not beyond the realm of possibility. For example, it is estimated that to provide a television channel at a frequency of 3000 mc would require a 1000-foot transmitting antenna, a 1000-foot satellite, a 140-foot receiving antenna, and a 50 kilowatt transmitter. It might be noted in passing that if an ordinary receiver were used instead of a maser amplifier, the power requirements would be about 10 times greater. The equivalent noise temperature of maser amplifiers, including external noise picked up by the antenna, has been variously estimated at 100°K to as low as 20°K. It is interesting to see confirmation from the fact that the first results with the maser-equipped radio telescope at the Naval Research Laboratory, reported just last month,[3] gave a figure of 85°K, and showed that 20°K could eventually be achieved. (Morris Handelsman, "Performance equations for a 'stationary' passive satellite relay (22,000-mile altitude) for communication," IRE TRANS. ON COMMUNICATIONS SYSTEMS, May, 1959.)

A novel hearing aid using a magnetic coupling has been suggested by recent work at the New York State Psychiatric Institute. A method has been developed whereby a tiny permanent magnet can be glued to the eardrum with water-soluble glue. A small coil of wire is placed close to the ear whose drum has been fitted with the magnet. When an audio frequency signal is fed to the coil, an alternating magnetic field is produced that will act on the permanent magnet and set it into vibration. Thus the eardrum is made to vibrate and stimulation of the ear occurs. In pathological cases where the middle ear is damaged or destroyed, the magnet could

[1] J. R. Pierce and R. Kompfner, ""Transoceanic communication by means of satellites," PROC. IRE, vol. 47, pp. 372–380; March, 1959.
[2] L. V. Berkner, "IRE Enters space," PROC. IRE, vol. 47, pp. 1048–1052; June, 1959.
[3] J. A. Giordmaine, L. E. Alsop, C. H. Mayer and C. H. Townes, "A maser amplifier for radio astronomy at X-band," PROC. IRE, vol. 47, pp. 1062–1069; June, 1959.

presumably be attached to some suitable structure of the inner ear and auditory stimulation might again be possible. A system using magnetic coupling may have an advantage over conventional hearing aids: the danger of audio feedback from the transducer to the hearing aid microphone is avoided and higher amplifications could be used. Also molds that tightly fit the outer ear canal would no longer be necessary. (J. Rutschmann, "Magnetic Audition—Auditory stimulation by means of alternating magnetic fields acting on a permanent magnet fastened to the eardrum," IRE TRANS. ON MEDICAL ELECTRONICS, March, 1959.)

Charles Lindbergh, it may surprise many to learn, achieved fame not only in the air but also in the medical research laboratory. Lindbergh had become interested in the idea of an artificial heart which could be substituted for the human heart during operations. Upon hearing of the work of the celebrated surgeon Alexis Carrel at the Rockefeller Institute, Lindbergh went there and offered his services as an amateur inventor and mechanic. Starting in 1930 he worked for five years as a volunteer in Carrel's laboratory, who was then doing pioneering work in trying to grow human organs in flasks in the hope of studying their function and of providing replacements for diseased organs of patients. Carrel's work required the development of a sterilizable glass pump, into which bacteria could not penetrate, for supplying nutritive fluids to the organ, and it was to this problem that Lindbergh devoted his ingenuity. This work led, in the middle of 1935, to the first successful experiment with an explanted organ—a thyroid gland. Later, in their best experiment, Carrel and Lindbergh kept a thyroid gland going for thirty days. About twenty Lindbergh pumps were made, several of which survive in various institutions, including the National Museum at Washington. Two or three investigators are in fact now using improved apparatus based on that of Carrel and Lindbergh. In time, interest in the culture of explanted organs may revive, and Lindbergh's early contributions may take on added significance. (G. W. Corner, "Organ culture at the Rockefeller Institute," IRE TRANS. ON MEDICAL ELECTRONICS, March, 1959.)

Antennas with nearly limitless bandwidths have at last become a practical reality. A special spiral design has resulted in a circularly polarized antenna having an essentially constant radiation pattern and input impedance over bandwidths greater than 20 to 1, a frequency range considered impossible only a few years ago. By way of comparison, conventional "broadband" antennas have a bandwidth of no more than 2 or 3 to 1. The antenna shape takes the form of a plane curve which spirals out from the center in an exponential fashion. The design is based on the principle that if the shape of an antenna can be specified entirely by angles, its performance will be independent of wavelength. To be totally frequency independent the antenna spiral would have to go out to infinity. In a practical antenna the diameter of the spiral need be only a half wavelength, measured at the lowest frequency of operation, and still be nearly independent of frequency. (J. D. Dyson, "The equiangular spiral antenna," IRE TRANS. ON ANTENNAS AND PROPAGATION, April, 1959.)

Nanosecond, for the benefit of those unfamiliar with the word, is a term meaning 10^{-9} second, or millimicrosecond. It is a term that is becoming increasingly common in the semiconductor and nuclear fields in connection with the generation and handling of extremely short pulses. In particular, nuclear research with high energy accelerators frequently requires time resolution of events which occur at nanosecond rates. An important element in nanosecond pulse circuits is the pulse transformer, which is used to match impedance of various devices having different impedance levels so that they may work in concert without degrading the transient response. Nanosecond pulse transformers can now be made for impedance matching in the range of 50 to 300 ohms having rise times of less than 5×10^{-10} seconds. (C. N. Winningstad, "Nanosecond pulse transformers," IRE TRANS. ON NUCLEAR SCIENCE, March, 1959.)

Books

The Radio Handbook, 15th Edition, edited by William I. Orr

Published (1959) by Editors and Engineers, Ltd., Summerland, Calif. 785 pages+13 index pages. Illus. $6\frac{1}{4}\times9\frac{1}{2}$. $7.50.

An appropriate title for this handy perennial would be "From Soup to Nuts." Astoundingly enough it starts with a description of the Arabic number system and how to add two numbers, and proceeds to a crescendo of the design of complex transistorized circuits and equipment. On the way it digresses to instruct in the use of a plane in trimming aluminum chassis.

Subjects covered include arithmetic, algebra, trigonometry, complex numbers, direct and alternating currents, vacuum tubes, transistors, all kinds of vacuum tube circuits, high-fi techniques, computers (both analog and digital), modulation theory and circuitry, single sideband transmission, propagation in general, antenna theory and construction, rotary beams, power supply design and construction, and electronic test equipment. Name it and you can find it.

Naturally, even in 800 pages, all of these subjects cannot be treated in depth. The treatments are in fact fairly uneven—digital computers including "and" "or" circuitry takes four and one-half pages, while vacuum tube amplifiers takes twenty-eight. In general, the depth of treatment is slanted toward the amateur experimenter, who wants to build equipment in his own home and to have at least a limited understanding of how it works. This job is very well done.

The fact that the book is in its fifteenth edition says more about its acceptance than any reviewer could.

KNOX MCILWAIN
Burroughs Corp.
Paoli, Pa.

The Technical Writer, by J. W. Godfrey and G. Parr

Published (1959) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 299 pages+6 index pages+31 appendix pages. Illus. $5\frac{1}{4}\times8\frac{1}{4}$. $8.50.

This book discusses the problems associated with writing, editing, and producing technical literature. It is oriented by the authors' particular experience toward the electrical engineering field. It is dedicated to Reginald O. Kapp, author of "Presentation of Technical Information," a book high in the rating of this reviewer. Although it owes a debt to the earlier book, it is by no means a revisitation; it stands on its own, offering its material according to its authors' sense of proportion and in its own time.

Its ten chapters are nicely balanced; they include discussions of the general subjects of writing, illustrating, printing, and editing. The tenth chapter discusses the problems of setting up a technical publications unit and clearly reflects what must be first-hand experience by the authors in this relatively new professional activity. Especially valuable are the chapters describing typewriting and printing equipment. The chapters on writing, style, and presentation which really form the heart of this type of book are frank appraisals of the writing problem that neither abrogate the rules of grammar nor dogmatize. The book is written in a less formal style that makes for grateful reading; yet, being written by two authors in Great Britain, it does not lack elegance. The American reader will, of course, find only academic interest in the numerous recommendations which are valid in the Isles but which are not acceptable here. However, these points seem trivial and, indeed, the authors seem at every turn to be thoroughly familiar with both our American styles and our equipment.

The book should be valuable reading for any electrical engineer who ever writes a report or articles, and any director of research or engineering who is concerned with report, proposal, or technical manual writing. It ought to be in the library of anyone engaged full time as a technical writer, for his own benefit and as outside authority against any lack of understanding on the part of a technical staff member.

J. D. Chapline
Philco Corporation
Philadelphia, Pa.

The Radio Amateur's Handbook, 36th Edition, 1959, edited by the Headquarters Staff of the American Radio Relay League

Published (1959) by the American Radio Relay League, West Hartford, Conn. 584 pages+15 index pages+32 pages of tables+110 pages of catalogue section. Illus. 6¾×9¾. Paper cover. $3.50.

It is always pleasant to review a hardy old perennial like the ARRL Handbook, now in its 36th edition. A book that has sold over three million copies is a record in any publishing world, and there must be a reason for such a phenomenal record.

The reason, of course, is well known. The Handbook is the standard manual of radio amateurs, a construction manual, a reference work, a training text for classroom or amateur shack, a catalog of current products—in short, a necessity.

New concepts, new types of construction, and new devices are introduced in each new edition so they may be easily assimilated into amateur (or professional) techniques. In this edition such matters as transistors, single sideband, and teletype take their places beside devices and methods which have existed since the beginning of ham radio.

No matter what an amateur does for his living, he should own the latest edition of this Handbook if only to read and read again the paragraphs on operating courtesy and how to be a good operator.

Keith Henney
McGraw-Hill Book Co.
New York, N. Y.

Recent Books

Environmental Requirements Guide for Electronic Component Parts, published by the Advisory Group on Electronic Parts and the Advisory Group on Electron Tubes, Office of the Director of Defense Research and Engineering. Available, free, from the Office of Technical Services, Dept. of Commerce, Washington 25, D. C. Summary of environmental requirements and test procedures specified by the three Military Departments for electronic components.

Grant, H. N., *Mobile Radio Telephones*. The Macmillan Co., 60 Fifth Ave., N. Y. 11, N. Y. $4.50.

James, Glenn and Robert C. James, Eds., *Mathematics Dictionary*. D. Van Nostrand Co., Inc., Princeton, N. J. $15.00. Definitions of over 7,000 mathematical terms, concepts and relationships and including multilingual indexes of Russian, German, French and Spanish equivalents.

Jelley, J. V., *Cerenkov Radiation and Its Applications*. Pergamon Press, 122 E. 55 St., N. Y. 22, N. Y. $10.00.

Klewe, H. R. J., *Interference between Power Systems and Telecommunications Lines*. St. Martin's Press, Inc., 103 Park Ave., N. Y. 17, N. Y. $12.50.

Semiconductor Abstracts, compiled by the Batelle Mem. Inst. John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $12.00.

Tepper, Marvin, *Fundamentals of Radio Telemetry*. J. F. Rider, Inc., 116 W. 14 St., N. Y. 11, N. Y. $2.95.

Abstracts of IRE Transactions

The following issues of Transactions have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
Aeronautical and Navigational Electronics	ANE-6, No. 1	$0.95	$1.45	$2.85
Antennas and Propagation	AP-7, No. 2	1.75	2.60	5.82
Audio	AU-7, No. 2	0.40	0.60	1.20
Broadcast and Television Receivers	BTR-5, No. 1	1.40	2.10	4.20
Communications Systems	CS-7, No. 1	1.85	2.75	5.55
Medical Electronics	ME-6, No. 1	1.80	2.70	5.40
Microwave Theory and Techniques	MTT-7, No. 2	2.20	3.30	6.60
Nuclear Science	NS-6, No. 1	1.50	2.25	4.50

*** Libraries and colleges may purchase copies at IRE Member rates.**

Aeronautical and Navigational Electronics

Vol. ANE-6, No. 1, March, 1959

Frontispiece—Harry R. Mimno (p. 2)

Editorial—E. A. Post (p. 3)

Radio-Compass Testing with Small Shielded Enclosures—A. S. Markham (p. 4)

A method is described by which radio compasses may be tested using small shielded enclosures (24 by 24 by 24 inches for many loop types) instead of screen rooms. This procedure results in economies of cost and space, as well as providing portability and rapid setup. Application of accepted methods of determining field strength for calibration purposes is discussed. Enclosure modifications for utilization of flush loops and the incorporation of rotation facilities for loop bearing-accuracy tests are also covered. Some approaches to enclosure design and construction are described, and examples of different types of existing and proposed units are shown. The possibility of using a small enclosure within an existing screen room as an aid in curing serious noise and interference problems, as an inexpensive substitute for renovating or replacing the screen room, or as a means for locating sources of noise or other disturbances, such as stray fields and RF leakage from signal generators, is also described.

Background and Principles of Tacan Data Link—B. Alexander, R. C. Renick, and J. F. Sullivan (p. 9)

The integration of Tacan with very-high-frequency omnidirectional range (VOR) to provide one common air-navigation and air-traffic-control system requires, for full usefulness, a method of automatic air-surface communication. Such a method, called the Tacan data link, has been devised, and is being flight-tested. Using this data link, messages can be received and sent to each of 120 aircraft every 2.67 seconds. Such messages would consist of navigational, aircraft-status, and traffic-control information. The data link employs the Tacan surface Beacon to carry both analog and digitally-coded messages, interpolating the coded pulse bursts, which last approximately 3 msec, 45 times every second. No additional transmitters or receivers are needed.

The Gyrovibrator—N. D. Diamantides (p. 16)

A new device for measuring angular rate is analyzed and its mechanical and electrical design features presented. Analytical proofs of three important characteristics of the instrument are given. These characteristics are: 1) a sinusoidally modulated-carrier voltage with the envelope representing the magnitude of the useful output signal; 2) zero drift as well as zero offset effects separated from the true output in the form of a dc quantity, and 3) a reference signal through which both the magnitude and the orientation of the input angular-velocity component in the plane of the spin are determined.

Radar Performance Degradation in Fog and Rain—H. E. Hawkins and O. La Plant (p. 26)

An equation is derived which can be used to compare the ability of a pulse radar to detect targets in rain and fog with its ability to detect similar targets in dry air. It is a common misconception that attenuation in the path between radar and target is the only consideration. However, clutter in the path between radar and target is the real concern. Clutter power backscattered from the storm in the immediate region of the target may be by far the most dominant degrading factor. The equation derived, and the resulting curves, take into consideration the combined effects of backscattering and attenuation and interpret the effects in terms of comparative dry-air and wet-air range capability. They enable 1) prediction of the relative performance of a given radar under various meteorological conditions, and 2) comparison of radars of different parameters operating under identical meteorological conditions.

The PAR-Scope: An Oscilloscope Display for Weather Radars—E. Kessler, III (p. 31)

The PAR-scope, an oscilloscope display for pulsed radars, uses gated, averaged, amplitude information as one coordinate, and azimuth or elevation as the other coordinate; its display is a profile of average reflectivity. It is operated with a scanning antenna, permitting simultaneous operation with the RHI or PPI display. Such combinations provide means for rapid, three-dimensional quantitative mapping of reflectivity in the vicinity of the radar. Theoretical analysis of PAR-scope performance considers the response of the scanning system to spatial variations of average reflectivity and to the rapid fluctuations characteristic of weather echoes. It is shown that reflectivity distributions of major interest can be easily represented to an accuracy within 1 db of that implicit in the radar calibration.

Correction to "Report on the Fifth Annual East Coast Conference on Aeronautical and Navigational Electronics" (p. 36)

Correspondence (p. 37)
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Book Reviews (p. 42)
PGANE News (p. 44)
Contributors (p. 46)
Suggestions to Authors (p. 48)

Antennas and Propagation

Vol. AP-7, No. 2, April, 1959

A 215-Mile 2720-MC Radio Link—L. H. Doherty and G. Neal (p. 117)

The results from the operation of a 215-mile, 2720-mc radio link are discussed. The link was operated for a period of twenty months. The yearly median signal level was 79 db below free space with a seasonal variation between 12 and 14 db in the hourly medians. If attention is confined to a single season the hourly medians have a log-normal distribution. No diurnal variation was observed. Probability distributions of signal amplitude based on 30-second samples were most commonly Rayleigh although some significant departures from this law did occur. A study of the time variation of the 30-second median shows the standard deviation of these medians five minutes apart to be about 1 db, and thirty minutes apart to be 2.6 db. The same type of analysis is also performed on the hourly medians. A diurnal variation is observed in the fading, with the midafternoon rate being almost twice that recorded during the early morning. Pulse distortion and meteorological correlations are discussed qualitatively.

Preliminary Results of 400-MC Radar Investigations of Auroral Echoes at College, Alaska—R. L. Leadabrand, L. Dolphin, and A. M. Peterson (p. 27)

Results of radar investigations of auroral ionization at a frequency of 398 mc are described. The radar detected auroral echoes at College, Alaska (near Fairbanks), even when the line of sight from the transmitter does not intersect the earth's magnetic field lines at perpendicular incidence. The auroral echoes were observed with off-perpendicular intersection angles as great as 10°. The requirement for near-perpendicular intersection of the radar beam with the earth's magnetic field is therefore met.

The auroral echoes at 398 mc were seen frequently. Occasionally they were very strong; those of highest amplitude were as much as 27 db above the receiver noise level. The echoes were detected over a relatively large echoing region corresponding to bearings within ±45° of geomagnetic north, and elevation angles of 0°–17°.

Two types of auroral echoes were observed—discrete and diffuse. The discrete echoes corresponded roughly to reflections from visual auroral forms seen at night. The diffuse echoes corresponded to reflections from a large echoing region that apparently existed most often during daylight hours. Estimates of the wavelength, λ, dependence of auroral echo power, although quite crude, are deduced as λ^{10} for the radar site at College, Alaska, and λ^5 for sites where perpendicular reflection can be obtained.

Correlation Function and Power Spectra of Radio Links Affected by Random Dielectric Noise—D. S. Bugnolo (p. 137)

The correlation function and corresponding power spectrum of an electromagnetic wave affected by random dielectric noise is related to the power spectrum of the source by an extention of the notions of time-variable linear networks. It will be shown that in general, the power spectrum of the received signal can be regarded as the output of a network characterized by a time-variable transfer function. The results are applied to a long line of sight radio link and used to predict the error in the received signal is a mean squared sense. This will be used to show that the rate of a source is bounded such that there exists a maximum rate R given a bandwidth δ and scattering parameters of the atmosphere.

Aperture-to-Medium Coupling on Line-of-Sight Paths: Fresnel Scattering—E. Levin, R. B. Muchmore, and A. D. Wheelon (p. 142)

An electromagnetic signal propagated through a medium of randomly varying dielectric exhibits random-phase fluctuations. The phase of the total signal is the average of all rays which strike the reflector and is therefore smoothed by a receiver-reflector combination of finite aperture. This paper presents a theoretical analysis and numerical results for this phase-smoothing for small phase perturbations. The receiver is assumed to be a circular parabolic reflector with a collecting feed at the focus. The propagation is described by Fresnel scattering and the one dimensional ray theoretical expressions employed. Closed form results are obtained for three separate space correlation models of the random dielectric medium. These results are valid so long as the scattering parameter $L\lambda l_0^{-2}$ is small.

The Inverse Scattering Problem in Geometrical Optics and the Design of Reflectors—J. B. Keller (p. 146)

The inverse scattering problem considered here is that of finding the shape of a reflector which produces a prescribed scattered wave. The scattered wave is characterized by its angular pattern, which determines the differential scattering cross section of the reflector. The problem is solved by means of explicit formulas for cylindrical and for rotationally symmetric objects. Plane, cylindrical, and spherical incident waves are considered. The general three dimensional object is also treated. The method of geometrical optics is used throughout.

On Scattering by Large Conducting Bodies—R. F. Harrington (p. 150)

Two sets of sources, equivalent in the sense that they produce the same field as does an illuminated conductor, are discussed. Both representations are suggestive of approximation. Crude approximations are made, yielding what are called "the physical optics solution," and the "image induction solution." It is shown that these two solutions are reciprocal to each other. This means that, given a source and an observer, the solution by one method is equal to the solution by the other method with the source and observer interchanged. Both solutions are amenable to further refinement if more accurate solutions are desired.

Asymmetrical Trough Waveguide Antennas—W. Rotman and A. A. Oliner (p. 153)

Line source radiators of the leaky wave type may be constructed readily with trough waveguide, which consists of a rectangular trough containing a symmetrically located center fin. Two types of such radiators are investigated: a continuously asymmetrical form suitable for near endfire radiation, and a periodically asymmetrical design permitting radiation through broadside. The propagation characteristics of the leaky wave on the continuously asymmetrical structure are determined theoretically by means of a transverse resonance procedure, and comparison is made with measured values. The influence of finite center fin thickness is also accounted for. Designs and radiation patterns are presented for a typical antenna of each type.

A Contribution to the Theory of Cylindrical Antennas—Radiation Between Parallel Plates—L. Lewin (p. 162)

A solution of the wave equation is obtained corresponding to an outgoing cylindrical wave between two parallel ground planes, and satisfying the boundary conditions appropriate to a center-fed, full height antenna at the origin. Expressions for the antenna current and impedance are obtained from a small radius approximation to these; formulas are obtained for comparison with those from usual antenna theory. Attention is directed to the expansion parameter and gap capacitance.

Radiation from Ring Sources in the Pres-

ence of a Semi-Infinite Cone—L. B. Felsen (p. 168)

A rigorous formulation for ring source Green's functions for a semi-infinite cone, carried out in terms of images in an infinitely extended angular transmission line, is evaluated asymptotically in the short-wave length range to yield geometric-optical, diffracted, and transition effects. Both scalar problems with Neumann or Dirichlet type boundary conditions, and electromagnetic problems appropriate to a perfectly conducting cone, are treated. The results are applied to calculate approximately the radiation pattern for a prescribed magnetic current distribution on the cone surface which can be taken to represent the excitation due to certain leaky wave type antenna arrays.

The Equiangular Spiral Antenna—J. D. Dyson (p. 181)

A circularly polarized antenna is described which makes possible bandwidths that a few years ago were considered to be impossible.

The design of the antenna is based upon the simple fundamental principle that if the shape of the antenna were such that it could be specified entirely by angles, its performance would be independent of wavelength. Since all such shapes extend to infinity it is necessary to specify at least one length for an antenna of finite size. The one length in this antenna, the arm length, need only be of the order of one wavelength at the lowest frequency of operation to obtain operation essentially independent of frequency, and the geometry of the antenna allows this arm length to be spiraled into a maximum diameter of one half wavelength or less. Antennas have been constructed that have an essentially constant radiation pattern and input impedance over bandwidths greater than 20 to 1.

The Influence of Gain and Current Attenuation on the Design of the Rhombic Antenna—R. P. Decker (p. 188)

With the increased use of the ionospheric scatter mode of propagation in the VHF range, the horizontal rhombic antenna is employed in many instances because of its simplicity, high performance, and low maintenance costs. Designers of these long rhombics have no doubt realized that the "maximum output" and "maximized" designs described by Harper and others do not take into account current attenuation due to radiation and cannot generally be employed when $l/\lambda > 8$ because of the requirement that the first ground factor maximum should agree closely with the free space vertical pattern maximum. This leads to the conclusion that power gain is the logical basis for design. In order to formulate an expression for power gain, a relation must be established between the radiation resistance with uniform current distribution and radiation resistance with exponential current distribution. The expression derived by Lewin is compared to that derived by Zuhrt. Gain curves are drawn using the formulas, and the "maximum output" and lobe alignment design are compared on the basis of gain with the design based on maximizing the vertical pattern function at the desired angle of radiation. It is found that only under certain conditions does the "maximum output" design have greater gain for the same leg length. The maximum gain condition is discussed together with optimum termination loss, attenuation rates, and surge impedance formulas for multiple wire rhombics.

In general, the analysis does not invalidate the design conditions previously derived, but rather increases the emphasis on the general alignment condition and gain and decreases the importance of the "maximum output" and "maximum alignment condition" which were derived on the basis of a constant input current and uniform current distribution.

Communications

Directivity of a Broadside Array of Isotropic Radiators—H. E. King (p. 197)

Directivity curves, of a uniform broadside array of isotropic sources, are shown to illustrate that gain increases almost linearly with aperture size until an optimum source spacing of approximately 0.9 λ is reached.

Modification of "Simplified Method for Computing Knife Edge Diffraction in the Shadow Region—L. J. Anderson and L. G. Trolese (p. 198)

Effect of Surface Reflections on Rain Cancellation of Circularly Polarized Radars—R. McFee and T. M. Maher (p. 199)

Laboratory Development Notes—Omnidirectional Vertically Polarized Paraboloid Antenna—E. O. Willoughby and E. Heider (p. 201)

Contributors (p. 204)

Audio

Vol. AU-7, No. 2, March–April, 1959

The Editor's Corner—"Third Person Passive"—Marvin Camras (p. 25)

PGA News (p. 26)

Audio Amplifier with Reduced Plate Dissipation—R. B. Dome (p. 29)

An audio frequency amplifier arrangement is described which results in decreased plate dissipation. The arrangement is suitable for either class A or B amplifiers. The scheme employed is to feed the grid of the amplifier an auxiliary signal as well as the desired audio frequency signal. The auxiliary signal is at a super-audible frequency and is automatically adjusted so that its peaks in the positive direction never cause the peak plate current to exceed the maximum current peak of the desired signal, nor shall its amplitude be so high as to cause clipping of the auxiliary signal on its negative peaks; in other words, the added wave should not affect the *average* value of the current generated by the desired signal. The plate circuit has its regular low-frequency load, but it also has a load circuit tuned to the super-audible frequency. The super-audible output may be dissipated as heat in a resistor.

Calculations have been made which show that for a given input and output at low frequencies, the maximum plate dissipation of a class A amplifier may be reduced to as low as 41 per cent of the maximum dissipation attained in conventional class A operation, and that the maximum plate dissipation of a class B amplifier may be reduced to as low as 50 per cent of the maximum dissipation attained in conventional class B operation.

Three Channel Stereo Playback of Two Tracks Derived from Three Microphones—P. W. Klipsch (p. 34)

Playback of two-track stereo source material with a derived center channel offers accurate reproduction of the original stereo geometry, and requires very simple implementation.

Essentially this two-track three-channel stereo depends on the principle that if two microphones are properly placed relative to each other and to the plane of the sound source, their combined output is that of a single microphone in the center, which output may be recovered by recombination of the two tracks.

When a physical third microphone in the center is employed to feed the two tracks, its recovery from two tracks depends on relative polarity and amplitudes and in one recombination method the center microphone could be cancelled instead of being reproduced.

An all-pass network may be used to shift the phase of one track so that on recombination the physical third microphone is always recovered, regardless of the manner in which it was mixed into the two tracks. The all-pass network produces 90° phase shift at only one frequency, but by choice of this frequency and the expectancy of additive polarity of original mixing, the center channel is recovered with excellent acuity, based on tests similar to those of Steinberg and Snow. Experimental recovery of a center output channel from a single center microphone feeding the two tracks, with flanking input signals zero, resulted in a center track output which was substantially indistinguishable from a normal monophonic reproduction.

Unilateral Transistor Amplifier—L. M. Vallese (p. 36)

A transistor audio amplifier with input impedance independent of the load and adjustable continuously to up a value of the order of $r_m/2$ is described. The amplifier consists of a unilateralized hook common collector configuration and has large power gain, very low output impedance. Its noise figure and dependence upon temperature, frequency, bias, and load are discussed.

Contributors (p. 40)

Broadcast and Television Receivers

Vol. BTR-5, No. 1, January, 1959

The Art of Getting Along—Wilfred Peterson (p. 1)

PGBTR Administrative Committee (p. 2)

Minutes of Administration Meeting, October 28, 1958 (p. 3)

A New AGC Circuit—F. J. Banovic and R. Miller (p. 5)

In radio an AGC circuit has three, sometimes conflicting, functions to perform. It should

1) be delayed enough so as not to affect sensitivity.

2) maintain a fairly constant signal strength at the input to the second detector to maintain a constant audio output under conditions where "fading" due to atmospheric variations is occurring. Automobile radios have an extra fading condition due to hills, bridges, varying ground conditions, fluctuating voltages, etc.

3) have enough control to limit the signal strength on strong signals so that overloading does not occur in the circuits preceding the manual volume control. That is, under conditions such as may be encountered near a strong transmitting station the circuits should not be driven past the non-linearity regions. Anyone who has driven a car close to a transmitter antenna with a radio tuned to that station has experienced the effect of RF overload.

Generally, the more stages that can be controlled by the AGC system, the better it can handle the preceding three conditions. This is obvious by considering that a small change in the control voltage will cause a large change in the system gain. In small transistorized receivers it is usually only possible to apply AGC control voltage to one stage. The reasons for this are discussed.

A block diagram of a typical six-transistor portable radio is shown. The circuits pertaining to AGC are analyzed in detail.

The Fuse in the Horizontal-Deflection System—W. Feingold (p. 10)

The role of the fuse in the horizontal-deflection circuit is discussed and a simple design change to reduce the current through it indicated. This modification allows the selection of a smaller fuse offering better protection on the one hand and greater freedom from sporadic failure on the other.

Application of RCA Drift Transistors to FM Receivers—J. W. Englund and H. Thanos (p. 13)

This paper discusses the application considerations involved in the use of drift transistors in the radio-frequency amplifier, oscillator, and intermediate-frequency amplifier stages of battery-operated FM broadcast receivers. Receiver design and performance are discussed in terms of individual stages, and data are presented on gain, bandwidth, signal-to-noise ratio, and frequency stability. The effects of ambient-temperature and supply-voltage variations are also described.

Television IF Selectivity and Adjacent Channel Interference—T. Matzek (p. 18)

One important aspect of a television IF amplifier is its selectivity, especially the ability to reject the adjacent channels. The maximum attenuation that is effective is limited by cross-modulation in the tuner, but this attenuation is seldom achieved. A novel trap circuit is presented that provides substantial improvement.

The Optimum Source Impedance and Noise Figures of Television Input Tubes with Various Circuits—L. E. Matthews (p. 22)

Measured optimum source impedances and noise figures are given for neutralized and unneutralized triodes and a pentode. Constant noise figure contours in the input impedance plane are given for each case at various frequencies. Measurement techniques are described and results discussed.

Television Receiver Color Decoder Design—D. Richman (p. 27)

This paper presents design principles and circuit details for a new color decoder which is intended to produce a high level of performance at reasonable cost.

The report presents 1) a discussion of design principles internal to the decoder; 2) a discussion of external or environmental factors such as IF shapes and burst gating; 3) a complete circuit diagram; and 4) measurements of performance characteristics.

The decoder is supplied with composite chrominance-plus-burst and a burst gate, and provides three color-difference outputs for application to a tricolor tube.

This decoder is demonstrably easy to tune and adjust in a receiver and requires a minimum of readjustment.

Technical features of interest include wide equiband chrominance, coupled equiangular demodulators, a high-gain APC loop with a solution to the balanced phase-detector problem, and a new simple control system called "Synchronous ACC" which uses a single DC control for synchronous automatic color-monochrome switching, noise-immune ACC, and two-mode sync. An accurate burst-gating system which is independent of horizontal phasing is also described.

The comparative ease of tuning the receiver with this decoder is due to the following characteristics: 1) the ACC system has high gain and does not affect hue; hence, the fine tuning is basically independent of the hue and saturation (the controls for hue and saturation need a minimum of "excess range"); 2) the APC loop has high DC gain, so that only limited hue control range is needed; 3) the APC and ACC systems are noise-immune and do not "cross-couple" their control effects—hence, the color controls do not need to be re-set with changes in signal level or channel; 4) because an accurate burst-gating arrangement is available, the color-control functions may be made independent of horizontal phase adjustment and less dependent on tuning; 5) as a consequence of the balanced detection system and the improved APC loop, pull-in is rapid and reliable; 6) the simple demodulation system and accurate demodulation angles preclude any question of correctness of color phase; and 7) the color-kill system is simple and reliable, requires no adjustments, and hence needs no tolerance potentiometers.

Detection of Asymmetric Sideband Signals in the Presence of Noise—T. Murakami and R. W. Sonnenfeldt (p. 46)

The first part of this paper analyzes three methods of detection of signals contaminated by fluctuation noise: 1) linear envelope detectors, 2) product detectors, 3) exalted carrier detectors.

A new ratio, the "video-to-noise-error ratio" is proposed for a more adequate quantitative evaluation of detector performance. Its use is justified by a large number of theoretical curves and experimental results. The theoretical and practical results show that an improvement in the output video-to-noise-error ratio is obtained by substitution of a product detector for an envelope detector. Improvements of 11 decibels have been measured. The second part of this paper analyzes the effects of impulse noise on the detection of asymmetric sideband signals. It is shown that product detectors eliminate the rectification of the noise envelope found in envelope detectors. This is of particular interest in asymmetric sideband systems where the noise output will be at relatively high frequencies. These can often be separated from lower frequency signals, making possible efficient impulse noise suppression.

Television Wireless Remote Control—R. Muniz (p. 76)

Basically, the wireless remote control device described is designed for convenience. To truly serve this purpose and to give adequate control of a television receiver from a remote point without wires, it is necessary to provide control over the essential customer adjustments. Therefore, the target or design objective was to provide the customer with means for selecting previously-programmed channels, for turning the set off or on, for having continuous control over picture brightness, and for continuous control over audio volume. It is also desirable to have the audio portion of the television receiver muted during the channel selection interval because otherwise the customer would be subjected to various disturbing sounds. With modern low-drift tuners having individual channel adjustments, adjustment of fine tuning from the remote point is probably unnecessary. Observation of television receivers under home operating conditions led to the selection of brightness control rather than contrast control, because it was found that with the modern keyed AGC circuits employed in the receivers to which this remote control would be applicable the need for customer adjustment of contrast was considerably minimized but that adjustments to compensate for room lighting conditions were brought about largely through adjustment of the brightness control. Obviously, to be commercially practical any system for providing this convenience has to be both relatively inexpensive and almost perfectly reliable. To this end an investigation of the various means for achieving the required end result were explored more or less fully.

Communications Systems

Vol. CS-7, No. 1, May, 1959

Quo Vadis—E. N. Dingley, Jr. (p. 1)

Frontispiece—Edward N. Dingley, Jr. (p. 2)

Performance of Digital Phase-Modulation Communication Systems—C. R. Cahn (p. 3)

This paper analyzes the performance of digital phase modulation systems in Gaussian noise and determines required signal-to-noise ratio as a function of the number of discrete phases and the desired error rate, under conditions of no fading. Both coherent detection with a locally-derived reference carrier and phase comparison detection are considered. The calculations show that multiphase modulation provides an efficient trade of bandwidth for signal-to-noise ratio in comparison with multilevel amplitude modulation. It is also found that phase comparison detection introduces about a 3-db degradation over coherent detection except with binary modulation, for which the degradation is less than 1 db for error rates not exceeding about 0.001.

Radio Channel Selection for Interference-Free Operation—J. Awramik, Jr. and W. M. Jewett (p. 7)

A procedure for the selection of mutually-interference-free (MIF) radio communication channels has been devised. Cross-modulation interference and various forms of intermodulation products were considered, as well as other interferences representative of real situations. The channel-selection process consists of the assignment of a priority to all previously-accepted channels over all later trial channels that result in an interference-producing condition. This results in the maximum number of MIF channels that can be obtained by any presently-known practical method from a given available bandwidth. Use of the procedure indicates that a 40–60-per cent increase in the maximum number of MIF channels may be achieved by reallocation of frequencies in the 225–400-mc band. Reassignment would not penalize any of the services entitled to use of the military VHF/UHF band. The use of a slide rule "Radio Interference Calculator," developed specifically for the problem, considerably reduces the computations required for MIF channel selection.

Performance Analysis of a Data Link System—A. B. Glenn (p. 14)

Because of the increased traffic densities and aircraft speeds contemplated in the near future, the present air-traffic-control system must be improved by a far more efficient and partially automatic system. To prevent further burden on already-burdened voice-communication facilities and frequencies, there is a requirement for more efficient means of transferring information between air and ground. This paper deals with the modulation and detection characteristics of a time-division multiplexed, digital radio link intended as an initial step in providing relief to the ATC communication problem.

As a result of both analytical and experimental work, two preferred modulation types have been chosen for further investigation. These types are frequency-shift-keyed amplitude modulation (FSK-AM) and frequency-shift-keyed carrier modulation. This analysis shows that there is little difference in performance of the FSK-AM and FSK systems for system frequency instabilities as low as 5 kc. Improvement of the system frequency instability to values below 5 kc shows a rapid improvement of the FSK system over the FSK-AM system. RCA is developing both systems for the Bureau of Research and Development, Federal Aviation Agency, in its AGACS (Experimental Automatic Ground/Air/Ground Communication System) project so that the Bureau can establish, in an operational environment, a comparison of the experimental and theoretical results.

Doppler Phenomena in Space Communications—F. J. Tischer (p. 25)

The Doppler phenomenon is studied under space-flight conditions. These conditions impose restrictions and introduce influences which otherwise are of negligible importance. Examples of undesired effects of the Doppler shift in space communications and of applications for useful purposes are given.

General relations are derived for the Doppler shift which permit consideration of the space-flight conditions. These conditions, one of which is flight through ionized medium and electron clouds, cause deviations from a regular Doppler shift. The deviations can be considered as errors, or they can be measured and explored for obtaining data about electron densities in space. Evaluation of Doppler measurements in satellite tracking is considered as an example.

Performance Equations for a "Stationary" Passive Satellite Relay (22,000-Mile Altitude)

for Communication—M. Handelsman (p. 31)

Performance equations are given for the use of a single fixed satellite used as a passive bounce point for almost hemispherical coverage of the earth for communication purposes. The antenna size and transmitter power per cps of bandwidth for desired SNR at RF are given for various types of passively-reflecting satellites. Possible limitations due to terrestrial and extraterrestrial noise and future possibilities using low-noise receivers are discussed.

An Analog Computer for Finding an Optimum Route Through a Communication Network—H. Rapaport and P. Abramson (p. 37)

This paper describes an analog computer capable of solving a variety of communication network problems; in particular, the problem of finding an optimum route through an arbitrary network. For example, in routing a call through a communications network, it may be desirable to determine that path (or paths) containing the smallest number of switching centers—or again, if in some predetermined sense, weighting factors have been assigned to each link in the network, it is then possible to determine that path over which the summation of the weights of the links is a minimum.

A 16-node multiply-connected prototype has been designed in which "time" is used as the analog of link weights. The utilization of this prototype to find minimum paths and the relative merit of alternate paths is described. The prototype also has the capability of simulating network vulnerability (link or node inhibition of destruction).

Prelimiting Band-Pass Filtering on Fading Radio Circuits—C. Buff (p. 42)

On channels subject to multipath propagation, the prelimiting bandpass filtering technique is shown to be a factor of great importance in determining the over-all performance. Minimum-loss, flat-topped filters create serious transient effects which add greatly to the error rate in data transmission. Because of adjacent channel requirements, an over-all Gaussian response is not applicable. A practical solution is considered to be a rounded-top filter composed of a series of synchronous-tuned single circuits in cascade. Tests and observations are described, leading to the application of these filters on an independent side band (ISB) telegraph circuit. A threefold improvement in the error rate of 45-baud printer operation resulted, as well as reduction of distortion for time-division multiplex operation.

Radio Link Communication Reliability: A Three-Part Design Problem—M. W. Green

The degree of reliability required for equipment used in a global communications system must be designed into the equipment from the inception of the original design specifications and implemented through all stages of development, procurement, production, and testing.

The design of a system based on three basic concepts, redundancy, rugged circuit design, and complete monitoring and alarm facilities, should result in a system capable of successful operation from the standpoint of system reliability in a global communication network.

Included are criteria for use by the engineer in designing equipment to attain the desired goals of optimum reliability and minimum maintenance.

Examples of actual equipment applications using one or more of these concepts are described, together with performance data obtained with systems using some of these techniques to indicate the improvement in system traffic reliability actually achieved.

Analysis of SSB Power Amplifiers—F. Assadourian (p. 53)

It is well known that one of the critical items in a single sideband (SSB) transmitter is the power amplifier with which are associated the basic problems of distortion, output power and plate dissipation. Tubes are available which may be used either as quasi-linear RF amplifiers operated class AB or B or as grid-modulated amplifiers operated class C. In the class-AB case, the SSB signal is introduced directly on the grid in modulated form. In the class-C case, the low-frequency envelope and high-frequency phase-modulated carrier of the SSB signal are initially separated and subsequently recombined through remodulation at the grid of the power tube. The performance of the power amplifier in these two modes of operation is compared theoretically for idealized tube characteristics. The analysis assumes in particular a quasi-linear plate current-grid voltage characteristic and is based on a two-tone SSB wave.

The results of the analysis are the following:

1) Class-B operation yields more average output ac power for a given plate supply voltage or peak envelope RF power and less for a given average plate dissipation. Efficiency and power tables are given. Typical normalized figures for average ac power output and average plate dissipation for two-tone SSB are 0.39 and 0.24 for class B and 0.26 and 0.11 for grid-modulated class C if plate voltage is fixed. The class-C figures change to 0.32 and 0.14 for equal peak envelope powers and to 0.57 and 0.24 for equal amounts of average plate dissipation.
2) Class-B operation is distortionless for two-tone SSB because of the idealized quasi-linear plate current-grid voltage characteristic.
3) The maximum distortion in class-AB operation is slightly less than that in grid-modulated class-C operation for two-tone SSB. Distortion tables are provided for the class-AB and class-C cases which are applicable when these two cases correspond to equal peak envelope powers. Typical spurious tones in the two-tone SSB output have maximum voltage amplitudes of around 0.08 for class-AB operation and 0.11 for grid-modulated class-C operation when normalized with respect to the output voltage amplitude of either equal tone.

Two-Dimensional Predictive Redundancy in a Television Display—A. V. J. Martin (p. 57)

The two-dimensional predictive redundancy of a television picture element is calculated as a function of the preceding point correlation for both interlaced and noninterlaced scanning. The resulting curves level off rapidly when the number of previous points taken into account is increased. This seems to indicate that while appreciable improvement would accrue if preceding point redundancy were used, there would be little more to earn by including the redundancies due to more distant points, especially in view of the necessarily increased complexity of circuitry.

An analysis of three-dimensional redundancy would probably result in similar conclusions.

Medical Electronics

Vol. ME-6, No. 1, March, 1959

Proceedings of the Conference on Methodology and Problems in Artificial Internal Organs, January 15, 1958.

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Microwave Theory and Techniques

Vol. MTT-7, No. 2, April, 1959

Message from the Editor (p. 188)
Frontispiece—Seymour B. Cohn (p. 189)
Breaking Through the Mental Barrier—Seymour B. Cohn (p. 190)

Reflection of a Pyramidally Tapered Rectangular Waveguide—K. Matsumaru (p. 192)

The reflection coefficient Γ of a pyramidally tapered rectangular waveguide is derived by assuming that the taper impedance is proportional to the height and guide wavelength and inversely proportional to the width of the taper cross section. It is shown that the loci of Γ, plotted in the K plane as a function of taper length for some conventional tapers, do not pass through the center of the chart at multiples of a half-guide wavelength as for an exponential line, but instead they converge almost concentrically. The frequency characteristic of the pyramidally tapered waveguide is compared with other types of tapers. Typical 7-kmc experimental results for several tapers differing in length are presented.

Cascade Directional Filter—O. Wing (p. 197)

A directional filter is a completely matched four-port which exhibits a directional and a filter-like frequency characteristic. This paper explores the properties of N-directional filters connected in cascade through sections of transmission lines. Analysis shows that if a directional filter admits the equivalent circuit representation offered here, its transfer functions are functions of only one parameter, a susceptance function. When the directional filters are cascaded in a certain way, the over-all transfer functions have the same form as before except that the susceptance function is now the sum of the susceptance functions of the component filters. The last property is an important one. Given a transfer function expressed in terms of a susceptance function, the network designer can expand the susceptance in partial fraction and realize the transfer function using directional filters in cascade, each being characterized by a much simpler susceptance.

Propagation in a Dielectric-Loaded Parallel Plane Waveguide—M. Cohn (p. 202)

A theoretical analysis of wave propagation in a parallel plane waveguide partially filled with a dielectric is performed. This transmission line is a symmetrical three-region structure consisting of two infinite parallel conducting planes with a dielectric slab of rectangular cross section between and contacting each of the planes. It has been found that TEM and TM modes cannot propagate on this structure. This investigation is concerned with TE modes, although hybrid modes can also propagate on this line. The lowest order TE mode, which is the dominant mode, has no cutoff and hence is inherently suited to extremely wide bandwidth operation. Equations have been presented for the field components, guide wavelength, cutoff criteria, power handling capabilities, wall losses, and dielectric losses as a function of the operating wavelength, waveguide dimensions, and material constants. In the case of the dominant mode, design curves covering a large range of wavelengths, dimensions, and dielectric constants are presented. For a loosely bound wave, the losses are comparable or less than those of conventional rectangular waveguide and the power handling capacity is an order of magnitude greater.

Electromagnetic Backscattering Measurements by a Time-Separation Method—C. C. H. Tang (p. 209)

The object of this research is to investigate the feasibility of adapting the conventional pulsed radar technique for close range backscattering measurements for obstacles of arbitrary shape and small scattering cross sections. The time-separation or microwave-pulse method described in this paper differs essentially from all previously used laboratory methods in that the scattered field does not mix with the incident field at the detector and is separated from it in time. The essential experimental arrangement of this method is similar to that of the CW magic-T method except that a source generating very short pulses is used instead of CW. Preliminary experimental data for thin circular metallic disks show that the pulse method is a feasible one, since the measured results are in close agreement with the theoretical values. Accurate back-scattering measurements for obstacles of arbitrary shape and small scattering cross sections should be obtainable by this method provided a short microwave pulse of high power level is available.

On Network Representations of Certain Obstacles in Waveguide Regions—H. M. Altschuler and L. O. Gladstone (p. 213)

Network representations for a class of obstacles in waveguide regions when the diffraction problem is of a vector type can be obtained by the use of E- and H-type modes. The special properties of these modes are discussed and highlighted by an example involving the network representation of a periodic strip grating in free space for oblique incidence. Transformations relating the different networks based on various modal representations in rectangular coordinate systems are also discussed.

Reflectors for a Microwave Fabry-Perot Interferometer—W. Culshaw (p. 221)

The advantages of microwave interferometers for wavelength and other measurements at millimeter wavelengths are indicated, and a microwave Fabry-Perot interferometer discussed in detail. Analogous to the cavity resonator, this requires reflectors of high reflectivity, small absorption, and adequate size. Stacked dielectric plates, and stacked planar or rod gratings are shown to be suitable forms of reflectors, and equations for the reflectivity, optimum spacing, and bandwidth of such structures are derived. A series of stacked metal plates with regularly spaced holes represents a good design of reflector for very small wavelengths. Fringes and wavelength measurements at 8-mm wavelength are given for one design of interferometer, these being accurate to 1 in 10^4 without any diffraction correction. For larger apertures and reflectors in terms of the wavelength, errors due to diffraction will decrease.

Precise Control of Ferrite Phase Shifters—D. D. King, C. M. Barruck, and C. M. Johnson (p. 229)

Hysteresis and thermal drifts can prevent accurate calibration of ferrite phase shifters. To provide a precise setting of phase in response to a control signal a servo system has been developed. This system utilizes a control frequency to determine uniquely the phase shift in a ferrite element. The desired phase shift is then a function only of control frequency and line length. Performance data are given for various operating conditions of the control system.

Tables for Cascaded Homogeneous Quarter-Wave Transformers—L. Young (p. 233)

Quarter-wave transformers are frequently required in microwave and UHF systems. An exact design procedure is known but involves lengthy calculations. Faced with the design of many such transformers, the calculations were programmed on an IBM 704 digital computer. The speed of computation is such that several hundred designs for 2, 3, and 4 section transformers were systematically computed in a few minutes. The results are reproduced here in tables, which should permit the calculation of most cases of practical interest by interpolation.

The Symmetrical Waveguide Synthesis of Circulators—B. A. Auld (p. 238)

A method for synthesizing symmetrical waveguide circulators by adjusting the eigenvalues of the scattering matrix is described. This procedure is particularly useful for the design of very compact circulators in the form of waveguide junctions containing ferrite obstacles. Permissible structural symmetries for a circulator are listed, and a standard form for the scattering matrix of a symmetrical circulator is defined. The synthesis procedure is then described in detail, stating the conditions to be imposed on the scattering matrix eigenvalues, and an expression is obtained for the changes in the eigenvalues due to the placing of anisotropic material within the junction.

By applying the theory to Allen's 4-port turnstile circulator, it is shown that the use of a matched turnstile junction and a reflectionless Faraday rotator is not essential. The theory is also applied to the design of novel 3- and 4-port circulators, and two 6-port circulators, one of which may be used as a 5-position waveguide switch, are described. Some experimental results are presented for a compact 3-port circulator in the form of an H-plane Y junction, in 1 inch by ½ inch waveguide, containing a ferrite post obstacle. This circulator, which operates with a bias field of approximately 25 oersted, has a useful bandwidth of 3 per cent. Greater bandwidths would be expected in a Stripline or a fin-line version of this device.

Delay Distortion in Crystal Mixers—T. Kawahashi and T. Uchida (p. 247)

Delay distortion is one of the most important characteristics in the frequency-modulated supermultichannel microwave repeater. With regard to receiving crystal mixers, the cause, shape, vanishing condition, etc., of delay distortion are analyzed, and the experiments show good agreement with the results of this analysis.

To eliminate this delay distortion, the electrical length between the crystal and the image suppression filter must be determined so that the image frequency impedance may not be infinite in a desired frequency band, or the intermediate frequency load impedance must be fixed at a certain definite value.

The Efficiency of Excitation of a Surface Wave on a Dielectric Cylinder—J. W. Duncan (p. 257)

This paper presents a theoretical and experimental study of the excitation of the lowest order TM surface wave on an infinite dielectric cylinder. The source is a circular filament of magnetic current within the dielectric rod. The integral solution for the field is evaluated as a contour integral by applying Cauchy's theorem. The far zone radiation field is obtained by means of a saddle point integration. Curves are presented which show excitation efficiency as a function of k_0a, the normalized circumferential length of the filament. A filament 0.83 wavelength in diameter will launch the TM mode with an efficiency of 95 per cent. A narrow annular slot in a large metal sheet was used to approximate the magnetic current filament and efficiency was measured using Deschamps' method for a two-point junction. The experimental measurements verify the theoretical analysis. In addition, it was found that the slot launching efficiency was essentially independent of the ground plane dimensions.

Proposal for a Tunable Millimeter Wave Molecular Oscillator and Amplifier—J. R. Singer (p. 268)

An atomic beam apparatus suitable for a millimeter wave generator is theoretically discussed. The beam consists of atoms having a net magnetic moment. The upper and lower Zeeman levels of the atomic beam in a magnetic field are spatially separated by an inhomogeneous magnetic field. The upper state atoms enter a cavity where transitions occur at a frequency determined by a static magnetic field. The resonant frequency of the cavity is set at the transition frequency. The positive feedback of the cavity allows operation as an oscillator. Some of the more important parameters for oscillator operation are evaluated. The upper frequency limit is determined primarily by the resonant structure design.

High-Speed Microwave Switching of Semiconductors—R. V. Garver (p. 272)

A relationship between low-power isolation and small-signal, low-frequency diode resistance is reported. A study of ambient heating indicates that with increasing temperature the diode characteristics tend to approach the line characteristic of the above relationship. Observed switching speeds of 1.5 to 3.0 mμs are reported. A theory is presented which agrees with the switching time data and predicts microwave switching times as low as 0.2 to 0.3 mμs. High speed switching is discussed with reference to significant parameters, *e.g.*, hole storage, internal heating, and pulse reverse diode characteristics.

A Logarithmic Transmission Line Chart—A. C. Hudson (p. 277)

A chart is presented which relates the real and imaginary components of the impedance at any position along a transmission line to the magnitude and location of the standing wave. In the present chart the ordinate is R/Z_0 plotted logarithmically and the abscissa is a function of X/R. Thus a change in the reference impedance becomes a simple vertical translation of any point. An auxiliary chart permits the direct determination of the length and impedance of transmission line required to match a given impedance.

The Far Fields Excited by a Point Source in a Passive Dissipationless Anisotropic Uniform Waveguide—A. D. Bresler (p. 282)

The direction of the net power flow associated with a propagating mode of an arbitrary passive dissipationless anisotropic uniform waveguide may be opposite to its direction of (phase) propagation. It is shown that when a point source is introduced into a waveguide in which this is the case, such propagating modes contribute to the fields excited by this source only in that direction for which their power flow is directed away from the source. In addition it is shown that the nonpropagating modes contribute to the total field only in that direction in which they decay with increasing distance away from the source to that the far fields are given by a super-position of propagating modes only. The proof given makes use of the known properties of the frequency dependence of the physical parameters of any linear passive system in which the causality restriction is satisfied.

Analysis of a Negative Conductance Amplifier Operated with a Nonideal Circulator—E. W. Sard (p. 288)

Negative conductance amplifiers are usually operated with a circulator in order to achieve greater gain-bandwidth products and stable operation. Typical circulators differ from ideal circulators in that the forward loss between ports is not zero, and the reverse isolation between ports is not infinite. The main effects of noninfinite isolation are shown to be a modified gain-bandwidth product and a change in output admittance of the circulator output port. These effects result principally from the finite isolation between the output and amplifier ports. The main effect of incidental dissipation has previously been shown to be an increase in system noise figure.

This paper considers only the effects caused by noninfinite isolation. A model of a lossless three-port circulator with noninfinite isolation is set up, and a negative conductance amplifier is considered to be connected to one port of this circulator. The magnitude of negative conductance is assumed to be limited to ensure a positive output conductance at the output port of the circulator (that is, the combination of negative conductance amplifier and nonideal circulator is assumed to be open-circuit stable). Subject to this assumption, the combination of negative conductance amplifier and nonideal circulator is then analyzed for its output admittance, available power gain, and effective input noise temperature.

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1959 National Symposium Program (p. 303)

Nuclear Science

Vol. NS-6, No. 1, March, 1959

From the Editor—R. F. Shea (p. 1)

A Bias Function Generator for the Zero Gradient Synchrotron—L. K. Wadhwa (p. 2)

An Analog function generator producing a biasing current for the permeability-tuned master oscillator of the zero gradient synchrotron is described. It produces the bias function with a precision of ± 0.02 per cent and is capable of delivering about 3 amperes of current into a predominantly resistive load of a quarter of an ohm. The input information to the function generator is obtained from proton accelerator parameters such as the magnetic guide field B_g and $\dot{B}_g$ and the guide magnet current I_g and $\dot{I}_g$. The total bias function is obtained as a result of adding outputs from the 1) B_g readout, 2) hump generator, and 3) biased diode network.

A specially developed electronic system and an analog-to-digital converter have been employed to make static and dynamic measurements with a view to determining the precision of the function generator.

Servologarithmic Amplifier for Reactor Instrumentation—W. J. Hartin (p. 11)

The instrument described is a servo which provides an output proportional to the logarithm of the input over 4 decades. Use of a special four-decade exponential feedback potentiometer makes the accuracy and repeatability much better than that obtainable with the usual logarithmic amplifiers. A period trip circuit is part of the instrument.

Study of the Feasibility of a Ferrite Modulation System for an FM Cyclotron—K. Enslein (p. 14)

This paper presents an analysis of a ferrite modulation system for a synchrocyclotron. It is shown that such a system is feasible at the present state of the art. The analysis is performed by numerical computation and is substantiated by a scaled experiment. Extensive data are obtained from the computation and are available in microfilm form. The computer program is listed.

Nanosecond Pulse Transformers—C. N. Winningstad (p. 26)

The transmission-line approach to the design of transformers yields a unit with no first-order rise-time limit since this approach uses distributed rather than lumped constants. The total time delay through the transmission-line-type transformer may exceed the rise time by a large factor, unlike conventional transformers. The extra winding length can be employed to improve the low-frequency response of the unit.

Transformers can be made for impedance matching, pulse inverting, and dc isolation within the range of about 30 to 300 ohms with rise times of less than 0.5×10^{-9} seconds, and magnetizing time constants in excess of 5×10^{-7} seconds. Voltage-reflection coefficients of 0.05 or less, and voltage-transmission efficiencies of 0.95 or better can be achieved.

A Chronotron for Relativistic Neutron Time-of Flight Measurements— R. H. Ragsdale and W. F. Stubbins (p. 31)

Neutrons in pulses produced at a 19 mc rate by the circulating beam of the 184-inch cyclotron striking an internal target have kinetic energies from 100 to 800 mev. Circulating-beam bunching limits the time duration of each neutron pulse. Reference signals are produced by a Çerenkov-radiator internal target or the radio-frequency voltage on the cyclotron dee. The chronotron coincidence circuits, capable of a resolving time of less than 10^{-9} second, are spaced with increasing time intervals to reduce reflection reinforcement. Saturation of coincidence circuits minimizes the dependence on signal amplitude. The 125-Ω cable is matched at sampling points by a series inductance tapped by tube input capacity.

Ion Phase Measurement Techniques on the Birmingham Cyclotron—M. Konrad (p. 35)

The experimental technique used for internal beam ion phase measurements on the Birmingham cyclotron is described. Ion phase measurements were carried out by measuring the target current waveform of a screened two-electrode target with an equipment based on a sampling technique, having a resolution of 1.5 mμsec and a sensitivity of 5 μa/cm deflection. The target current waveform was displayed on a cathode-ray-tube screen against the sine of the cyclotron RF phase or in polar coordinates.

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Electronic and Radio Engineer*, incorporating *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

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ACOUSTICS AND AUDIO FREQUENCIES

534.522.1 **1751**

Diffraction of Light by Ultrasonic Waves—Oblique Incidence and Sound Intensity—S. Parthasarathy and C. B. Tipnis. (*Nature, London*, vol. 182, pp. 1795–1796; December 27, 1958.) The diffraction is considerably modified if the intensity of the sound is increased above that observed in the previous experiment (1049 of April).

534.612-8:537.228.1 **1752**

Piezoelectric Crystal Probe for the Measurement of Ultrasonic Power and the Investigation of an Ultrasonic Field—N. Ségard, J. Cassette, and F. Cocquerez. (*Compt. rend. Acad. Sci., Paris*, vol. 247, pp. 873–876; September 22, 1958.) An X-cut quartz crystal is used in a triode circuit.

534.78 **1753**

Experiences Gained in the Development of a Vocoder and the Measurements of Intelligibility Achieved with it—G. Krohm. (*Z. angew. Phys.*, vol. 10, pp. 56–65; February, 1958.) An intelligibility of about 80 per cent was achieved with the equipment described using a nominal bandwidth of 300 cps.

534.79 **1754**

Subjective Measurements of the Influence of Peak Content in Band-Pass Noise on the Sensation of Loudness—H. Niese and J. Köhler. (*Hochfreq. und Elektroak.*, vol. 66, pp. 150–160; March, 1958.) Tests were made with two types of noise of variable bandwidth, *a*) filtered white noise, and *b*) noise generated using an FM method to eliminate amplitude fluctuations which might give rise to conflicting assessments of loudness. See also 3381 of 1957 (Zwicker *et al.*).

The Index to the Abstracts and References published in the PROC. IRE from February, 1958 through January, 1959 is published by the PROC. IRE, May, 1959, Part II. It is also published by *Electronic and Radio Engineer*, incorporating *Wireless Engineer*, and included in the March, 1959 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

534.86:621.396.712.3 **1755**

The Acoustic and Technical Characteristics of Reverberation Plates—W. Kuhl. (*Rundfunktech. Mitt.*, vol. 2, pp. 111–116; June, 1958.) A device for producing artificial reverberation is described which consists of a large thin sheet of tin-plate electrodynamically excited in the flexural mode, the oscillations being picked up by a piezoelectric microphone. The reverberation time may be varied over the range 0.8–5 seconds by altering the distance between the oscillating plate and a movable porous damping plate.

534.861:621.396.712.3 **1756**

The Acoustic Design of the New Studio in Karlsruhe—L. Keidel. (*Rundfunktech. Mitt.*, vol. 2, pp. 106–110; June, 1958.) Details are given of the construction adopted to eliminate traffic noise.

621.395.61 **1757**

Theory of First- and Second-Order Gradient Receivers—C. Smetana. (*Hochfreq. und Elektroak.*, vol. 66, pp. 143–150; March, 1958.) The characteristics of differential microphones are derived with particular reference to their noise-suppression capabilities.

621.395.613.32 **1758**

Life Tests of the Microphone Carbon in Practical Uses—H. Hirabayashi, H. Toyoda, H. Shibata, and T. Ayakawa. (*Rep. elect. Commun. Lab., Japan*, vol. 6, pp. 211–220; June, 1958.) A study of the effects of different methods of preparing the carbon powder.

621.395.623.7:537.523.3 **1759**

The Operation of Ionic Loudspeakers—G. Bolle. (*Nachrichtentech. Z.*, vol. 11, pp. 172–178; April, 1958.) The theory and design of corona-wind loudspeakers are given, and their performance is compared with that of other types.

621.395.623.7:621.318.2 **1760**

Permanent Magnets in Audio Devices—R. J. Parker. (IRE TRANS. ON AUDIO, vol. AU-6, pp. 15–21, January/February, 1958.) Abstract, PROC. IRE, vol. 46, p. 1440; July, 1958.)

621.395.623.7.001.4 **1761**

Procedures for Loudspeaker Measurements—P. J. A. H. Chavasse and R. Lehmann. (IRE TRANS. ON AUDIO, vol. AU-6, pp. 56–67; May/June, 1958.) Translation of paper abstracted in 8 of 1954.

621.395.623.7.002 **1762**

Progress in the Construction of Loudspeakers—F. K. Schröder. (*Nachrichtentech. Z.*, vol. 11, pp. 169–172; April, 1958.) The improvement of performance characteristics by impregnation of the diaphragm and the fitting of copper rings in the air gap is discussed.

621.395.623.7.012:681.142 **1763**

Loudspeaker Enclosure Calculations—M. V. Callendar. (*Wireless World*, vol. 65, pp. 162–163; April, 1959.) An analog network is described for studying the performance of a loudspeaker in a bass-reflex cabinet.

681.84:534.851 **1764**

Pickup for Low Record Wear—J. Walton. (*Wireless World*, vol. 65, pp. 182–185; April, 1959.) A design for a crystal pickup to track within the elastic limit of the record is described The effective mass is 0.6 mg. See also 2330 of 1957 (Darlow).

ANTENNAS AND TRANSMISSION LINES

621.372.2:538.569.21.3 **1765**

Transmission Lines with Inhomogeneous Attenuation for the Low-Reflection Absorption of Electromagnetic Waves—K. L. Lenz. (*Z. angew. Phys.*, vol. 10, pp. 17–25; January, 1958.) Measurements were made on a recurrent network of 32 Π-sections to determine by an analog method the optimum distribution of losses in an absorber for a given minimum residual reflection. For calculations on homogeneous lines see 1068 of April (Lenz and Zinke).

621.372.8 **1766**

Nonlocal Reflection in Waveguides of Variable Cross Section—V. Pokrovskiĭ, F. Ulinich, and S. Savvinykh. (*Dokl. Ak. Nauk SSSR*, vol. 24, pp. 304–306; January 11, 1959.) A brief mathematical analysis of the reflection and scattering due to irregularities in the shape of a waveguide considered as a whole. See also 3326 of 1958.

621.372.8:621.3–71 **1767**

Waveguides for Use in Low-Temperature Cryostats—A Caistor, S. J. Fray, and W. C. Hopper. (*J. Sci. Instr.*, vol. 36, p. 144; March, 1959). A design which minimizes the conduction of heat into the cryostat.

621.372.831 **1768**

Reflection of Tapered Waveguides—K. Matsumaru. (*Rep. elect. Commun. Lab., Japan*, vol. 6, pp. 235–239; June, 1958.) An approximate study of the characteristics of conically tapered guides and rectangular-to-circular tapers.

621.372.832.8 1769
Low-Loss L-Band Circulator—F. R. Arams, and G. Krayer. (PROC. IRE, vol. 47, p. 442; March, 1959.) Insertion loss averages 0.3 db between 1200 and 1450 mc when the magnetic field around the ferrite is optimized for each frequency. Reverse isolation is ≥ 30 db, and input voltage SWR ≤ 1.11.

621.372.852.22 1770
Modes in Rectangular Guides Filled with Magnetized Ferrite—G. Barzilai and G. Gerosa. (*Nuovo Cim.*, vol. 7, pp. 685–697; March 1, 1958. In English.) The analysis is based on the characteristic equation for a rectangular waveguide with a ferrite slab placed against one side wall, the slab being magnetized in a direction parallel to the wall and perpendicular to the waveguide axis.

621.372.852.22 1771
Rectangular Guide Ferrite Phase Shifters Employing Longitudinal Magnetic Fields—P. A. Rizzi and B. Gatlin. (PROC. IRE, vol. 47, pp. 446–447; March, 1959.)

621.372.852.3 1772
The Homogeneous Rectangular Waveguide with Attenuating Foil—H. Buseck and G. Klages. (*Arch. elekt. Übertragung*, vol. 12, pp. 163–168; April, 1958.) The influence of an axial foil on the waveguide transmission characteristics is analyzed assuming metallic contact between foil and waveguide walls. Discrepancies between measured and calculated values are attributed to the absence of electrical contact between the surfaces.

621.396.67:621.315.668.2 1773
Aalen Television Tower of the Süddeutscher Rundfunk—(*Rundfunktech. Mitt.*, vol. 2, pp. 143–144; June, 1958.) A reinforced-concrete tower of 80-m height supports a 61-m mast for the VHF and television antennas.

621.396.676 1774
Slot-Antenna Array for Missiles and Aircraft—E. J. Wilkinson. (*Electronics*, vol. 32, pp. 56–57; February 27, 1959.) Circular polarization is achieved by combining a folded dipole and a slot radiator. Impedance properties and radiation patterns are given.

621.396.677:621.397.62 1775
A Second Band-III Programme?—The Aerial Problem—F. R. W. Strafford. (*Wireless World*, vol. 65, pp. 171–174; April, 1959.) Problems associated with the use of existing antennas for the reception of programs separated in frequency by no more than three channels are examined.

621.396.677.5 1776
Calculated Radiation Resistance of an Elliptical Loop Antenna with Constant Current—J. Y. Wong and S. C. Loh. (*J. Brit. IRE*, vol. 19, pp. 89–91; February, 1959.) Extension of the analysis given in 20 of January (Loh and Wong) to derive formulas for loops comparable in size to the wavelength.

621.396.677.73 1777
The Construction of Horn-Type Aerials with Parabolic Reflectors—L. Calligaris. (*Alta Frequenza*, vol. 27, pp. 410–432; June/August, 1958.) Design and constructional details and methods of erection are given for the antennas used in the radio links Milan-Palermo and Rome-Pescara.

621.396.677.83:621.396.65 1778
Some Remarks on Passive Repeaters—F. Cappuccini. (*Alta Frequenza*, vol. 27, pp. 263–268; June/August, 1958.) The use of the curves obtained by Jakes (1243 of 1953) for calculating the attenuation of reflectors in radio links is described.

AUTOMATIC COMPUTERS

681.142 1779
Digital Computers Available in Britain—C. H. Lees. (*Brit. Commun. Electronics*, vol. 5, pp. 942–949; December, 1958.) Information about 27 computers is given in tabulated form.

681.142 1780
Digital Differential Analysers—G. C. Rowley. (*Brit. Commun. Electronics*, vol. 5, pp. 934–938; December, 1958.) Principles of operation are described and details are given of the design of the Avro D.D.A. machine ADDAM II.

681.142:537.227 1781
Ferroelectrics and Computer Storage—M. Prutton. (*J. Brit. IRE*, vol. 19, pp. 93–99; February, 1959. Discussion pp. 100–102.) The polarization reversal process in ferroelectric single crystals and its application to data storage is reviewed. An optical system for reading information from a ferroelectric store is described.

681.142:538.221 1782
Magnetics for Computers—a Survey of the State of the Art—J. A. Rajchman. (*RCA Rev.*, vol. 20, pp. 92–135; March, 1959.) A review of the application of magnetic materials to storage and switching devices.

681.142:621.314.7 1783
Transistorized-Core Memory—R. E. McMahon. (IRE TRANS. ON INSTRUMENTATION, vol. I-6, pp. 157–160; June, 1957.)

681.142:621.374.33:621.314.7 1784
Transistors Provide Computer Clock Signals—S. Schoen. (*Electronics*, vol. 32, pp. 79–72; February 27, 1959.) Switching circuits capable of high speed and controlling peak currents capable of high speed and controlling peak currents up to 5 amperes are described.

681.142:621.395.625.2 1785
Digital Storage on Punched Tape—M. E. Theis. (*Trans. Soc. Instrum. Technol.*, vol. 10, pp. 178–182; December, 1958.)

681.142:612.395.625.3 1786
The Storage and Processing of Digital Data on Magnetic Tape—D. W. Willis. (*Trans. Soc. Instrum. Technol.*, vol. 10, pp. 182–189; December, 1958.)

681.142:621.396.822 1787
The Noise Problem in a Coincident-Current Core Memory—F. McNamara. (IRE TRANS. ON INSTRUMENTATION, vol. I-6, pp. 153–156; June, 1957.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.048.75 1788
The Printed Circuit—C. Brinkmann. (*Elektrotech. Z. Ed. B*, vol. 10, pp. 461–467; December 21, 1958.) Summary of manufacturing and assembly techniques.

621.316.5 1789
Circuits for Ternary Switching Variables—E. Mühldorf. (*Arch. elekt. Übertragung*, vol. 12, pp. 176–182; April, 1958.) Applications of a ternary switching algebra (see 1636 of May) are considered and the synthesis of a ternary adder is described.

621.318.4:621.318.134 1790
Design of Toroidal Coils with Ferrite Cores Operating in the Audio-Frequency Range—L. I. Rabkin and Z. I. Novikova. (*Radiotekh. Elektron.*, vol. 2, pp. 762–768; June, 1957.) A method of determining the optimum relation between core dimensions is outlined. This gives the minimum coil size for a given Q-factor, or the maximum Q-factor for a given size.

621.318.435.34 1791
Auto Self-Excited Transductors—E. L. Clarke. (*Instrum. Practice*, vol. 12, pp. 1093–1100; October, 1958.) Basic circuits are examined nonmathematically, considering only resistive loads.

621.318.57:621.396.963.3 1792
Coincidence Diodes gate Electronic Switch—J. B. Beach. (*Electronics*, vol. 32, pp. 66–68; February 20, 1959.) A transistor switching circuit for radar indicators is described. Six channels are used in each corodinate axis of a cro presentation.

621.319.4:537.56:538.63 1793
Hydromagnetic Capacitor—Anderson, Baker, Bratenahl, Furth, and Kunkel. (See 1853.)

621.319.4.011.4 1794
Accurate Determination of the Capacitance of Rectangular Parallel-Plate Capacitors—D. K. Reitan. (*J. Appl. Phys.*, vol. 30, pp. 172–176; February, 1959.) The subarea method is recast and applied to derive a universal curve for a square-plate capacitor. Values calculated by other methods are compared.

621.372.01 1795
Elements of Electronic Circuits: Part 1—Time Constant and Differentiation—J. M. Peters. (*Wireless World*, vol. 65, pp. 156–158; April, 1959.) First of a series of articles reviewing basic electronic circuits, with emphasis on physical explanations of their operation.

621.372.049.621.314.7 1796
Analogous Transistor System Design and Nodal Methods of Construction with Applications to Research Equipment and Prototype Evaluation—R. F. Treharne. (*Proc. IRE Australia*, vol. 19, pp. 319–347; July, 1958.) Transistor action and circuits are described in terms of the thermionic valve analogy. A modular technique of circuit construction based on this principle is described, in which stages are assembled individually using a method which simplifies their interconnection.

621.372.4:621.372.5 1797
On the Reactance Theorem—H. Wolter. (*Arch. elekt. Übertragung*, vol. 12, pp. 158–162; April, 1958.) Any passive quadripole with Foster-type short-circuit and open-circuit impedances at either end transforms any Foster-type two-terminal network into a Foster-type two-terminal network again. Polygons consisting of Foster-type two-terminal networks are also considered. See also 3006 of 1958.

621.372.5 1798
Group Delay and Group Velocity—W. P. Wilson. (*Electronic Radio Eng.*, vol. 36, pp. 145–146; April, 1959.) The concepts are defined and their relation to the transfer function of a network is given. See also 2004 of 1958 (Gouriet).

621.372.5 1799
Radio Engineering Use of the Minkowski Model of the Lorentz Space—E. F. Bolinder. (PROC. IRE, vol. 47, p. 450; March, 1959.)

621.372.5 1800
The Condition of Passivity for the Linear Quadripole with Complex Characteristic Impedances—E. R. Berger. (*Arch. elekt. Übertragung*, vol. 12, pp. 149–157; April, 1958.) Reciprocal and non-reciprocal two-port networks are considered.

621.372.5.029.6:621.317.341 1801
An Analysis of Lossy Symmetrical Qudaripoles in the Decimetre and Centimetre Wavelength Ranges using Voltage Node Displacements—F. Gemmel. (*Arch. elektr. Übertragung*, vol. 12, pp. 169–172; April, 1958.) See also 1114 of April.

621.372.54:621.397.62 1802
A Combined Pulse-Width Filter for Television Receivers—W. Schröder. (*Elektron. Rundschau*, vol. 12, pp. 115–118; April, 1958.) Synchronization by means of a combined integrating and differentiating filter network is described. Performance data are tabulated.

621.372.543.2 1803
Intermediate-Frequency Circuits with Three Coupled Resonators—G. B. Stracca. (*Alta Frequenza*, vol. 27, pp. 304–346; June/August, 1958.) The operation of triple-tuned wide-band band-pass filters is analyzed and design formulas are tabulated. For data on double-tuned filters see 3063 of 1957.

621.373.421.13 1804
Theory of the Crystal-Controlled Inductive Three-Terminal Circuit—G. Becker. (*Arch. eleckt. Übertragung*, vol. 12, pp. 183–191; April, 1958.) Conditions of oscillation, equivalent circuits, and methods of compensation for the crystal-controlled Hartley oscillator are discussed.

621.373.421.13:621.314.7:538.569.4 1805
Transistorized, Crystal-Controlled Marginal Oscillator—R. L. Garwin, A. M. Patlach, and H. A. Reich. (*Rev. Sci. Instrum.*, vol. 30, pp. 79–80; February, 1959.) Circuit details of a small, nonmicrophonic oscillator unit for nuclear-magnetic-resonance observations.

621.373.431.1 1806
Operating Conditions of the Symmetrical Multivibrator—N. A. Zheleztsov and M. I. Feigin. (*Radiotekh. Elektron.*, vol. 2, pp. 751–761; June, 1957.) An approximate method for the division of a multidimensional phase space into subspaces is described and is applied to analyze the operation of a symmetrical multivibrator taking account of parasitic capacitance and grid current. Three modes of operation are considered.

621.373.431.1 1807
Multivibrator with Negative Feedback—I. Sh. Libin. (*Radiotekh. Elektron.*, vol. 2, pp. 809–810; June, 1957.) A description is given of a multivibrator circuit with very good frequency stability.

621.373.52 1808
Improved RC Oscillator—L. H. Dulberger. (*Electronics*, vol. 32, p. 62; March 6, 1959.) A modified version of the bridged-*T* circuit described by Sulzer (2943 of 1951); this oscillator operates at a single frequency in the range 4 cps–350 kc.

621.373.52 1809
The Conditions for the Onset of Oscillations in Transistor Oscillators—R. J. Paul. (*Nachr. Tech.*, vol. 8, pp. 109–116; March, 1958.) The condition for self-oscillation is established and oscillators are considered in two groups, with frequency either dependent on or independent of transistor parameters.

621.373.52:538.569.4 1810
Transistorized Nuclear-Resonance Magnetic-Field Probe—J. R. Singer and S. D. Johnson. (*Rev. Sci. Instrum.*, vol. 30, pp. 92–93; February, 1959.) Description of a marginal-oscillator type of nuclear-resonance detector.

621.373.52:621.395.44 1811
Frequency-Stable Transistor Oscillators in Carrier-Frequency Techniques—W. Hüfner. (*Nachr. Tech.*, vol. 8, pp. 117–122; March, 1958.) The design of a Meacham-bridge oscillator is given.

621.374.32:621.387 1812
Reversible Dekatron Counter—W. K. Hsu. (*Wireless World*, vol. 65, pp. 190–192; April, 1959.) A counter circuit is described which has two inputs, one of which allows addition to and the other subtraction from an existing count.

621.374.34 1813
The Static Characteristics of the Cathode—Coupled Limiter (Clipper)—J. Schulz. (*Frequenz*, vol. 12, pp. 114–117; April, 1958.)

621.375.223.029.33:621.397.62 1814
Cathode Compensation—H. D. Kitchin. (*Electronic Radio Eng.*, vol. 36, pp. 122–128; April, 1959.) The design of a cathode-compensated pentode video stage is discussed, particularly the selection of tubes for the cathode resistor and capacitor and the use of a "bleed" resistor.

621.375.226.012.6 1815
Response of Cascaded Double-Tuned Circuits—Y. Peless. (*Electronic Radio Eng.*, vol. 36, pp. 134–140; April, 1959.) The transient and steady-state responses are developed in terms of the location of the poles of the transfer function. The results can also be applied to networks with a response asymmetrical about the band center, but with a narrow relative bandwidth.

621.375.4 1816
Design of Transistor RC Amplifiers—R. P. Murray. (IRE Trans. on Audio, vol. AU-6, pp. 67–76; May/June, 1958. Abstract, Proc. IRE, vol. 46, pp. 1888–1889; November, 1958.)

621.375.4.018.7 1817
Transistor Nonlinearity: Dependence on Emitter Bias Current in *P-N-P* Alloy-Junction Transistors—D. R. Fewer. (IRE Trans. on Audio, vol. AU-6, pp. 41–44; March/April, 1958. Abstract, Proc. IRE, vol. 46, p. 1774; October, 1958.)

621.375.9:538.569.4 1818
Radiation Damping Effects in Two-Level Maser Oscillators—A Yariv, J. R. Singer, and J. Kemp. (*J. Appl. Phys.*, vol. 30, p. 265; February, 1959.) Analytical note on the modulation effects occurring in a spontaneously radiating inverted two-level spin system.

621.375.9:621.3.011.23 1819
Phase-Distortionless Limiting by a Parametric Method—A. E. Siegman. (Proc. IRE, vol. 47, pp. 447–448; March, 1959.) Nearly ideal limiting can be obtained by using the signal to be limited as the "pump"-frequency signal in a parametric device.

621.375.9:621.3.011.4 1820
Nonlinear-Capacitance Amplifiers—L. S. Nergaard. (*RCA Rev.*, vol. 20, pp. 3–17; March, 1959.) An account of the physical principles and the design of variable-capacitance amplifiers. The effective noise temperatures achieved are compared with those of other low-noise amplifiers and of terrestrial and extraterrestrial noise sources.

621.375.9+621.372.632:621.385.029.6 1821
A Three-Frequency Electron-Beam Parametric Amplifier and Frequency Converter—Louisell. (See 2065.)

621.375.9:621.385.029.6 1822
Gain, Bandwidth and Noise in a Cavity-Parametric Amplifier using an Electron Beam—Wade and Heffner. (See 2066.)

621.375.9:621.385.029.6:621.372.2 1823
Travelling-Wave Couplers for Longitudinal Beam-Type Amplifiers—R. W. Gould. (See 2067.)

621.375.9.029.6:537.311.33 1824
The Physical Principles of a Negative-Mass Amplifier—H. Krömer. (Proc. IRE, vol. 47, pp. 397–406; March, 1959.) Negative effective masses for relatively low energies may be obtained if the energy contours are re-entrant near the edge of the frequency band, as is the case for heavy holes in germanium and certain other semiconductors with degenerate band edges. Operation at frequencies up to 1000 kmc (0.3 mm λ) is envisaged. See also 2354 of 1958.

621.376.22.029.64:621.318.134 1825
Microwave Ferrite Modulators for High Signal Frequencies—A. L. Morris. (*J. Brit. IRE*, vol. 19, pp. 117–129; February, 1959.) Methods are suggested for avoiding ferromagnetic resonance. Skin effects are overcome by using the waveguide as a modulating helix. Performance details of two experimental modulators for *X*-band frequencies are given.

621.376.32:538.569.4 1826
Frequency Modulator for a Marginal Oscillator—D. A. Jennings and W. H. Tanttila. (*Rev. Sci. Instrum.*, vol. 30, pp. 137–138; February, 1959.) A voltage-sensitive capacitor with dc bias is used in the tank circuit of the oscillator.

621.376.4:621.314.7:621.398 1827
The Accuracy Obtainable with Transistors in Pulse Amplitude and Pulse Width Modulation—E. Schenck. (*Nachrichtentech. Z.*, vol. 11, pp. 191–196; April, 1958.) The effects of ambient temperature and power supply fluctuations on transistor PM circuits for telemetry applications are analyzed. The design of a PAM/PWM converter is given.

GENERAL PHYSICS

535.215:535.34 1828
Photoconductivity as a Function of Optical Absorption—A. M. Goodman. (*J. Appl. Phys.*, vol. 30, pp. 144–147; February, 1959.) A theoretical analysis of the dependence of photoconductivity on optical absorption, based on the concept of a "schubweg" or mean range for the optically liberated charge carriers of each sign.

537.122 1829
Correlation Function for a System of Particles Carrying Like Charges—V. P. Galaiko and L. E. Pargamanik. (*Dokl. Ak. Nauk. SSSR*, vol. 123, pp. 999–1002; December 21, 1958.) A description of a mathematical method for the determination of correlation functions for systems of particles carrying like charges. This method could be extended to charges of different sign and also to the kinetic theory of charged particles. See also 3825 of 1957 (Tyablikov and Tolmachev).

537.122:537.2 1830
A Dielectric Formulation of the Many-Body Problem: Application to the Free Electron Gas—P. Nozières and D. Pines. (*Nuovo Cim.*, vol. 9, pp. 470–490; August 1, 1958. In English.)

537.226:539.2 1831
Theory of the Contribution of Excitation to the Complex Dielectric Constant of Crystals—J. J. Hopfield. (*Phys. Rev.*, vol. 112, pp. 1555–1567; December 1, 1958.)

537.291 1832
Graphical-Analytical Construction of the Space Trajectory of Charged Particles in

Magnetic Fields—N. I. Shtepa. (*Radiotekh. Elektron.*, vol. 2, pp. 790–795; June, 1957.) Two methods are described for plotting the trajectory of relativistic charged particles: an acceleration method and the "radii of curvature" method (see 2720 of 1957).

537.525 1833
Effect of Space Charge in Cold-Cathode Gas Discharges—A. L. Ward. (*Phys. Rev.*, vol. 112, pp. 1852–1857; December 15, 1958.) Townsend's basic ionization equations for cold-cathode discharges between parallel plates are modified by Poisson's equation to account for space-charge effects. Numerical calculations have been made for argon.

537.525.029.6:551.510.52 1834
High-Frequency Breakdown in Air at High Altitudes—A. D. MacDonald. (PROC. IRE, vol. 47, pp. 436–441; March, 1959.) Breakdown field is computed for 100 mc, 3, 10, 20 and 35 kmc from atmospheric data, the validity of which is discussed. Considerably more power per unit area can be transmitted at the higher frequencies.

537.533 1835
Longitudinal Oscillations of Electron-Ion Beams—P. V. Polovin and N. L. Tsintsadze. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2615–2623; November, 1957.) Investigation of the stability of electron-ion beams leads to a differential equation that is difficult to solve. A qualitative method based on self-conjugate differential operators is used, which avoids solution of this equation.

537.533 1836
Compensation of Space Charge in an Electron Beam—V. I. Volosok and B. V. Chirikov. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2624–2630; November, 1957.) Measurements of the electric field of the space charge in the beam were carried out to evaluate the lifetime of the virtual cathode.

537.533.74 1837
Calculation of the Spatial and Angular Distribution of a Stream of Particles with Multiple Scattering—F. Lenz. (*Z. angew. Phys.*, vol. 10, pp. 31–34; January, 1958.) Scattering of electron beams is considered taking account of absorption and retardation effects.

537.534.8 1838
Energy Spectrum of Secondary Electrons Emitted by a Metal under the Action of a Fast Ion Beam—F. Pradal and R. Simon. (*Compt. rend. Acad. Sci., Paris*, vol. 247, pp. 438–441; July 28, 1958.) Spectra have been analyzed by means of a magnetic spectrograph.

537.54:538.6:523.75 1839
Solar Proton Stream Forms with a Laboratory Model—Bennett. (See 1864.)

537.56:538.56 1840
A Transit-Time Relation for Plasma Electron Oscillations—K. G. Emeleus and D. W. Mahaffey. (*J. Electronics Control*, vol. 5, pp. 559–560; December, 1958.)

537.56:538.566 1841
Microwave Propagation in Hot Magnetoplasmas—J. E. Drummond. (*Phys. Rev.*, vol. 112, pp. 1460–1464; December 1, 1958.) The refractive indexes for circularly polarized waves propagating along the magnetic field in an ionized gas at high temperature are calculated. They are found to depend sensitively on electron density and temperature.

537.56:538.566 1842
Conductivity of Plasmas to Microwaves—E. A. Desloge, S. W. Matthysse, and H. Margenau. (*Phys. Rev.*, vol. 112, pp. 1437–1440; December, 1 1958.) A new derivation of an expression previously obtained for plasma conductivity [see 1702 of 1958 (Margenau)], and an alternative expression which avoids previous difficulties with negative conductivities.

537.56:538.63:621.319.4 1843
Hydromagnetic Capacitor—O. Anderson, W. R. Baker, A. Bratenahl, H. P. Furth, and W. B. Kunkel. (*J. Appl. Phys.*, vol. 30, pp. 188–196; February, 1959.) Very high dielectric constants can be achieved with an ionized gas in a strong magnetic field. When an orthogonal electric field is applied, resultant particle drift stores electrical energy. Dielectric constants from 10^6 to 10^8 have been measured in a coaxial capacitor using a rotating plasma disk. Potential use in fast-discharge work is considered.

538.56:621.372.413:537.122 1844
On the Question of Quantum Effects in the Interaction of Electrons with High-Frequency Fields in Resonators—V. L. Ginzburg and V. M. Fahn. (*Radiotekh. Elektron.*, vol. 2, pp. 780–789; June, 1957.) It is shown that the quantum effect corresponds to the interaction of electrons with a neutral oscillatory field in the resonator. Methods of quantum mechanics are avoided by using the Nyquist formula [see, *e.g.*, *Phys. Rev.*, vol. 101, pp. 1620–1626; March 15, 1956. (Weber)].

538.56.029.66 1845
The Band between Microwave and Infrared Regions—I. Kaufman. (PROC. IRE, vol. 47, pp. 381–396; March, 1959.) Difficulties that have hitherto prevented microwave generation in the 300–3000 kmc region are discussed. Schemes which might overcome these difficulties are considered.

538.566 1846
Multistage Resonance Absorbers for Centimetre Electromagnetic Waves—H. J. Schmitt and W. Futtermenger. (*Z. angew. Phys.*, vol. 10, pp. 1–7; January, 1958.) A three-stage dipole resonance absorber is derived from the two-stage absorber described in 1064 of 1957 (Schmitt) by the addition of a second dipole grid at a distance of $\lambda/8$ in front of the metal surface. The frequency characteristics of the reflection coefficient are calculated and compared with the results of measurements.

538.566 1847
The Absorption of Centimetre Electromagnetic Waves in Artificially Anisotropic Media—R. Pottel. (*Z. angew. Phys.*, vol. 10, pp. 8–16; January, 1958.) Two methods of equalizing the complex permeability and dielectric coefficient of a medium to obtain absorption at cm λ are discussed. In one case the medium consists of thin parallel layers, in the other use is made of gyromagnetic resonance in ferrite material subjected to a static magnetic field. Experimental results are given.

538.566 1848
The Group Velocity of Damped Waves—L. A. Vainshtein. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2606–2614; November, 1957.) The group velocity is shown to be equal to $S_z/\bar{\omega}$ where S_z is the component of the Poynting vector along the direction of propagation and $\bar{\omega}$ the energy density. The electromagnetic energy and group velocities are calculated for a simple medium, a plasma and a dielectric with bound electrons.

538.566 1849
Kinematics of Growing Waves—P. A. Sturrock. (*Phys. Rev.*, vol. 112, pp. 1448–1503; December 1, 1958.) Analytical treatment of the problem of distinguishing between amplifying and evanescent waves.

538.566:535.42]+534.26 1850
Asymptotic Development of Double Integrals Encountered in Diffraction Theory—N. Chako. (*Compt. rend. Acad. Sci., Paris*, vol. 247, pp. 436–438; July 28, 1958.)

538.566:535.42 1851
A New Method for the Solution of a Problem of Diffraction of Electromagnetic Plane Waves at an Unlimited Rectilinear Slit, and Related Problems—G. A. Grinberg. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2595–2605; November, 1957.) Using the method described, a good approximation to the exact solution of the diffraction problem can be obtained for normally incident plane waves of wavelength much smaller than or equal to the width of the slit.

538.569.4 1852
Radio-Frequency Spectra of Hydrogen Deuteride in Strong Magnetic Fields—W. E. Quinn, J. M. Baker, J. T. LaTourrette, and N. F. Ramsey. (*Phys. Rev.*, vol. 112, pp. 1929–1940; December 15, 1958.)

538.569.4 1853
Steady-State Free Precession in Nuclear Magnetic Resonance—N. Y. Carr. (*Phys. Rev.*, vol. 112, pp. 1693–1701; December 1, 1958.) Description of a new technique for observing nuclear magnetic resonance.

538.569.4:535.33 1854
An *X*-Band Spectrometer for the Demonstration of Paramagnetic Resonances—W. Stieler. (*Z. angew. Phys.*, vol. 10, pp. 89–95; February, 1958.) A bridge-type spectrometer with frequency stabilization is described.

538.569.4:537.311.62 1855
Theory of Cyclotron Resonance in Metals—Rodriguez. (*Phys. Rev.*, vol. 112, pp. 1616–1620; December 1, 1958.) Analysis of the low-temperature case in which the mean free path of the electrons is greater than the skin depth.

538.569.4:538.222 1856
Multiple Quantum Transitions in Paramagnetic Resonance—P. P. Sorokin, I. L. Gelles, and W. V. Smith. (*Phys. Rev.*, vol. 112, pp. 1513–1515; December 1, 1958.) At high RF field strengths the normal paramagnetic resonance spectrum of Mn^{++} in cubic MgO is found to be modified by additional absorption lines, which are interpreted as double quantum transitions ($\Delta M=2$) occurring between nearly equally spaced energy levels at high RF power. Calculation of the relative absorption intensity for such transitions agrees with experimental results.

538.569.4:621.375.9:535.61—1/2 1857
Infrared and Optical Masers—A. L. Schwawlow and C. H. Townes. (*Phys. Rev.*, vol. 112, pp. 1940–1949; December 15, 1958.) Theoretical aspects of maser-like devices for wavelengths much shorter than 1 cm are discussed. The short-wavelength limit for practical devices is examined. Design principles as illustrated by a system for the infrared region using potassium vapor.

539.2:537.311.31 1858
The Effective Radius of the Electron in Crystal Lattices—A. F. Kapustinskiĭ. (*Dokl. Ak. Nauk, SSSR*, vol. 124, pp. 1265–1266; February 21, 1959.) The effective radius of electrons in metals determined by different methods is found to be 0.78 ± 0.02 Å.

539.2:537.311.31 1859
Relation between the Constants Characterizing the Interaction of Electrons with Phonons and Impurities, in Metals—V. L. Bonch-Bruevich. (*Dokl. Ak. Nauk, SSSR*, vol. 124, pp. 1233–1235; February 21, 1959.) A mathematical analysis.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.165 1860
Unusual Cosmic-Ray Intensity Fluctuations Observed at Southern Stations during October 21–24, 1957—K. G. McCracken and N. R. Parsons. (*Phys. Rev.*, vol. 112, pp. 1798–1801; December 1, 1958.) The fluctuations observed have unusual features which suggest the existence of a short-lived and highly directional anisotropy of the primary radiation during the period immediately preceding a Forbush-type decrease.

523.165:061.3 1861
International Convention on Cosmic Rays—(*Nuovo Cim.*, vol. 8, suppl. no. 2, pp. 125–804; 1958. In English.) The text is given of over 90 papers presented at a convention held at Varenna, June 21–26, 1957. The subject matter is divided into the following groups: *a*) solar and geomagnetic effects on cosmic rays, *b*) problems of origin, *c*) composition of primary radiation, *d*) air showers, and *e*) interactions of cosmic radiation.

523.165:523.75 1862
Latitude Variation of 27-Day Cosmic-Ray Intensity Decreases—R. R. Brown. (*Nuovo Cim.*, vol. 9, pp. 197–207; July 16, 1958. In English.) The latitude variation of decreases in cosmic-ray intensity due to the modulation effects of a geocentric nebula of disordered magnetic fields has been measured. Experimental data provided by neutron monitor observations during a period of intense solar activity show that equatorial variations exceed calculated values by a factor of two or more.

523.3:621.396.96 1863
Radio Observations of the Lunar Surface—J. K. Hargreaves. (*Proc. Phys. Soc.*, vol. 73, pp. 536–537; March 1, 1959.) Information about the moon's surface can be obtained from a statistical consideration of radar scattering from the surface.

523.75:537.54:538.6 1864
Solar Proton Stream Forms with a Laboratory Model—W. H. Bennett. (*Rev. Sci. Instrum.*, vol. 30, pp. 63–69; February, 1959.) A "Störmertron" tube is described for simulating the streams of charged particles in the earth's dipole magnetic field. Photographs of stream forms and contacts are presented. The same techniques can be used to study stream forms in other complicated magnetic fields.

550.385:523.75 1865
Recurrent Geomagnetic Storms and Solar Prominences—R. T. Hansen. (*J. Geophys. Res.*, vol. 64, pp. 23–35; January, 1959.) An examination of the association of prominence areas with days of recurrent storms during the period 1917–1944. The identification of *M* regions with solar prominences is not confirmed.

550.389.2:629.19 1866
Motion of a Satellite around an Unsymmetrical Central Body—T. E. Sterne. (*J. Appl. Phys.*, vol. 30, p. 270; February, 1959.) Comment on 1541 of May (Newton).

550.389.2:629.19 1867
A Determination of the Coefficient *J* of the Second Harmonic in the Earth's Gravitational Potential from the Orbit of Satellite 1958 β2—M. Lecar, J. Sorenson, and A. Eckels. (*J. Geophys. Res.*, vol. 64, pp. 209–216; February, 1959.)

550.389.2:629.19 1868
Vanguard I I.G.Y. Satellite (1958 Beta)—R. L. Easton, and M. J. Votaw. (*Rev. Sci. Instrum.*, vol. 30, pp. 70–75; February, 1959.) The instrumentation used, the measurements made, and the uses of the satellite are described. Information on satellite temperatures and rotation rate is given. Orbital data are being used for measuring the oblateness of the earth and for correcting mapping errors.

550.389.2:629.19 1869
Information by Radio from the Satellites—J. A. Ratcliffe. (*J. IEE*, vol. 4, pp. 603–608; November, 1958.) An account of some early results obtained mainly from observations of satellite 1957α.

550.389.2:629.19:061.3 1870
I.G.Y. Conference in Moscow: Soviet Papers Presented at the Rocket and Satellite Symposium—J. W. Townsend, Jr. (*Science*, vol. 129, pp. 80–84; January 9, 1959.) Summary of preliminary data presented at the 5th General Assembly of C.S.A.G.I., Moscow, July 30–August 9, 1958.

550.389.2:629.19:551.510.53 1871
Densities and Temperatures of the Upper Atmosphere Inferred from Satellite Observations—G. F. Schilling and T. E. Sterne. (*J. Geophys. Res.*, vol. 64, pp. 1–4; January, 1959.) The atmospheric density between 180 and 400 km altitude appears to be appreciably higher than that derived from rocket data. The temperature in this region must therefore be higher than that given by present model atmospheres. The density is still below that derived from meteor data.

550.389.2:629.19:551.510.53 1872
Investigation of the Upper Atmosphere by means of the Third Artificial Earth Satellite—V. I. Krasovskiĭ. (*Priroda Mosk.*, pp. 71–78; December, 1958.) A brief description of geophysical investigations carried out in U.S.S.R. The temperature variation and the distribution of the electron density with height up to about 500 km are discussed, and the collision of an artificial satellite with micrometeorites is considered.

550.389.2:629.19:551.510.535 1873
The Ionosphere and Artificial Earth Satellites—Ya. L. Al'pert. (*Priroda, Mosk.*, pp. 71–77; October, 1958.) Two methods for investigating the ionosphere are briefly described; the Doppler shift of emitted signals and the observation of the times of "radio rise" and "radio setting" of the satellites. From the difference between the optical and the radio observations of the rising and setting, the electron concentration can be calculated.

550.389.2:629.19:551.510.535 1874
Satellite Doppler Measurements and the Ionosphere—J. A. Thomas and F. H. Hibberd. (*J. Atmos. Terrest. Phys.*, vol. 13, pp. 376–379; February, 1959.) Simultaneous Doppler measurements of satellite 1957α at frequencies of 20 and 40 mc were shown to be in agreement with an approximate model of the ionosphere for the same period.

550.389.2:629.19:621.396.43 1875
Transoceanic Communication by Means of Satellites—Pierce and Kompfner. (See 2012.)

551.510.53:551.524.7 1876
Temperatures in the High Atmosphere—F. S. Johnson. (*Ann. géophys.*, vol. 14, pp. 94–108; January/March, 1958. In English.) Temperature distribution is discussed in relation to rocket and radio observations on the assumption that heating of the ionosphere is due primarily to solar radiation and that temperature distribution is controlled by thermal conduction. A model is presented which indicates a very rapid temperature rise between 100 and 200 km. This would produce a density distribution in agreement with observations.

551.510.53:551.593 1877
The Temperature in the Atmospheric Region Emitting the Nightglow OI 5577 Line and in Regions above Faint Auroral Arcs—E. B. Armstrong. (*J. Atmos. Terrest. Phys.*, vol. 13, pp. 205–216; February, 1959.)

551.510.535 1878
Effect of Vertical Drifts on the Nocturnal Ionization of the Lower Ionosphere—M. N. Rao and A. P. Mitra. (*J. Atmos. Terrest. Phys.*, vol. 13, pp. 271–290; February, 1959.) Vertical ionic drifts present in quiet conditions and enhanced during a magnetic storm result in a vertical redistribution of ionization and thus change the apparent recombination coefficient at a given height. The resulting $N(h)$ profiles are calculated and shown to explain the sudden cessation of night-time echoes of LF radio waves at times of magnetic disturbance reported by Lindquist (724 of 1954).

551.510.535 1879
A Survey of the Present Knowledge of Sporadic-*E* Ionization—J. A. Thomas and E. K. Smith. (*J. Atmos. Terrest. Phys.*, vol. 13, pp. 295–314; February, 1959.) Discussion of the various techniques of observing E_s, and the classification of the phenomenon by geographical zones (auroral, temperate and equatorial), and by type, as evidenced by the trace recorded on an ionogram. The diurnal and seasonal variations in the three zones are shown to be quite distinct. Various theories of the nature of the ionization formations and the agencies producing them are considered, and the subjects requiring special experimental and theoretical attention are indicated Over 100 references

551.510.535 1880
Lunar Tides in the Sporadic *E*-Layer at Ibadan—R. W. Wright and N. J. Skinner. (*J. Atmos. Terrest. Phys.*, vol. 13, pp. 217–221; February, 1959.) "An analysis of the lunar semi-diurnal tides in $f'E_s$ and $h'E_s$ is made for Ibadan. The results are presented in the form of harmonic dials. Comparisons are made between these results and those of other stations."

551.510.535 1881
Ionospheric Measurements made at Halley Bay—W. H. Bellchambers and W. R. Piggett. (*Nature, London*, vol. 182, pp. 1596–1597; December 6, 1958.) Graphs of the monthly median values of f_0F_2 at Halley Bay (75°31'S, 26°36'W) for July (winter), September (equinox), and December (summer) are given. A large diurnal variation of electron density occurs in winter. The seasonal maximum of electron density during the day is found at the equinoxes.

551.510.535 1882
The Diurnal and Annual Variations of f_0F_2 over the Polar Regions—S. C. Coroniti and R. Penndorf. (*J. Geophys. Res.*, vol. 64, pp. 5–18; January, 1959.) The variations have been studied for about 15 stations over a three year period and the following results have been found: *a*) the diurnal variation is largest in winter and smallest in summer; the maximum values of f_0F_2 occur at noon in the northern and at midnight in the southern hemisphere; *b*) there is a regular annual variation for a given time of day; f_0F_2 in northern winter has deep minima in the morning and evening; *c*) in summer the latitude differences in f_0F_2 are small but become large in winter, up to 0.85 mc per degree; *d*) the lines of equal critical frequency lie between circles of geographic and geomagnetic latitude.

551.510.535:523.78 1883
Ionospheric Changes at Singapore during the Solar Eclipse of 14th December 1955—C. M. Minnis. (*J. Atmos. Terrest. Phys.*, vol. 13, pp. 346–350; February, 1959.) The eclipse oc-

curred in the late afternoon and no simple interpretation of the changes in the E and F_1 layers can be given. The critical frequency of the F_2 layer did not change but there were considerable eclipse effects in the lower part of the F layer.

551.510.535:621.396.11 1884
An Analysis of Drifts of the Signal Pattern associated with Ionospheric Reflections—D. G. Yerg. (*J. Geophys. Res.*, vol. 64, pp. 27–31; January, 1959.) A statistical treatment to determine the drift and random velocity components from the signals observed on three spaced receivers. Particular attention is given to the random motion in order to examine the detailed movement of the pattern. The results suggest a ruled pattern undergoing fluctuations in contour spacing as the pattern drifts across the receiver site.

551.510.535:621.396.11 1885
Triple Splitting of the *F* Echoes—R. Satyanarayana, K. Bakhru, and S. R. Khastgir.—(*J. Atmos. Terrest. Phys.*, vol. 13, pp. 201–204; February, 1959.) Polarization measurements of triple-split echoes at Banaras (25°N, 83°E) agree with measurements at high latitudes. The process concerned in producing triple-splitting at low latitudes is discussed.

551.510.535(98) 1886
Arctic Measurements of Electron Collision Frequencies in the *D* Region of the Ionosphere—J. A. Kane. (*J. Geophys. Res.*, vol. 64, pp. 133–139; February, 1959.) Simultaneous measurement of refractive indexes and the difference in the absorption of the two magneto-ionic components of a 7.75-mc CW signal transmitted from a rocket, allowed the electron collision frequency profile to be determined. The results of two midday flights indicate that the frequencies are lower by a factor of three than the unpublished theoretical values calculated by Nicolet.

551.594.5 1887
Electric Field Theory of Aurorae—G. C. Reid. (*Nature, London*, vol. 182, pp. 1791–1792; December 27, 1958.) A qualitative description of the development of a typical auroral display following the growth of an electric field in the ionosphere.

551.594.5 1888
An Artificial Aurora—P. H. Fowler and C. J. Waddington. (*Nature, London*, vol. 182, p. 1728; December 20, 1958.) The auroral display at Apia reported by Cullington (1569 of May) could not have been caused by direct radiation from the explosion of a nuclear bomb at Johnston Island (17°N, 169°W), which is approximately 2200 miles away, but was probably due to charged particles from the explosion.

551.594.5:621.396.11.029.62:551.510.535 1889
A Bistatic Radio Investigation of Auroral Ionization—Collins and Forsyth. (See 2001.)

LOCATION AND AIDS TO NAVIGATION

621.396.933 1890
B.O.A.C.'s Comet: Communications and Navaids—(*Brit. Commun. Electronics*, vol. 6, pp. 34–35; January, 1959.) A general report of the radio equipment used.

621.396.933.1 1891
A Low-Drain Distress Beacon for a Crash Position Indicator—D. M. Makow, H. R. Smyth, S. K. Keays, and R. R. Peal. (*J. Brit. IRE*, vol. 19, pp. 135–147; March, 1959.) A pulsed transmitter with trickle-charged batteries and an internal capacitor antenna operates at 243 mc for approximately 100 hours. Cased in shock-absorbing foam, the device forms part of an aerofoil designed to fall at a safe speed and land clear of wreckage when automatically released from a crashing aircraft.

621.396.933.23 1892
B.L.E.U.'s Automatic Landing System—(*Brit. Commun. Electronics*, vol. 5, p. 927; December, 1958.) A system developed by the Blind-Landing Experimental Unit of the Royal Aircraft Establishment. See 464 of February.

621.396.962.25/.3:621.396.969.11 1893
A Comparison between Pulse and Frequency Modulation Echo-Ranging Systems—L. Kay. (*J. Brit. IRE*, vol. 19, pp. 105–113; February, 1959.) The FM system is more flexible in its design parameters than the pulse system. Where the echo/background ratio is not important the FM system can provide a higher information rate than the pulse system.

621.396.962.3:621.396.665 1894
The Effects of Automatic Gain Control Performance on the Tracking Accuracy of Monopulse Radar Systems—J. H. Dunn and D. D. Howard. (Proc. IRE, vol. 47, pp. 430–435; March, 1959.) An analysis including practical data shows that a short-time-constant fast-acting AGC will minimize tracking noise. Servo bandwidth should be as small as possible.

621.396.962.38 1895
A Practical Application of Phase-Measuring Techniques to Precision Angle and Distance Measurements—W. J. Thompson. (IRE Trans. on Instrumentation, vol. I-6, pp. 12–17; March, 1957.) Proc. IRE, vol. 45, p. 898; June, 1957.)

621.396.965:621.318.134 1896
Volumetric Scanning of a Radar with Ferrite Phase Shifters—F. E. Goodwin and H. R. Serf. (Proc. IRE, vol. 47, pp. 453–454; March, 1959.)

621.396.969.11 1897
VOR—Compatible Doppler Omnirange, Design Considerations—P. G. Hansel. (Proc. IRE, vol. 47, pp. 443–444; March, 1959.)

621.396.969.14:656.1 1898
The Telefunken Traffic Radar—H. Lueg, W. Schallehn, and H. Toedter. (*Elektrotech. Z., Ed B*, vol. 10, pp. 385–390; October 21, 1958.) Description of the design and operation of Doppler radar equipment for the measurement of the speed of road vehicles.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.5 1899
Theory and Design of Getter-Ion Pumps—L. Holland. (*J. Sci. Instr.*, vol. 36, pp. 105–116; March, 1959.) Includes a survey of the sorption and deposition properties of a range of getter materials. 63 references.

535.215 1900
Gain-Bandwidth Product for Photoconductors—A. Rose and M. A. Lampert. (*RCA Rev.*, vol. 20, pp. 57–68; March, 1959.) This product is shown to be proportional to an enhancement factor M dependent on the trap distribution. For many materials the maximum value of M is unity; much greater values may, however, be obtained under certain circumstances. See also 1901.

535.215 1901
Properties of Deep Traps derived from Space-Charge-Current Flow and Photoconductive Decay—R. W. Smith. (*RCA Rev.*, vol. 20, pp. 69–78; March, 1959.) The gain-bandwidth product is evaluated from measurements on a CdS single crystal. By using different light levels and a range of voltages, the Fermi level is moved to scan the deep-lying states.

535.215 1902
Gains, Response Times, and Trap Distributions in Powder Photoconductors—H. B. DeVore. (*RCA Rev.*, vol. 20, pp. 79–91; March, 1959.) Measurements have been made on CdS and CdSe powders and the data compared with the analytic expression given by Rose and Lampert (1900) relating the gain-bandwidth product to the dielectric relaxation time.

535.215:546.47-31 1903
The Field Effect in Insulating ZnO Powder—W. Ruppel. (*Z. Phys.*, vol. 152, pp. 235–241; July 28, 1958.) Lifetime and trap concentration are calculated from the results of field-effect measurements on photoconductive ZnO powder. See also 877 of March (Ruppel *et al.*).

535.215:546.482.21 1904
The Spectral Distribution of Photoconductivity in CdS Single Crystals—K. W. Böer and H. Gutjohr. (*Z. Phys.*, vol. 152, pp. 203–213; July 28, 1958.) The distribution of photoconductivity is investigated in relation to crystal structure and as a function of applied voltage.

535.215:546.482.21:548.5 1905
Vaporization-Crystallization Method for Growing CdS Single Crystals—D. R. Boyd and Y. T. Sihvonen. (*J. Appl. Phys.*, vol. 30, pp. 176–179; February, 1959.)

535.215:546.482.31 1906
Characteristics of the Increased Conductivity of Cadmium Selenide Single Crystals under X-Ray Excitation: Parts 1 and 2—S. V. Svechnikov. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2492–2506; November, 1957.)

535.215:546.482.41:539.23 1907
High-Voltage Photovoltaic Effect—B. Goldstein and L. Pensak. (*J. Appl. Phys.*, vol. 30, pp. 155–161; February, 1959.) See 1756 and 1757 of 1958.

535.37:546.472.21 1908
On Certain Chromatic Aspects of the Photoluminescence of a ZnS-Cu Phosphor—J. P. Leroux and P. Thureau. *Comp. rend. Acad. Sci., Paris*, vol. 247, pp. 924–926; September 29, 1958.) The color of the light emitted by a ZnS-Cu phosphor depends on the intensity of excitation and on the surface density of the irradiated powder.

535.376 1909
Electroluminescence—V. E. Oranovskiĭ. (*Priroda, Mosk.*, vol. 11, pp. 17–22; November, 1958.) Description of the process of luminescence of substances such as Zn_2SiO_4, BN, $BaTiO_3$, $SrTiO_3$, and TiO_2, and particularly ZnS and ZnSe and their application to television screens.

535.376:546.281.26 1910
Impurity Bands and Electroluminescence in SiC *P-N* Junctions—L. Patrick and W. J. Choyke. (*J. Appl. Phys.*, vol. 30, pp. 236–248; February, 1954.) A study of the electroluminescence of certain SiC *p-n* junctions, between 77°K and 830°K, and over a range of 10^4 in current density, has been used to verify and extend a three-part model of the junctions derived from electrical measurements.

537.227 1911
Anomalous Polarization in Ferroelectrics and Other Oxides—J. D. Hurd, A. W. Simpson, and R. H. Tredgold. (*Proc. Phys. Soc.*, vol. 73, pp. 448–454; March 1, 1959.) It is shown that a number of metal oxides have a large relaxation polarization which cannot be explained in terms of ferroelectricity or of the Maxwell-Wagner effect. Experimental results indicate

that the materials behave as solid-state secondary cells.

537.227 1912
Nonferroelectric Phase Transitions in Solid Solutions Formed in (Ca, Sr) (Ti, Zr)O_3 and Na(Nb, Ta)O_3 Systems—G. A. Smolenskiĭ, V. A. Isupov, A. I. Agranovskaya, and E. D. Sholokhova. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2528–2534; November, 1957.) An investigation of the temperature dependence of permittivity.

537.227:546.431.824–31 1913
Effect of γ-Ray and Pile Irradiation on the Coercive Field of $BaTiO_3$—I. Lefkowitz and T. Mitsui. (*J. Appl. Phys.*, vol. 30, p. 269; February, 1959.)

537.311.31 1914
Solution of Bloch's Integral Equation for Metal Electrons in an Electric Field for the Whole Temperature Range—D. Langbein. (*Z. Phys.*, vol. 152, pp. 123–142; July 11, 1958.)

537.311.33 1915
Decay of Excess Carriers in Semiconductors—K. C. Nomura and J. S. Blakemore. (*Phys. Rev.*, vol. 112, pp. 1607–1615; December 1, 1958.) "A discussion is given of the nonlinear differential equations which govern the decay of excess carriers with arbitrary densities. The form of decay is explored for situations where the Fermi level is in the same half of the energy gap as the recombination level; criteria are established for both strong and weak trapping in addition to recombinative action. Analytic results are augmented and illustrated by numerically computed decay curves for a variety of circumstances. The separate solutions for holes and electrons are combined to show various kinds of behavior for photoconductive lifetime."

537.311.33 1916
Effects of Carrier Injection on the Recombination Velocity in Semiconductor Surfaces—G. C. Dousmanis. (*J. Appl. Phys.*, vol. 30, pp. 180–184; February, 1959.) Predictions of theory based on the Shockley-Read model are illustrated with curves of surface recombination velocity as a function of the the fractional excess carrier density where, for Ge, experimentally determined surface parameters are used.

537.311.33 1917
Narrowing the Energy Gap in Semiconductors by Compensation—F. Stern and J. R. Dixon. (*J. Appl. Phys.*, vol. 30, pp. 268–269; February, 1959.) The addition of large and equal numbers of both donor and acceptor impurity atoms has the effect of lowering the bottom of the conduction band and raising the top of the valence band, thus narrowing the energy gap while maintaining a low carrier concentration. This has been verified by doping InAs with S and Zn.

537.311.33 1918
Change of Semiconductor Properties with Fusion—A. I. Gubanov. (*Zh. Tekh. Fiz.*, pp. 2510–2516; November, 1957.) The variations in width of the forbidden zone and effective masses of the charge carriers on fusion are considered in relation to the variations produced by deformation of the crystal.

537.311.33 1919
Distribution of Non-equilibrium Charge Carriers in the Base Region of a *P-N* Junction with a High Injection Coefficient—M. I. Iglitsyn, Yu. A. Kontsevoĭ, and A. I. Sidorov. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2458–2460; November, 1957.) Stationary conditions of *p-n* junctions in semiconductors with an arbitrary injection level and lifetime depending upon this level are expressed by a system of equations, and the distribution of minority-carrier concentration in the base region is shown graphically.

537.311.33 1920
Vitreous Semiconductors—T. N. Vengel' and B. T. Kolomiets. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2484–2491; November, 1947.) Eleven compounds in the system As_2Se_3—As_2Te_3 have been investigated, and their photoelectric and thermoelectric properties are examined in relation to their chemical composition. See also 2796 of 1957 and 1766 *n* of 1958 (Gory-unova and Kolomiets).

537.311.33:535.215 1921
Measurement of Lifetime by the Photoconductive Decay Method—B. K. Ridley. (*J. Electronics Control*, vol. 5, pp. 549–558; December, 1958.) General equations have been derived considering both steady-state and pulsed initial distributions. The decay is exponential only for zero surface recombination. With high surface recombination velocities the loss of exponential nature of the decay leads to an error in the measured value of filament lifetime. This error has been found numerically and graphically for various lifetimes and penetration depths.

537.311.33:537.533 1922
On the Electrostatic Electron Emission of Semiconductors—I. I. Gofman, B. G. Smirnov, G. S. Spirin, and G. N. Shuppe. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2662–2663; November, 1957.) Note of an experimental investigation of W_2C showing that the field-emission current/voltage characteristic is in qualitative agreement with theory [see 717 of 1956 (Stratton)].

537.311.33:538.63 1923
The Theory of Electrical and Photoelectric Effects for Three Carriers in a Magnetic Field—A. K. Walton and T. S. Moss. (*Proc. Phys. Soc.*, vol. 73, pp. 399–412; March 1, 1959.) "Formulas for the Hall, magnetoresistance, photoelectromagnetic, and Dember effects due to electrons and slow and fast holes are derived for the cases of energy-independent relaxation time and lattice scattering. The results are discussed for the particular case of germanium."

537.311.33:546.26–1:537.32 1924
The Thermoelectric Power of a Semiconducting Diamond—H. J. Goldsmid, C. C. Jenns, and D. A. Wright. (*Proc. Phys. Soc.*, vol. 73, pp. 393–398; March 1, 1959.) The thermoelectric power of *p*-type semiconducting diamond was measured between 220°K and 700°K. The phonon-drag component had a value of about 2.5 mv/°K at room temperature and varied with temperature approximately as $^{-3.6}$.

537.311.33:546.28 1925
Diffusion of Impurities into Evaporating Silicon—R. L. Batdorf and F. M. Smits. (*J. Appl. Phys.*, vol. 30, pp. 259–264; February, 1959.) A diffusion technique in a vacuum system where the only ambient is the vapor of a diffusing impurity is described and results of measurements are quoted.

537.311.33:546.28 1926
Mechanism of the Formation of Donor States in Heat-Treated Silicon—W. Kaiser, H. L. Frisch, and H. Reiss. (*Phys. Rev.*, vol. 112, pp. 1546–1554; December 1, 1958.) A mechanism is proposed which accounts quantitatively for existing kinetic and extra-kinetic data on the system. It depends on reactions with atomically dissolved oxygen introduced during the process of crystal growing.

537.311.33:546.28 1927
A Comparison of the Theory of Impact Ionization with Measurements on Silicon *P-N* Junctions—F. W. G. Rose. (*J. Electronics Control*, vol. 6, pp. 70–73; January, 1959.) There is a voltage range over which the inverse current rises linearly with voltage on a log-log graph. It is shown that over this range the current is mainly due to impact ionization.

537.311.33:546.289 1928
Investigation of the Field Effect and Surface Recombination in Germanium Samples—A. V. Rzhanov, Yu. F. Novototskii-Vlasov, and I. G. Neizvestnyi. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2440–2450; November, 1957.) The investigation shows that the action of ozone on Ge gives rise to "fast" surface states some of which are recombination states. The density and energy positions of all states introduced by ozone are estimated. The cross section of capture of an electron by recombination states is also evaluated. Preliminary data are derived concerning the dependence of surface recombination velocity on the electrostatic surface potential, from which the ratio of the effective cross sections of capture of an electron and a hole are obtained.

537.311.33:546.289 1929
The Lifetime of Non-equilibrium Charge Carriers in Germanium with Arbitrary Injection Levels—M. I. Iglitsyn, Yu. A. Kontsevoĭ, and A. I. Sidorov. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2461–2468; November, 1957.) An investigation of the dependence of lifetime on the concentration of nonequilibrium charge carriers at different temperatures for Ge specimens alloyed with Sb. At room temperature, the lifetime decreases with increasing injection level for samples of high resistance and increases for samples of low resistance. The type of level (donor or acceptor), their position in the forbidden band, the energy of ionization, and the ratio of probabilities of recapture of electrons and holes can be determined.

537.311.33:546.289 1930
Measurement of Short Lifetimes of Charge Carriers in Germanium—L. S. Smirnov. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2469–2471; November, 1957.) Measurements were carried out on Ge plates with a large-area *p-n* junction. The non-equilibrium carriers were excited by a monochromatic light source near the surface of the semiconductor, and the lifetime of the charge carriers, in the range $2\times10^{-6}-10^{-8}$s, was determined from the short-circuit current and the number of light quanta absorbed. The temperature dependence of lifetime and the position of recombination levels were also evaluated.

537.311.33:546.289 1931
The Influence of an Intense Electric Field on Germanium-Diode Transparency—Yu. I. Ukhanov and S. G. Shul'man. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2507–2509; November, 1957.) Measurements were made on a Ge *p-n* junction with an infrared beam directed perpendicularly to the applied electric field using 1–3 μs pulses to give current densities up to 20 ma/mm². The infrared transparency varied proportionally with the reverse current. Cooling the specimen to 78°K had no effect on the transparency. With a forward current a decrease in transparency was observed.

537.311.33:546.289 1932
Dependence of the Lifetime of Injected Current Carriers on the Concentration of Antimony Impurity in Germanium—V. E. Lashkarev, V. G. Litovchenko, N. M. Omel'-yanovskaya, P. N. Boudarenko, and V. I. Strikha. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2437–2439; November, 1957.) Report of an investigation carried out close to the limit of solubility of Sb in Ge. Results indicate that the Sb impurity atoms are not directly effective as re-

combination centers but that the recombination is due to deeply embedded and uncontrolled impurities originally in the Ge or introduced with the Sb.

537.311.33:546.289 1933
Investigation of the Recombination of Current Carriers in Germanium with Iron Impurity K. D. Glinchuk, E. G. Miselyuk, and N. N. Fortunatova. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2451–2457; November, 1957.) The acceptor level situated at 0.27 ev from the conduction band in *n*-type Ge can be removed by annealing at 450°–500°C. This results in a sharp increase in the lifetime of nonequilibrium charge carriers. It may be explained by the deactivation of iron atoms following their expulsion from Ge lattices.

537.311.33:546.289 1934
The Effect of Annealing on Local Levels and the Lifetime of Non-equilibrium Current Carriers in Germanium with Iron Impurity—K. D. Glinchuk, E. G. Miselyuk, and N. N. Fortunatova. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2666–2667; November, 1957.)

537.311.33:546.289 1935
Orientation Control for Germanium Wafers—B. J. Coughlin, G. L. Davis, and R. L. Kingsnorth. (*J. Sci. Instr.*, vol. 36, pp. 144–145; March, 1959.) A method for the accurate mounting of ingots for cutting and for determining the orientation of small wafers.

537.311.33:546.289 1936
Precision Measurement of the Lattice Constant of Germanium Single Crystals by the Method of Kossel and van Bergen—G. Mack. (*Z. Phys.*, vol. 152, pp. 19–25; July 11, 1958.) See also 1937.

537.311.33:546.289 1937
Precision *X*-Ray Investigations on Germanium-Indium *P-N* Alloy Junctions—G. Mack. (*Z. Phys.*, vol. 152, pp. 26–33; July 11, 1958.) Report of measurements made by the method of Kossel and van Bergen.

537.311.33:546.289 1938
Light-Induced Plasticity in Germanium—G. C. Kuczynski and R. N. Hochman. (*J. Appl. Phys.*, vol. 30, p. 267; February, 1959.)

537.311.33:546.681.19 1939
Piezoresistance in *N*-Type GaAs—A. Sagar. (*Phys. Rev.*, vol. 112, p. 1533; December 1, 1958.) "Piezoresistance and elastoresistance coefficients of *n*-type GaAs were determined at room temperature. The results are consistent with a spherical energy-band model as predicted by Callaway [1816 of 1957] from theory."

537.311.33:546.682.86 1940
Band Structure of InSb—T. Igo, E. Yamaka, and M. Yatani. (*Rep. elect. Commun. Lab., Japan*, vol. 6, pp. 205–210; June, 1958.) A short discussion on the calculation of the band structure.

537.311.33:546.817.241 1941
Investigation of the Thermoelectric Properties of Lead Telluride—E. Z. Gershtein, T. S. Stavitskaya, and L. S. Stil'bans. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2472–2483; November, 1957.) An investigation of the properties of degenerate and nondegenerate samples shows the dependence of the length of the free path of electrons on temperature and energy. The possibility of a correlation between mobility and the temperature dependence of the width of the forbidden zone is also examined.

537.311.33:546.873.241 1942
The Electrical Conductivity and Hall Coefficient of Bismuth Telluride—B. Yates. (*J. Electronics Control*, vol. 6, pp. 26–38; January, 1959.) These quantities have been examined, for a wide range of doping concentrations, over the temperature range 1.3°–660°K. The results cannot be explained by simple temperature-dependent scattering in the conduction or the valence band, nor are they in complete agreement with a model based on two scattering mechanisms operating in one band. A qualitative explanation of the results for *n*-type specimens, over a restricted temperature range, is possible on the basis of an impurity-band model.

537.311.33:621.923.7 1943
Improved Machine for Lapping Very Thin Slices of Semiconductor Materials—D. Baker. (*J. Sci. Instrum.*, vol. 36, pp. 145–147; March, 1959.)

537.312.8:539.234 1944
Magnetic Resistance Variation in Vapour-Deposited Nickel Films as a Function of Film Thickness and Structure—W. Hellenthal. (*Z. Phys.*, vol. 151, pp. 421–430; June 2, 1958.) A quantitative link with conditions observed in solid material is obtained if account is taken of the increase in resistivity as a function of film thickness and structure, and the decrease of spontaneous magnetization.

537.581:546.77:538.63 1945
Effect of a Magnetic Field on Thermionic Emission from Molybdenum—J. Greenburg. (*Phys. Rev.*, vol. 112, pp. 1898–1900; December 15, 1958.) Experiment shows that an applied field of up to 6000 G has no effect on the saturation current density.

538.22 1946
Antiferromagnetism of $CuF_2 \cdot 2H_2O$—G. S. Verma and K. Tokunaga. (*Phys. Rev.*, vol. 112, pp. 1521–1522; December 1, 1958.) "The perpendicular and parallel magnetic susceptibilities have been calculated for $CuF_2 \cdot 2H_2O$ on the basis of Nakamura's theory. The computed values of molar susceptibility for the same compound have been compared with the recent measurements of Bozorth and Nielsen (3163 of 1958) and are found to be in good agreement."

538.22:538.569.4 1947
Evidence for Antiferromagnetism in $Cu_3(CO_3)_2(OH)_2$—R. D. Spence and R. D. Ewing. (*Phys. Rev.*, vol. 112, pp. 1544–1545; December 1, 1958.) Proton resonance measurements at low temperatures are reported. At 1.86°K a transition takes place and the resonance pattern below this temperature indicates an antiferromagnetic state.

538.22:538.569.4 1948
Magnetic Resonance Line Shapes at the Onset of Saturation—D. F. Holcomb. (*Phys. Rev.*, vol. 112, pp. 1599–1603; December 1, 1958.) Results are given of measurements of magnetic resonance absorption line shapes and widths in Li metal and CaF_2 crystals, as a function of RF power level, up to the region of appreciable saturation.

538.22:538.569.4 1949
Paramagnetic Resonance of Fe^{3+} in Sapphire at Low Temperatures—G. S. Bogle and H. F. Symmons. (*Proc. Phys. Soc.*, vol. 73, pp. 531–532; March 1, 1959.)

538.221 1950
Search for New Heusler Alloys—D. P. Morris, R. R. Preston, and I. Williams. (*Proc. Phys. Soc.*, vol. 73, pp. 520–523; March 1, 1959.) A brief report is given of investigations on silver and gold ternary alloys.

538.221:534.213–8 1951
Ultrasonic Wave Propagation in a Nickel Single Crystal—J. de Klerk. (*Proc. Phys. Soc.*, vol. 73, pp. 337–344; March 1, 1959.) An improved pulse technique has been used to investigate the dynamic elastic constants and energy losses with and without an applied magnetic field. Energy losses are substantially reduced when the material is magnetized to saturation.

538.221:538.632 1952
Spin-Orbit Coupling and the Extraordinary Hall Effect—C. Strachan and A. M. Murray. (*Proc. Phys. Soc.*, vol. 73, pp. 433–447; March 1, 1959.) Quantum-mechanics transport theory is used to evaluate the magnitude of the extraordinary Hall coefficient.

538.221:538.632 1953
Two Hall Effects of Iron-Cobalt Alloys—F. P. Beitel, Jr., and E. M. Pugh. (*Phys. Rev.*, vol. 112, pp. 1516–1520; December 1, 1958.) The ordinary and extraordinary Hall coefficients, and the resistivity of Fe-Co alloys have been measured at 77°K, 169°K and room temperature. The results are analyzed in terms of two models for the electronic structure.

538.221:538.652:621.372.41 1954
Ferrites for Magnetostriction Oscillators in Filter Circuits—S. Schweizerhof. (*Nachrichtentech. Z.*, vol. 11, pp. 179–185; April, 1958.) The performance of ferrite rings with improved temperature characteristics and high Q values is discussed.

538.221:539.23 1955
Magnetization Reversal by Rotation and Wall Motion in Thin Films of Nickel-Iron Alloys—E. M. Bradley and M. Prutton. (*J. Electronics Control*, vol. 6, pp. 81–96; January, 1959.) Measurements of the 400-cps, hysteresis loops on unixaial films indicate that both coherent rotation and domain wall motion can occur depending on film orientation and thickness.

538.221:621.318.134 1956
Domain Behaviour in some Transparent Magnetic Oxides—R. C. Sherwood, J. P. Remeika, and H. J. Williams. (*J. Appl. Phys.*, vol. 30, pp. 217–225; February, 1959.) Magnetiic domains were observed by means of the Faraday effect and by the Bitter technique in a number of compounds with the spinel, magnetoplumbite, and perovskite-like structures.

538.221:621.318.134 1957
Some Investigations on Li-Zn Ferrites—Kh. S. Valeev, N. G. Drozdov, and A. L. Frumkin. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2517–2527; November, 1957.) As a result of an investigation of the influence of composition and temperature on the magnetic properties of Li-Zn ferrites, materials were obtained having low losses in the range 20–75 mc. The permeability at a wavelength of 3.2 cm was found to be less than 1.

538.221:621.318.134 1958
Domain Wall Motion and Ferrimagnetic Resonance in a Manganese Ferrite—J. F. Ditton, Jr., H. E. Earl, Jr. (*J. Appl. Phys.*, vol. 30, pp. 202–213; February, 1959.) Very simple domain walls were driven th ough single crystals of a high-resistivity Mn ferrite. A considerable temperature range was covered, through which the material constants varied substantially, as did the losses encountered. Ferrimagnetic resonance experiments are reported for the same material over the same temperature range.

538.221:621.318.134:621.357.7 1959
Preparing Ferrites by Continuous Electrolytic Co-precipitation—H. B. Beer and G. V. Planer. (*Brit. Commun. Electronics*, vol. 5, pp. 939–941; December, 1958.) A new method is

outlined having the advantage of greater economy and simplicity, with increased homogeneity and chemical purity of the product.

621.318.2 1960
Permanent Magnet Stability—J. E. Gould. (*Instrum. Practice*, vol. 12, pp. 1083–1091; October, 1958.) Consideration is given to the influence on magnetization of external magnetic fields, thermal effects, mechanical shock and nuclear radiation. Under normally steady conditions, magnetization changes occur which are proportional to the logarithm of time and can be minimized by choice of material, working permeance, and prestabilization.

621.318.2-492 1961
Powdered Magnets—G. Sideris. (*Elektronics*, vol. 32, p. 69; February 27, 1959.) The properties of sintered alloy, sintered oxide, and pressed-powder permanent magnets are tabulated.

669.71:621.795 1962
Aluminium Finishes for Use in Electronics—W. E. Pocock. (*Electronics*, vol. 32, pp. 58–59; February 20, 1959.) A survey of the properties and applications of various surface finishes.

MATHEMATICS

517.93 1963
Parametric Excitation—N. Minorsky. (*Compt. rend. Acad. Sci., Paris*, vol. 247, pp. 406–408; July 28, 1958.) Simplified solutions of the equation $\ddot{x}+b\dot{x}+x+(a-cx^2)x\cos 2t+cx^3=0$ are obtained by a development of the method described earlier (2951 of 1951).

518 1964
Some Properties of Strongly Connected Graphs—B. Roy. (*Compt. rend. Acad. Sci., Paris*, vol. 247, pp. 399–401; July 28, 1958.) Two general theorems are stated concerning flow-diagram analyses.

MEASUREMENTS AND TEST GEAR

531.76:621.374.32 1965
Vernier Chronotron times Nuclear Particle Flight—H. W. Lefevre and J. T. Russell. (*Electronics*, vol. 32, pp. 44–47; March 6, 1959.) A time interval analyzer with a resolution better than 10^{-9} seconds is described.

621.317.2:621.373.4 1966
Timed-Signal Generator with Flexible Output—D. E. Minow. (*Electronics*, vol. 32, pp. 52–53; March 6, 1959.) Details of a portable unit with two output channels delivering pulses of controllable duration, amplitude, carrier content and repetition rate.

621.317.3:537.311.33:621.3–71 1967
Cryostat for Measuring the Electrical Properties of High-Resistance Semiconductors at Low Temperatures—W. H. Mitchell and E. H. Putley. (*J. Sci. Instrum.*, vol. 36, pp. 134–136; March, 1959.) Resistances of up to $10^{11}\Omega$ may be measured at temperatures down to 2°K, using a vibrating-reed electrometer. The lower limit for Hall mobility is 10 cm²/v per second.

621.317.34.029.63/.64:621.372.5 1968
Three-Point Method of Measuring UHF Quadripoles—J. Šinejkal and L. Mollwo. (*Hochfreq. und Elektroak.*, vol. 66, pp. 167–169; March, 1958. Comment on 1213 of 1958 and author's reply.

621.317.35:621.372.54 1969
Calculation of the Resolving Power of Automatic Frequency Analysers—N. V. Terpugov. (*Radiotekh. Elektron.*, vol. 2, pp. 796–806; June, 1957.) A method is described for calculating the dynamic frequency characteristics of filter systems. Factors and coefficients for evaluating the characteristics are given and results of an experimental investigation of nine types of filter are tabulated.

621.317.373 1970
Phase-Angle Measurement—P. Kundu. (*Electronic Radio Eng.*, vol. 36, pp. 150–154; April, 1959.) The signals are applied to a heptode mixer whose differential anode current with respect to the reference value for quadrature inputs is a measure of phase.

621.317.373:621.3.05 1971
Measurement of Phase Difference on Long Power Transmission Lines—G. Zito. (*Alta Frequenza*, vol. 27, pp. 378–400; June/August, 1958.) The apparatus described can be used for determining, with an error not exceeding ±1 per cent, the phase relations between different power supply systems feeding a network of television stations.

621.317.42:537.311.33 1972
A New Method of Measuring Magnetic Field Intensity—G. E. Pikus and O. V. Sorokin. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2647–2651; November, 1957.) This method is based on the change of concentration of charge carriers in a thin semiconductor specimen located in a magnetic field and passing an alternating current. By virtue of a linear relation between the applied voltage and the magnetic field intensity the latter can be estimated in the range 5×10^{3}–10^{-5} oersteds.

621.317.44:538.632 1973
Magnetic Field Probe of High Sensitivity and Resolution—B. Kostyshyn and D. D. Roshon, Jr. (PROC. IRE, vol. 47, p. 451: March 1959.) Note on the performance characteristics of a miniature Hall-effect probe of Bi

621.317.71:621.314.7 1974
Transistor Junction Temperature—H. Sutcliffe and D. J. Matthews. (*Electronic Radio Eng.*, vol. 36, pp. 143–144; April, 1959.) A circuit is described for measurement of the temperature-dependent base leakage current in a class-C transistor stage.

621.317.742:621.317.755 1975
4000-Mc/s-Band Wide-Band VSWR Scanner—Y. Ninomiya, N. Miyamoto, and A. Yanagi. (*Rep. elect. Commun. Lab., Japan*, vol. 6, pp. 154–157; May, 1958.) Voltage SWR is displayed on an oscilloscope over the frequency range 3600–4200 mc.

621.317.75:621.374 1976
Amplitude Slicer for Signal Analysis—T. A. Bickart. (*Electronics*, vol. 32, pp. 64–65; February 27, 1959.) Description of a circuit providing a rectangular output pulse whose width is proportional to the time during which the input signal lies between specified voltage levels.

621.317.755 1977
The Cathode-Ray Oscilloscope: a Survey—J. F. Golding. (*Brit. Commun. Electronics*, vol. 6, pp. 27–33; January, 1959.) Abridged specifications of oscilloscopes available in U. K. are tabulated.

621.317.755.087.5 1978
Automatic Recorder for Cathode-Ray Oscillography—H. Lindner and U. Gladhorn. (*Arch. tech. Messen*, pp. 79–82 and 171–174; April, August, 1958.) An oscilloscope camera for film or drum recording is described. With an adaptor the camera can be used at right angles to the cathode ray tube screen. Details are also given of an electronic control unit for the provision of triggering and brightening pulses.

621.317.763:621.314.7 1979
Transistorized Absorption Wavemeter—G. W. Short. (*Wireless World*, vol. 65, pp. 193–196; April, 1959.) A description of a wavemeter incorporating a modulating oscillator and covering the frequency range 1–100 mc.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

526.2:621.396.9 1980
Electronic Principles of the Tellurometer—T. L. Wadley. (*Trans. S. Afr. Inst. Elec. Eng.*, vol. 49, pt. 5, pp. 143–161; May, 1958. Discussion, pp. 161–172.) A detailed description of the instrument. See also 3250 of 1957 and 874 of 1958 (Hammond).

531.767:621.396.96 1981
Radar Meter helps enforce Traffic Laws—J. Barker. (*Electronics*, vol. 32, pp. 48–49; March 6, 1959.) A battery-powered system is described, converting Doppler shift to give a direct reading of vehicle speed.

538.569.4:621.372.8 1982
Low-Power Microwave Reflection Bridge—R. C. Rempel and H. E. Weaver. (*Rev. Sci. Instrum.*, vol. 30, p. 137; February, 1959.) A standard microwave bridge for detection of electron paramagnetic resonance is modified by introducing an adjustable ferrite isolator in the sample arm.

621.313.334:621.318.57:621.314.7 1983
Four Transistor Inverter drives Induction Motor—W. H. Card. (*Electronics*, vol. 32, pp. 60–61; February 20, 1959.) Direct-current motors used in low-pressure or explosive environments can be replaced with induction motors by employing transistors as controlled switches to provide two-phase square-wave output from a single dc source.

621.384.6 1984
Longitudinal Space-Charge Effects in Particle Accelerators—C. E. Nielsen and A. M. Sessler. (*Rev. Sci. Instrum.*, vol. 30, pp. 80–89; February, 1959.) "The modification of the single-particle theory of particles subject to RF acceleration caused by electrostatic repulsion between particles is calculated."

621.384.611 1985
A Fixed-Frequency Cyclotron with One Dee—R. Bock, A. Doehring, J. Jänecke, O. Knecht, L. Koester, H. Maier-Liebnitz, C. Schmelzer, and U. Schmidt-Rohr. (*Z. angew. Phys.*, vol. 10, pp. 49–55; February, 1958.) The design and construction of a 12-mev deuteron accelerator installed at Heidelberg are described.

621.384.7:537.54 1986
High-Current Ion Source—R. G. Meyerand, Jr. and S. C. Brown. (*Rev. Sci. Instrum.*, vol. 30, pp. 110–111; February, 1959.) An ion source of simple design and construction is described capable of producing a pulsed ion beam of $\frac{1}{2}$A.

621.385.833 1987
A High-Resolution Emission Microscope for Viewing Surfaces with Electrons Released by Ultraviolet Radiation—W. Koch. (*Z. Phys.*, vol. 152, pp. 1–18; July 11, 1958.) The 40-kv microscope described has an electron-optical magnification of 700 with a resolution of 1000 Å.

621.385.833 1988
Electrostatic Charging of the Photosensitive Material in Electron Microscopes—E. Kinder. (*Z. angew. Phys.*, vol. 10, pp. 95–98; February, 1958.) Methods of eliminating or reducing excessive charges are discussed.

621.387.4:621.395.625.3 1989

Magnetic Recording of Pulse-Amplitude Data—J. Baumgardner. (*Rev. Sci. Instrum.*, vol. 30, pp. 134–135; February, 1959.) The limitations and advantages of the direct recording of pulses on magnetic tape are considered.

621.387.462 1990

Diamond Conduction Counters with Small Electrode Separations—F. C. Champion and S. B. Wright. (*Proc. Phys. Soc.*, vol. 73, pp. 385–392; March 1, 1959.) Measurements of the charge pulse-height/applied-field curve at electrode spacings from 10 μ to 1 mm show variations in the characteristics of the curves which cannot be explained in terms of trapping field distribution or field distortion due to dark current.

621.397.3:621.39 1991

Automatic Character Recognition—K. Steinbuch. (*Nachrichtentech. Z.*, vol. 11, pp. 210–219 and 237–244; April, May, 1958.) Detailed investigation of the problems of recognizing written or printed characters, particularly numerals. A number of scanning methods are reviewed. 30 references including patents.

621.398:621.376.55 1992

Telemetry Demodulator using Modified AND Gate—L. Weisman. (*Electronics*, vol. 32, pp. 54–57; February 20, 1959.) A description of a pulse-position telemetry system, with details of the demodulator channels.

PROPAGATION OF WAVES

621.396.11:523.53 1993

Observations of Direction of Arrival of Long-Duration Meteor Echoes in Forward Scatter Propagation—T. Hagfors and B. Landmark. (*J. Geophys. Res.*, vol. 64, pp. 19–22; January, 1959.) The angular distribution of enduring meteor bursts is shown to be similar to that observed for short-duration specular reflections and in marked contrast to that of the turbulent background component. It is concluded that the long-duration echoes are from specularly reflecting meteor trails and not from trails broken up by atmospheric turbulence.

621.396.11:551.311.122:537.226 1994

The Electromagnetic Properties of Glacier Ice—M. Lafargue and R. Millecamps. (*Compt. rend. Acad. Sci., Paris*, vol. 247, pp. 884–886; September 22, 1958.) Experiments were made on em wave propagation at frequencies in the range 50–400 kc in a glacier. The ice appears to behave as a low-loss dielectric. At 150 kc reception was possible at a distance of 5 km from a 1 w transmitter on the glacier.

621.396.11:551.510.52 1995

Calculation of Multiple Dispersion in U.S.W. Scatter Propagation in the Troposphere—D. M. Vysokovskiĭ. (*Radiotekh. Elektron.*, vol. 2, pp. 807–809; June, 1957.) An approximate evaluation of energy loss due to scatter.

621.396.11:551.510.535 1996

Propagation of Electrical Waves along a Plasma Layer Bounded by a Dielectric with a Longitudinal Magnetic Field—W. O. Schumann. (*Z. angew. Phys.*, vol. 10, pp. 26–31; January, 1958.) See also 1881 of 1957.

621.396.11:551.510.535 1997

An Analysis of Drifts of the Signal Pattern associated with Ionospheric Reflections—Yerg. (See 1884.)

621.396.11:551.510.535 1998

The Effects of Ionospheric Irregularities and the Auroral Zone on the Bearings of Short-Wave Radio Signals—H. A. Whale. (*J. Atmos. Terrest. Phys.*, vol. 13, pp. 258–270; February, 1959.) For sunlit paths less than 15,000 km most of the daily variation of bearing of F_2-propagated signals is said to arise from refraction in the F_1 region. During night-time, obstructing patches of E-region ionization are suggested as the cause. For paths longer than 15,000 km the direction of the transmitter beam may be relatively unimportant in determining the incoming bearing, and large observed bearing changes are supposed to arise from absorbing and scattering processes in the auroral regions. A method for plotting the shape of the absorbing parts of the auroral zones is described.

621.396.11:621.396.67 1999

Transmission of Power in Radio Propagation—J. R. Wait. (*Electronic Radio Eng.*, vol. 36, pp. 146–150; April, 1959.) As the transmission of power between antennas can be influenced appreciably by the immediate neighborhood of the antennas, it is suggested that the total transmission loss should be divided into two parts, one of which should account for this effect.

621.396.11.029.6 2000

Symposium on Long-Distance Propagation above 30 Mc/s—(*Proc. IEE*, vol. 105, Pt. B, suppl. no. 8, pp. 1–191; 1958.) The following papers were read at the IEE Symposium held in London, January 28, 1958.

Ionospheric Forward-Scatter Propagation:

a) **Survey of the Gibralter-United Kingdom Ionospheric Scatter Measurements**—F. A. Kitchen and G. Millington (pp. 2–6).

b) **A Scatter-Signal Analyser**—P. H. Cutler and D. Williams (pp. 7–11).

c) **The Choice of Aerial Height for Ionospheric Scatter Links**—E. Fitch and R. Ruddlesden (pp. 12–18).

d) **The Structure of High-Frequency Ionospheric Scatter Signals**—D. Williams (pp. 19–26).

e) **Radio Interference as a Factor in Ionospheric Scatter Communication**—G. A. Isted (pp. 27–35).

f) **Analysis of Gibraltar-United Kingdom Ionospheric Scatter Signal Recordings**—G. A. Isted (pp. 36–44).

g) **Polar-Diagram Requirements for Aerials for Communication by Ionospheric Scatter**—D. H. Shinn (pp. 45–52).

h) **The Angular Distribution of Energy Received by Ionospheric Forward Scattering at Very High Frequencies**—W. C. Bain (pp. 53–55).

i) **The Direction and Amplitude of Reflections from Meteor Trails and Sporadic-*E* Ionization on a 1740-km North-South Path at Very High Frequencies**—R. W. Meadows (pp. 56–64).

j) **Short Bursts of Amplitude of a 50-Mc/s Wave Received over a Distance of 480 km**—G. S. Kent (pp. 65–69).

k) **Amplitude of Very-High-Frequency Signals Reflected from the Sporadic-*E* Layer in North-West Europe**—P. J. Brice (pp. 70–72).

Discussion (pp. 73–78). Tropospheric Propagation Beyond the Horizon:

l) **Guglielmo Marconi and Communication Beyond the Horizon: a Short Historical Note**—G. A. Isted (pp. 79–83).

m) **A Survey of Tropospheric Wave Propagation Measurements by the B.B.C., 1946–1957**—R. A. Rowden, L. F. Tagholm, and J. W. Stark (pp. 84–90).

n) **The Measurement and Prediction of V.H.F. Tropospheric Field Strengths at Distances Beyond the Horizon**—J. K. S. Jowett (pp. 91–96).

o) **The Effects of Atmospheric Discontinuity Layers up to and including the Tropopause on Beyond-the-Horizon Propagation Phenomena**—B. J. Starkey, W. R. Turner, S. R. Badcoe, and G. F. Kitchen (pp. 97–105).

p) **Some Investigations of Metre-Wave Radio Propagation in the Transhorizon Region**—F. A. Kitchen, E. G. Richards, and I. J. Richmond (pp. 106–116).

q) **The Reduction of Threshold by the Use of Frequency Compression**—A. J. Buxton and M. O. Felix (pp. 117–121).

r) **Propagation Measurements at 3480 Mc/s over a 173-Mile Path**—B. C. Angell, J. B. L. Foot, W. J. Lucas, and G. T. Thompson (pp. 128–142).

s) **Some Tropospheric Scatter Propagation Measurements and Tests of Aerial Siting Conditions at 858 Mc/s**—G. C. Rider (pp. 143–152).

t) **The Long-Range Propagation of Radio Waves at 10-cm Wavelength**—W. R. R. Joy (pp. 153–157).

u) **Radio Propagation Far Beyond the Horizon at about 3.2-cm Wavelength**—W. R. R. Joy (pp. 158–164).

v) **A Review of Tropospheric Scatter Propagation Theory and its Application to Experiment**—M. A. Johnson (pp. 165–176).

w) **The Estimation of Transmission Loss in the Transhor zon Region**—E. G. Richards (pp. 177–183).

Discussion (pp. 122–126, 184–188).

621.396.11.029.62:551.510.535:551.594.5 2001

A Bistatic Radio Investigation of Auroral Ionization—C. Collins and P. A. Forsyth. (*J. Atmos. Terrest. Phys.*, vol. 13, pp. 315–345; February, 1959.) The scattering of radio waves in the upper atmosphere at times of auroral disturbance has been studied by means of some twenty 30–50-mc radio systems in Canada, each having transmitter and receiver about 1000 km apart. At least four different kinds of auroral events are distinguishable. Of these, two appear to be associated with different phases of visible aurora, the third with a later stage in the auroral process which is not observed visually, and a fourth with the recurrent daytime absorption which often precedes auroral disturbance. In these four events evidence is found for three separate scattering mechanisms, each of which has been proposed previously as the principal source of radar echoes from aurora.

621.396.11.029.63:621.396.812.3 2002

Propagation Tests at 1000 Mc/s with Diversity Reception between Monte Penice and Monte Venda—P. Quarta. (*Alta Frequenza*, vol. 27, pp. 219–225; June/August, 1958.) Analysis and discussion of test results obtained in 1954 over a 196-km path. See also 3522 of 1956 (Vecchiacchi).

621.396.8.029.62 2003

Long-Distance V.H.F. Reception—H. V. Griffiths. (*Wireless World*, vol. 65, pp. 179–181; April, 1959.) An analysis of observations made since 1946 shows that interference on band I is due to three modes of propagation, F-layer, sporadic-E, and tropospheric. The intensity of the interfering signal has not been measured, but the frequency of occurrence is correlated with the sunspot number for reception via the F layer and sporadic E.

RECEPTION

621.376.23 2004

The Optimum Detector with Log I^0 Characteristic for the Detection of Weak Signals in Noise—B. S. Fleyshman. (*Radiotekhn. Elektron.*, vol. 2, pp. 726–734; June, 1957.) The characteristics of the optimum log I_0 detector [see *e.g.* 2782 of 1953 (Middleton)] are calculated. Inaccuracies in earlier calculation of the I_0 expansion are pointed out and the application of the derived expression to radar is considered.

621.376.3 2005

Passage of Random Noise Signals through a Detector considering Biasing and Limiting

Effects—E. G. Logachev. (*Radiotekhn. Elektron.*, vol. 2, pp. 735–750; June, 1957.) A general expression is derived for the correlation function of the noise current at the output of a detector in the low-frequency region.

621.396.621.54:621.385.029.6 2006

Carcinotron Harmonics boost Receiver Range—C. H. Currie. (*Electronics*, vol. 32, pp. 58–61; February 27, 1959.) A continuous frequency coverage from 30 mc to 75 kmc is obtained by using harmonic mixing. The carcinotron local oscillator operates in the band 2–4 kmc and supplies two separate RF sections.

STATIONS AND COMMUNICATION SYSTEMS

621.376.018.78 2007

The Amplitude and Frequency of a Modulated Carrier Wave—A. Ditl. (*Hochfreq. und Elektroak.*, vol. 66, pp. 160–167; March, 1958.) Both AM and FM systems are considered. Signal distortion due to linear distortion of the carrier, the distortion in SSB systems, the effect of short pulses on the output signal, the interference between FM carriers, and the effect of reflection in FM systems are investigated with examples.

621.391 2008

Discrimination between Several Signals in Telecommunication—P. Béthoux. (*Compt. rend. Acad. Sci., Paris*, vol. 247, pp. 412–415; July 28, 1958. Mathematical treatment of signal discrimination in noise.

621.391:621.376.56 2009

Signal/Noise Ratio in Pulse-Code Modulation Systems: Use of the "Ideal Observer" Criterion—J. W. R. Griffiths. (*J. Brit. IRE*, vol. 19, pp. 183–186; March, 1959.) The "ideal observer" criterion is applied to determining the probability of error in selecting a single pulse in a background of noise. The results are similar to those obtained by a method due to Flood (3970 of 1958).

621.391.1:621.396.14 2010

Theoretical Considerations about the Merits of the Normal Binary Telegraph Code and the So-Called Gaussian Code, and Special Methods of Detection for Both Codes—K. Posthumus. (*Tijdschr. ned. Radiogenoot.*, vol. 23, pp. 55–82; March, 1958.)

621.396.2:621.394.4 2011

Multichannel V.F. Telegraph Systems for H.F. Networks—J. V. Beard. (*Point to Point Telecommun.*, vol. 3, pp. 29–48; October, 1958.) A general description of a two-tone frequency-diversity system which gives, under conditions of selective fading, a tenfold reduction in error rate compared with a comparable FM system using the same bandwidth.

621.396.43:550.389.2:629.19 2012

Transoceanic Communication by means of Satellites—J. R. Pierce and R. Kompfner. (PROC. IRE, vol. 47, pp. 372–380; March, 1959.) "A satellite in a polar orbit at a height of 3000 miles would be mutually visible from Newfoundland and the Hebrides for 22.0 per cent of the time and would be over 7.25° above the horizon at each point for 17.7 per cent of the time. Out of 24 such satellites, some would be mutually visible over 7.25° above the horizon 99 per cent of the time. With 100-foot-diameter spheres, 150-foot-diameter antennas, and a noise temperature of 20°K, 85 kw at 2000 mc or 9.5 kw at 6000 mc could provide a 5-mc base band with a 40-db signal-to-noise ratio."

621.396.65 2013

Radio-Link Equipment for 60–120 Channels—G. Strocchi. (*Alta Frequenza*, vol. 27, pp. 269–291; June/August, 1958.) Three types of Italian equipment are described, including that used in the radio link Milan-Palermo [2230 of 1958 (Peroni)].

621.396.65 2014

Convention on Radio Links—(*Alta Frequenza*, vol. 27, pp. 177–432; June/August, 1958.) Second issue covering the proceedings of a convention held in Rome, June 5–8, 1957. First issue: 2229 of 1958. Abstracts of some of the papers are given individually; titles of others are as follows:

a) **The Trans-Appennine Radio Link**—G. Monti-Guarnieri (pp. 179–218).

b) **Realization and Future Application of Pulse Techniques in Radio Communication Networks**—R. Cabessa (pp. 226–235, in French).

c) **Diversity Systems and their Influence on the Economic Operation of Radio Links**—P. Clavier (pp. 236–244, in French).

d) **Some Design Problems in F.M. Broad-Band Microwave Systems**—B. Håård (pp. 245–262, in English).

e) **The Evaluation of Transmission Quality in Multichannel F.M. Radiotelephony Links**—I. Medici (pp. 347–362.)

f) **Field-Strength Recordings and Performance of Very-Short-Wave Radio Links**—J. A. Smale (pp. 363–377, in English).

621.396.71.029.55 2015

The Olifantsfontein and Derdepoort Radio Stations of the Department of Posts and Telegraphs—A. Birrell. (*Trans. S. Afr. Inst. Elec. Eng.*, vol. 49, pt. 6, pp. 177–228; June, 1958. Discussion, pp. 228–231.) A detailed description of the Olifantsfontein transmitting and Derdepoort receiving stations and of their part in the radio telephone and telegraph services of the Union of South Africa is given. A VHF radio link operating at 100 mc for line-of-sight transmission of standard-frequency and time signals from the Union Observatory to Olifantsfontein is under construction; the signals are to be broadcast on 5, 10, 15, and 20 mc.

621.396.712.3 2016

Broadcasting Equipment at the New Studio in Karlsruhe—W. Hoffmann. (*Rundfunktech. Mitt.*, vol. 2, pp. 100–105; June, 1958.) Details are given of the control room, recording and distribution installations.

621.396.712.3:534.861 2017

The Acoustic Design of the New Studio in Karlsruhe—Keidel. (See 1756.)

621.396.73 2018

The Megacoder—a High-Speed, Large-Capacity Microminiature Decoder for Selective Communication—H. Kihn and W. E. Barnette. (*RCA Rev.*, vol. 20, pp. 153–179; March, 1959.) The device can be preset to respond to any one of a million possible code combinations for use in FM personal paging systems.

SUBSIDIARY APPARATUS

621.3.087.9:621.318.4 2019

I.R.E. Standards on Static Magnetic Storage: Definitions of Terms, 1959—(PROC. IRE, vol. 47, pp. 427–430; March, 1959.) Standard 59 IRE 8.S1.

621.3.087.9:621.395.625.3 2020

Magnetic Head reads Tape at Zero Speed—M. E. Anderson. (*Electronics*, vol. 32, pp. 58–60; March 6, 1959.) Design details of a system enabling recorded HF signals to be played back at speeds low enough for the output to be fed to a pen-recorder without deterioration in signal/noise ratio.

621.314.58:621.314.7 2021

Development of the Transistor Inverter at 20 kc/s using Power Transistors—W. A. Martin. (IRE TRANS. ON INSTRUMENTATION, vol. I-6, pp. 118–122; June, 1957.)

621.314.63:621.39 2022

Physical and Electrical Properties of Silicon Rectifiers for Communications Applications—H. L. Rath. (*Elektron. Rundschau*, vol. 12, pp. 119–122; April, 1958.) Small junction-type diodes with a current-carrying capacity of up to 1 A are considered.

621.352:541.135.6 2023

Current Integration with Solion Liquid Diodes—R. N. Lane and D. B. Cameron. (*Electronics*, vol. 32, pp. 53–55; February 27, 1959.) The construction and characteristics of electrochemical diodes using iodine potassium-iodide solution are described (see 1563 of 1958). These can be used as electrical current integrators and flow or pressure detectors.

TELEVISION AND PHOTOTELEGRAPHY

621.397.611:778.5 2024

The Film Recording of Television Transmissions in the German Federal Republic—J. Goldmann and H. Funk. (*Rundfunktech. Mitt.*, vol. 2, pp. 129–136; June, 1958.) Review of methods used and description of installations.

621.397.611.2 2025

A Vidicon Camera for Industrial Colour Television—I. L. P. James. (*J. Brit. IRE*, vol. 19, pp. 165–180; March, 1959. Discussion, pp. 181–182.) The main features of a simultaneous color system employing three vidicons are described and signal amplifiers and line and field scanning circuits are discussed. The picture quality obtained is adequate for 625-line broadcast standards and the equipment is universally applicable for general industrial use.

621.397.611.2 2026

Contribution on the Problem of Portable Television Cameras for Outside Broadcasts—H. Fix. (*Rundfunktech. Mitt.*, vol. 2, pp. 120–128; June, 1958.) The design and application of portable television cameras are discussed and three cameras, including the French Type CP103 [see 3650 of 1958 (Polonsky)], are compared.

621.397.62 2027

Inexpensive Sound for Television Receivers—R. B. Dome. (*Electronics*, vol. 32, pp. 66–68; February 27, 1959.) The system provides AM compression, a high-level AF output, FM detection and cancellation of the AM fundamental frequency.

621.397.62:621.376.33:621.314.7 2028

TV Sound Detector uses Drift Transistor—M. Meth. (*Electronics*, vol. 32, pp. 62–64; February 20, 1959.) Circuit details of a sensitive, oscillating linear-slope detector giving improved performance at low signal levels compared with a passive detector.

621.397.621:535.623 2029

Results with an Experimental Colour Television System using Controlled Colour Filters—V. A. Babits. (*Brit. Commun. Electronics*, vol. 6, p. 15; January, 1959.) In a brief report, reference is made to problems to be overcome. For an account of similar work see 621 of February (Wells).

621.397.7(71/73) 2030

Television Station List—M. I. Schiller. (*Radio-Electronics*, vol. 30, pp. 106–107; January, 1959.) A list of U. S., Canadian and Mexican stations correct to December 1, 1958 giving call sign, location and channel number.

621.397.8 2031

A Method for the Measurement of Random

Fluctuations in Television—D. Waechter. (*Rundfunktech. Mitt.*, vol. 2, pp. 117–119; June, 1958.) In the comparison method described adjustable random noise is superimposed on a small area of the picture under test.

621.397.8.083:535.623 2032

A Simple Method of Mixing a Colour Subcarrier of Variable Frequency with a Black and White Picture—G. Bolle. (*Frequenz*, vol. 12, pp. 103–108; April, 1958.) Test equipment is described for investigating the interference caused by the color subcarrier of the N.T.S.C. system in monochrome picture reproduction. A constant-amplitude carrier is continuously variable in the frequency ranges 1.5–3.0 mc and 3.0–4.5 mc.

621.397.9:629.136.3 2033

Single-Line Television—F. H. Harris and J. Ainsworth. (*Rev. Sci. Instrum.*, vol. 30, pp. 76–78; February, 1959.) Description of equipment constructed to show the practicability of using a vidicon camera use for measuring the space orientation of spin-stabilized rockets.

TRANSMISSION

621.376.32 2034

A Reactance-Valve Frequency Modulator—F. Carassa. (*Alta Frequenza*, vol. 27, pp. 292–303; June/August, 1958.) A portable modulator unit used for feeding television outside-broadcast programs into a radio-link network is described.

TUBES AND THERMIONICS

621.314.63 2035

The Problem of Representation of a Semiconductor Diode in the Form of a Series Connection of Two Nonlinear Inertia Elements and the Applicability of the Pulse Method of Voltage Division—Yu. K. Barsukov. (*Zh. Tekh. Fiz.*, vol. 27, pp. 2262–2267; October, 1957.) The relation of the over-all I/V characteristic of a Ge diode to the I/V characteristic of the p-n junction and the volume of Ge is considered, and the applicability of a pulse method for determining the division of voltage in the circuit representation of the diode is discussed.

621.314.63 2036

The Inductive Behaviour of *P-N* Rectifiers under High Forward-Current Loads—E. Spenke. (*Z. angew. Phys.*, vol. 10, pp. 65–88; February, 1958.) A simplified model of a p-n junction and its equivalent circuit are used as a basis for detailed calculations of the ac characteristics, to account for the inductive component of the rectifier impedance.

621.314.63 2037

Semiconductor—Semiconductor "Point-Contact" Diode—A. Levitas and I. Ladany. (*J. Appl. Phys.*, vol. 30, pp. 267–268; February, 1959.). Description of a technique developed for the fabrication of a junction device made from a single-crystal Ge bar containing a grown p-n junction, with an external appearance similar to a point-contact diode.

621.314.63.012.6:621.317.6 2038

Evaluating Logarithmic Diodes—A. Gill. (*Electronics*, vol. 32, pp. 64–67; March 6, 1959.) Note on a method of determining the low-level characteristic of a semiconductor diode, using a sawtooth input voltage and obtaining a cro trace showing di/dt as a function of current i.

621.314.63:621.316.722.1 2039

Characteristics of Silicon Junction Diodes as Precision Voltage-Reference Devices—K. Enslein. (IRE TRANS. ON INSTRUMENTATION, vol. I-6, pp. 105–118; June, 1957.)

621.314.63:621.318.57 2040

Two-Terminal Solid-State Switches—T. P. Sylvan. (*Electronics*, vol. 32, pp. 62–63; February 27, 1957.) Characteristics of commercial p-n-p-n and p-n-p-m semiconductor diodes are tabulated. See *e.g.*, IRE TRANS. ON ELECTRON DEVICES, vol. ED-5, pp. 13–18; January, 1958. (Philips and Chang.)

621.314.7+621.385.3 2041

Simple General Analysis of Amplifier Devices with Emitter, Control, and Collector Functions—E. O. Johnson and A. Rose. (PROC. IRE, vol. 47, pp. 407–418; March, 1959.) The photoconductor, unipolar and bipolar transistors, vacuum triode, analog transistor, and beam-deflection tube are considered. The characteristics are compared and discussed for particular applications.

621.314.7 2042

Contribution on the Representation of the A.C. Characteristics of the Earthed-Base Transistor—H. Schneider. (*Nachrtech.*, vol. 8, pp. 126–129; March, 1958.)

621.314.7 2043

On the Lifetime and Diffusion Constant of the Injected Carriers and the Emitter Efficiency of a Junction Transistor—S. Deb and A. N. Daw. (*J. Electronics Control*, vol. 5, pp. 514–530; December, 1958.) Experimental data are given for three types of p-n-p alloy-junction transistor. Variations of minority carrier lifetime and diffusion coefficient with emitter current, temperature, and carrier injection level are examined in relation to theory.

621.314.7:621.3.012.029.63 2044

U.H.F. Transistor Data—H. Tulchin. (*Electronics*, vol. 32, p. 57; March, 1959.) Characteristics are presented for eight commercially available transistors with operating frequencies above 300 mc.

621.314.7:621.317.71 2045

Transistor Junction Temperature—Sutcliffe and Matthews. (See 1974.)

621.314.7:621.318.57 2046

Solid-State Thyratrons Available Today—T. P. Sylvan. (*Electronics*, vol. 32, pp. 50–51; March 6, 1959.) Characteristics of three-terminal switching transistors are tabulated.

621.314.7:621.396.822 2047

The Influence of Inductive Source Reactance on the Noise Figure of a Junction Transistor—E. R. Chenette. (PROC. IRE, vol. 47, pp. 448–449; March, 1959.) Theoretical predictions for the equivalent noise resistance agree well with measurements, but the correlation reactance does not, except at low frequencies. Reasons for this are given.

621.314.7.001.4 2048

Complete Linear Characterization of Transistors from Low through Very High Frequencies—H. G. Follingstad. (IRE TRANS. ON INSTRUMENTATION, vol. I-6, pp. 49–63; March, 1957. Abstract. PROC. IRE, vol. 45, p. 898; June, 1957.)

621.314.7.002.2 2049

Techniques of Transistor Production—R. N. Wheaton. (*Proc. IRE, Australia*, vol. 19, pp. 358–369; July, 1958.) Requirements of a transistor suitable for HF operation are discussed and fabrication methods are reviewed, in particular alloying and diffusion techniques. Characteristics of different types of transistor are tabulated.

621.314.7.012.8 2050

Unified Representation of Junction-Transistor Transient Response—A. Harel and J. F. Cashen. (*RCA Rev.*, vol. 20, pp. 136–152; March, 1959.) A general mathematical formula is derived which is applicable to any circuit configuration of the transistor.

621.314.7.016.35 2051

Methods of Calculation for the Stabilization of Transistor Circuits at Variable Temperature—K. Lunze. (*Nachr. tech.*, vol. 8, pp. 98–108; March, 1958.)

621.314.7.078:621.316.825 2052

Application of Negative-Temperature-Coefficient Resistors to Temperature Stabilization of Transistor Circuits—C. Wright. (*Proc. IRE, Australia*, vol. 19, pp. 374–376; July, 1958.)

621.383.032.217.2 2053

Interference Photocathodes of Increased Yield with Freely Variable Maximum Spectral Response—K. Deutscher. (*Z. Phys.*, vol. 151, pp. 536–555; July 1, 1958.) Full report of an investigation noted earlier (2271 of 1958).

621.383.27 2054

The Linearization of Multipliers at High Anode Currents—H. J. Kopp and W. Petzold. (*Z. angew. Phys.*, vol. 10, pp. 34–36; January, 1958.) A method of obtaining a linear relation in the measurement of light intensity by means of photomultipliers is described.

621.383.5 2055

Bismuth-Tellurium Photovoltaic *P-N* "Sandwich" Layer—T. Piwkowski. (*Nature, London*, vol. 182, pp. 1793–1794; December 27, 1958.) A layer of Te evaporated onto a layer of Bi backed by a glass plate was found to act as a barrier-type photocell.

621.383.5 2056

Influence of Selenium Microstructure on Photocell Characteristics—T. K. Lakshmanan. (*J. Appl. Phys.*, vol. 30, pp. 265–266; February, 1959.)

621.385.029.6 2057

The Cut-Off Characteristics of Magnetrons (Static Regime)—W. Fulop. (*J. Electronics Control*, vol. 5, pp. 531–548; December, 1958.) Experiments show strong emission dependence and indicate that the anomalous current flow arises from electron interaction. The electron ensemble is shown to be far removed from thermal equilibrium and to approach this condition, though never reaching it, only at extremely high emission currents. Indications are given of theoretical trends needed to account for the experimental results.

621.385.029.6 2058

On Space-Charge Waves—D. H. Trevena. (*J. Electronics Control*, vol. 6, pp. 50–64; January, 1959.) The existence is shown of four sets of space-charge waves in a uniform electron beam focused by an axial magnetic field with arbitrary uniform cathode flux. The magnetic field affects the results only through the cathode flux. Expressions for the plasma frequency reduction factors are given.

621.385.029.6 2059

The Theory of the Formation of Electron Beams—V. T. Ovcharov. (*Radiotekh. Elektron.*, vol. 2, pp. 696–704; June, 1957.) A method is described for calculating the electric field inside an electron beam with prescribed trajectories and magnetic field taking into account the charge of the beam. By an appropriate choice of orthogonal curvilinear coordinates for the trajectory an ordinary second-order differential equation is obtained.

621.385.029.6 2060

The Effect of the Inclination of the Focusing Electrodes on Electron-Beam Formation—R. J. Lomax. (*J. Electronics Control*, vol. 6, pp. 39–49; January, 1959.) It is shown that variations of the angle at which Pierce electrodes meet the cathode give rise to variations in the angle at which the beam emerges, and that the current density at the periphery of the cathode depends markedly on this angle.

621.385.029.6 2061

Periodic Electrostatic Focusing of Laminar Parallel-Flow Electron Beams—W. W. Siekanowicz and F. E. Vaccaro. (PROC. IRE, vol. 47, pp. 451–452; March, 1959.)

621.385.029.6 2062

Electron Waves in Retarding System Non-linear Equations for Travelling-Wave Valves—L. A. Vaĭnshteĭn. (*Radiotekh. Elektron.*, vol. 2, pp. 688–695; June, 1957.) Gneralization of the results of an earlier analysis (338 of 1957) with application to the nonlinear theory of travelling-wave tubes.

621.385.029.6 2063

The Engineering of Low-Noise Travelling-Wave Tubes—F. J. Bryant. R. B. Coulson, and J. K. Fowler. (*Brit. Commun. Electronics*, vol. 6, pp. 20–35; January, 1959.) Design features and operating characteristics of *C*-band and *L*-band amplifiers are given.

621.385.029.6 2064

The Exponential Gun—a Low-Noise Gun for Travelling-Wave-Tube Amplifiers—A. L. Eichenbaum and R. W. Peter. (*RCA Rev.*. vol. 20, pp. 18–56; March, 1959.) The gun section between the cathode region and the circuit input is treated as a transmission-line matching transformer. The exponential gun is shown to be the best solution for the gun requirements of a low-noise travelling-wave tube.

621.385.029.6:[621.375.9+621.372.632 2065

A Three-Frequency Electron-Beam Parametric Amplifier and Frequency Converter—W. H. Louisell. (*J. Electronics Control*, vol. 6, pp. 1–25; January, 1959.) Analysis of Louisell and Quate (2273 of 1958) is generalized for the case of an electron beam in which the pump frequency ω need not be twice the signal frequency ω_2. If $\omega_1 \neq \omega_2$, where ω_2 is the idler frequency generated in the beam, growing and decaying fast space-charge waves can be excited equally, independent of the phase of the pump relative to the signal. The threshold of modulation needed to produce gain increases with beam size. If $\omega = \omega_1 - \omega_2$, the device acts as a frequency converter.

621.385.029.6:621.375.9 2066

Gain, Bandwidth and Noise in a Cavity-Type Parametric Amplifier using an Electron Beam—G. Wade and H. Heffner. (*J. Electronics Control*, vol. 5, pp. 497–509; December, 1958.) A modulated beam flowing across the gap of a resonant cavity changes the gap capacitance at the modulating frequency. It is shown that complete cancellation simultaneously of the two uncorrelated noise sources in the beam, while feasible in principle, is virtually impossible in practice. Conflicting requirements of large beam current for acceptable capacitance variation and large plasma wavelength for optimum noise cancellation lead to practical minimum noise figures of about 3 db. Design data are given for an amplifier with pump frequency 2 kmc, a gain of about 15 db at 500 mc and bandwidth 43 kc. A noise figure of 3.4 db could be achieved with some difficulty.

621.385.029.6:621.375.9:621.372.2 2067

Travelling-Wave Couplers for Longitudinal Beam-Type Amplifiers—R. W. Gould. (PROC. IRE, vol. 47, pp. 419–426; March, 1959.) The theory is developed and applied to the design of couplers for parametric amplifiers. Matrices for velocity jumps and drift regions are given, and conditions for the removal of beam noise from the fast space-charge wave are derived.

621.385.032.213 2068

Development of Thermionic Cathodes—B. M. Tsarev. (*Radiotekh. Elektron.*, vol. 2, pp. 675–687; June, 1957.) The general requirements of thermionic cathodes are discussed in relation to current applications in vacuum tubes. A classification is given of different types of cathode in use and under investigation, and the characteristics of cathode materials are tabulated.

621.385.032.213.13 2069

A Study of the Moulded Nickel Cathode—C. P. Hadley, W. G. Rudy, and A. J. Stoeckert. (*J. Electrochem. Soc.*, vol. 105, pp. 395–398; July, 1958.) "Research work on the moulded nickel cathode is described. Results are given regarding the effects on emission and life of variations in nickel powder, alkaline-earth carbonates, reducing agents, sintering, and aging. Data on pulsed emission are presented."

621.385.032.213.63 2070

On the Energy Distribution of Electrons from Antimony-Caesium Cathodes—A. I. Pyatnitskiĭ. (*Radiotekh. Elektron.*, vol. 2, pp. 714–725; June, 1957.) Results of measurements of the photocurrent and secondary emission of a Cs-Sb cathode and a Cs-Ag layer show the basic difference between their I/V characteristics, confirming the presence in Cs-Sb of low-energy secondary electrons. The number of these electrons is dependent on the quantity of Cs in the cathode.

621.385.3:621.365.5 2071

Intermittent Use of Oscillator Valves in RF Heating Generators—E. G. Dorgelo and J. C. van Warmerdam. (*Electronic Applic.*, vol. 18, pp. 41–47, April, 1958; *Mullard Tech. Commun.*, vol. 4, pp. 173–178; December, 1958.)

621.385.3:621.365.5 2072

Output and Load Resistance of Oscillating Triodes in RF Heating Generators—E. G. Dorgelo. (*Electronic Applic.*, vol. 18, pp. 19–26, January, 1958; *Mullard Tech. Commun.*, vol. 4, pp. 179–185; December, 1958.) A method is described for calculating the output power and other operating conditions of triode oscillators as a function of their load resistance.

621.385.832:621.396.662 2073

Electron-Beam Voltage-Indicator Tube EM84—A. Lieb. (*Elect. Commun.*, vol. 35, pp. 76–82; 1958.) A specially designed tuning indicator with a ZnO phosphor forming two fluorescent bands along the tube axis is described, which can be used in conjunction with color filters or a printed scale to measure voltages to an accuracy within 5–15 per cent.

MISCELLANEOUS

538.569.2.047 2074

Health Hazards from Powerful Radio Transmissions—D. H. Shinn. (*Nature, London*, vol. 182, pp. 1792–1793; December 27, 1958.) Field strength contours in decibels for a paraboloidal antenna in free space are given, and the danger area is calculated on the basis of the theory of Schwan & Li (537 of 1957).

538.569.2.047 2075

Researching Microwave Health Hazards—F. Leary. (*Electronics*, vol. 32, pp. 49–53; February 20, 1959.) A summary of the effects of high-intensity microwave radiation on the human body.

621.38.004.15 2076

Designing for Reliability in Electronic Instrumentation—R. E. Fishbacher. (*J. Electronics Control*. vol. 5, pp. 471–482; November, 1958.) Discussion of the main factors which contribute to the design of reliable electronic instruments.

001.891:621.396 2077

Radio Research 1957: The Report of the Radio Research Board and the Report of the Director of Radio Research—H. M. Stationery Office, London, Eng., publishers, 43 pp., 1958, 3s. 6d. (*Nature, London*, vol. 182, pp. 1558–1559; December 6, 1958.)