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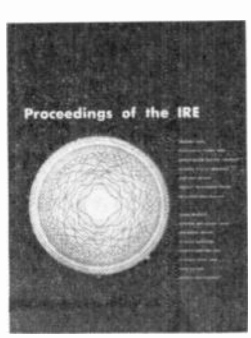
ADVERTISING SECTION



THE COVER—The interesting pattern pictured on the cover is, in reality, the stator of an electrostatic generator of an electronic organ. A number of basic waveforms, each having an harmonic content representative of a class of organ tone, are engraved on a metallized disc. The appropriate waveform is scanned by a rotor to give the organ the desired timbre.

A thorough discussion of electronic music and instruments is presented in the review paper beginning on page 457 of this issue.

Photo—*John Compton Organ Co., London, England.*

Scanning the Issue

Electronic Music (Le Caine, p. 457)—Continuing its series of invited review papers on selected topics, the PROCEEDINGS presents this month an excellent discussion of electronically generated music. The reader will find himself particularly at home with this subject because of the correspondence of pitch, loudness and timbre in music to frequency, amplitude and harmonic spectrum in radio engineering. The discussion brings out the subtle differences between electronically pure music and music as rendered by an artist playing a musical instrument, the various means of achieving these subtle effects in electronic instruments, and some of the unusual instruments and sounds which are now being experimented with.

Transistors versus Vacuum Tubes (Fink, p. 479)—Student member and Fellow alike will enjoy and profit by reading this skillfully drawn comparison of the transistor and the vacuum tube. In a very readable manner the author assesses their present abilities and disabilities and at the same time examines their inherent traits with an eye towards likely avenues of future development. In so doing, he presents an excellent survey of the state of the art for the novice and constructive suggestions for the expert.

The Cryotron—A Superconductive Computer Component (Buck, p. 482)—This paper presents a very novel development which may be of major importance in the future and certainly will be of great immediate interest to the radio engineering profession. The development makes use of the two little-used physical phenomena: first, the resistance of a material gradually decreases as its temperature is lowered until, at a point near absolute zero, the resistance falls off sharply to zero and the material becomes superconductive; secondly, the temperature at which this transition will occur can be lowered by applying a magnetic field to the material, or to look at it another way, if the temperature is held constant the resistance of the material can be switched from its superconductive to normal state by applying a magnetic field. Thus a device, which in its simplest form consists merely of a straight piece of wire about one inch long with a single-layer control winding wound over it, can easily be built which has both power and current gain and which, as this paper shows, can be used in flip-flop and many other computer circuits. Thanks to the recent development of helium liquefiers, the once formidable task of obtaining temperatures near absolute zero is now readily achieved. It is to be hoped that this development, in addition to offering the possibility of a new component to the computer engineer, will open wide the gates to further progress in the virgin field of low-temperature electronics.

Factors Affecting Reliability of Alloy Junction Transistors (Wahl and Kleimack, p. 494)—The transistor has from the beginning offered the promise of being a highly reliable and stable device. These expectations have been borne out as far as the interior of the transistor is concerned but some unexpected difficulties have arisen in regard to surface effects. About three years ago it was found that transistors then in use were unaccountably deteriorating. A careful investigation revealed that water vapor collecting on the surface was the cause. From then on transistors were hermetically sealed under dry room-temperature conditions and the problem was apparently solved. However, it has now been found from severe aging tests that hermetically sealed transistors still show serious deterioration of operating characteristics. In this paper the authors make the important finding that water vapor and oxygen *individually* have rather drastic and generally opposite effects on transistor characteristics which, under normal conditions, counteract each other. However, under prolonged exposure to heat one or the other is driven off the surface leaving a destructive unbalance. It is concluded that for normal applications hermetic sealing is adequate, but for severe requirements involving high temperatures a vacuum baking process, which the authors have found to be adequate, will be necessary.

Microwave Detector (Mendel, p. 503)—A new microwave detector which has bandwidth and power-handling capabilities that greatly exceed those of any present detector has been developed by utilizing an hitherto unwanted instability in an electron beam focusing system developed two years ago for traveling-wave tubes. The focusing system consists of a row of alternately-poled permanent magnets spaced along the electron path, which periodically focus the beam as the electrons travel towards the collector. It has been previously noticed that for certain combinations of magnetic field strength, electron velocity and spacing between magnets, a violent defocusing occurs and electron transmission stops. The author has found that by designing the mechanism to operate on the brink of this "stop-band" when there is no rf modulation on the beam, the introduction of a modulated signal will alter the velocities of the electrons sufficiently so that they fall onto the sloping edge of the stop-band region, thus causing variations in collector current corresponding to the modulating signal. This development not only makes a valuable contribution to microwave detection but also suggests a method for improving the efficiency of traveling-wave tubes and for mixing microwave signals at high power levels.

Increasing the Reliability of Electronic Equipment by the Use of Redundant Circuits (Creveling, p. 509)—The reliability of electronic equipment which employs large numbers of tubes and components is often seriously affected because there are so many parts that can fail. A modern bomber, for example, may employ as many as 10,000 tubes. Since the life expectancy of a tube is at best no more than 10,000 hours, the bomber crew faces the alarming prospect of tubes failing at the rate of one an hour. With little likelihood that tube life will be increased greatly in the future, it would seem on the face of it that as equipment becomes more complex reliability must inevitably decrease. Such is not the case, however, as is shown in this investigation of another avenue of improving reliability which, paradoxically, calls for using even more tubes. By incorporating duplicate circuits in equipment so that they will automatically take over in the event of a failure in the original, it is found that a substantial improvement in reliability can be obtained. This easily-followed discussion is of importance not only for the specific improvements in reliability it suggests but also for its broader implication that the use of more components per se does *not* decrease reliability.

Transformer Miniaturization Using Fluorochemical Liquids and Conduction Techniques (Kilham and Ursch, p. 515)—The techniques presented in this paper make possible a substantial reduction in the size of transformers used in electronic equipment through the use of newly developed coolants which can withstand much higher temperatures than standard transformer oils and yet which have equally good dielectric properties. The increasing need for small components and components capable of giving satisfactory performance at elevated temperatures makes this a

(*Continued on p. 100A*)

Poles and Zeros

Proceedings vs Transactions—II. Last month we developed the proposition that papers selected for publication in the PROCEEDINGS should have significance extending beyond the interests of a particular Professional Group. Papers qualify in this trans-group domain when the subject matter is of common interest to several types of specialists.

For example, a paper on an outstanding new electronic component is presumably of primary interest to the Professional Group on Component Parts, because the members of the PGCP are the designers and producers of said component. But the majority of the *users* of the component, like as not, belong to *other* Professional Groups or to no Group at all. These latter members (customers of the PGCP, so to speak) might languish in ignorance for months, if the PROCEEDINGS studiously avoided publication of papers on components, merely because the PGCP has a first-rate TRANSACTIONS.

This situation arises whenever any new development, arising within the sphere of interest of one Group, can profitably be put to use by members of other Groups. For "new development" we may read new theory, method of measurement, natural phenomenon, computation, magnetic or dielectric material, component, circuit, administrative procedure, system, etc. Such a list covers the major interests of the 24 Professional Groups, but none is the exclusive concern of any one Group.

More often than not, a technical development of *vital* interest to one Professional Group is, or can be if described in appropriate language, of *substantial* interest to a large part of the IRE Membership. Norris Tuttle, a member of the Editorial Board who has given much thought to this matter, puts it this way: "One reason for the continued existence of the PROCEEDINGS as an outstanding technical journal is that so large a part of the technology of radio engineering is the common property of everyone in the field, regardless of his specialization. It is unreasonable to expect an IRE member to join four or five Professional Groups to get material of general interest."

The question of PROCEEDINGS vs TRANSACTIONS publication thus reduces not to subject matter in itself, but to single-group vs multi-group interest. If there is a strong component of inter-group significance in a paper, that paper should be directed to the PROCEEDINGS. To put a paper in its proper place, all those concerned—authors, program committees, PROCEEDINGS reviewers, TRANSACTIONS editors—must weigh carefully the question "How broad a range of interest is represented?" and recommend accordingly.

The channel of referring papers from the PROCEEDINGS to the TRANSACTIONS is well established. The reverse channel should not be neglected. It is in no way a disloyal act for a TRANSACTIONS editor to refuse a top-notch paper in favor of PROCEEDINGS publication. He needs to be satisfied only on two points: (a) that the paper in question is sound and has the broad significance discussed above, and (b) that its language is not so specialized as to prevent its being readily assimilated by a reasonably large fraction of the membership. For more on this last point, the source of complaints from PROCEEDINGS readers of many grades and persuasions, see this space next month.

Access. IRE members who are concerned with the mounting problem of storing past volumes of Institute publications and gaining rapid access to their contents will be interested in the Conference on the Practical Utilization of Recorded Knowledge held last January at Western Reserve University. Many professional groups, notably in law and medicine, have just about given up trying to keep abreast of new developments and are asking for an automatic library system that will classify and store information until needed, and then retrieve it rapidly and accurately.

A good friend of IRE who attended the conference writes as follows: "It really was an eye-opener that such a diversified group is really working on methods to accelerate absorption and to develop an adequate indexing and retrieval system for the vast amount of technical information now being published. . . . About the only people not represented were the electronics and communications people. . . . Most of this is still dream stuff, but it has to come, and very likely it will have to start with the technical societies. For my money, IRE should be a leader in this field, rather than leaving it to the metallurgists and the pharmaceutical groups."

Editors of appropriate Professional Group TRANSACTIONS, please copy. Pieces of this problem would seem to fall in the range of the PG's on instrumentation, telemetry and remote control, information theory, engineering management, electronic computers, production techniques, and automatic control, to mention only seven of our 24 Groups.

Shift. The secretary's report for 1955 contains a noteworthy statistic; the hub of IRE section activity is no longer New York City. The largest IRE Section is that in Los Angeles, by 147 members out of about 4000. Los Angeles also nosed out New York City in the number of meetings held during the year, 24 to 23. Take it away, California!

Donald G. Fink

EDITOR, 1956

Donald G. Fink was born on November 8, 1911, at Englewood, New Jersey. He received the Bachelor's degree in electrical engineering from the Massachusetts Institute of Technology in 1933, and the Master's degree in electrical engineering from Columbia University in 1942.

During 1933–1934 Mr. Fink was a research assistant in the M.I.T. departments of electrical engineering and geology. From 1934 to 1952 he was on the editorial staff of *Electronics*, becoming Editor-in-Chief in November, 1946. Since 1952 he has been with the Philco Corporation as Director of Research for radio, television and appliances.

Mr. Fink participated in the development of the loran system in 1941–1943, when on leave to the Radiation Laboratory at M.I.T. In 1946, he was a civilian consultant on the staff of the Commander, Joint Task Force One, in charge of preparing damage reports on all electronic material and test facilities for the Bikini bomb tests.

He was a member of Panel 1 of the first National Television System Committee and editor of its *Proceedings*. He was also vice-chairman of the second National Television System Committee from 1950 to 1952, and chairman of Panel 12 and its editorial committee.

He is a member of Tau Beta Pi, Sigma Xi and Eta Kappa Mu, and a Fellow of the IRE, American Institute of Electrical Engineers and of the Society of Motion Picture and Television Engineers.

Electronic Music*

HUGH LE CAINE†

The following paper is one of a planned series of invited papers, in which men of recognized standing will review recent developments in, and the present status of, various fields in which noteworthy progress has been made.—*The Editor*

***Summary*—The three attributes of musical sound, pitch, loudness and timbre, are discussed in relation to their counterparts in an electrical signal, frequency, amplitude and harmonic spectrum. Although electrical devices can control these quantities simply according to the explicit instructions contained in written music, that is, to provide the "bare essentials" of a musical performance, it is more difficult to produce the complex patterns of frequency, amplitude and harmonic spectrum actually found in a musical performance on well known instruments. Control of the build up and decay of each note, arbitrary deviations of pitch, vibrato of varying rates and amounts, and a "choir effect" of random pattern of beats are required. Various means of achieving these effects are discussed, first in the electronic organ and then in various "monophonic" or single-note instruments. Finally, coded-performance of "synthetic music" devices are described, including those used by the Musique Concrète Group of Paris, and the Cologne Studio for Electronic Music.**

Introduction

When an electronic engineer looks at music with a view to designing an electronic musical instrument, many things about music strike him as extremely fortunate. The music with which we are familiar may be represented to a good approximation by describing three attributes of the auditory sensation which vary with time: pitch, loudness, and timbre. Each of these variables has a counterpart which is well known in electronics, apart from sound. *Pitch* corresponds to frequency, a property of periodic phenomena; *loudness* to the amplitude of electrical oscillations, and *timbre* to the harmonic spectrum obtained by Fourier analysis. Methods of measuring and controlling frequency, amplitude, and harmonic spectrum are well known.

To begin with, a "musical" sound is usually thought of as a purely periodic one. The sound made by a card held against the teeth of a circular saw while it is being rotated produces a less musical sound than a siren, because in the first case the sound-producing process is less capable of exact repetition from one cycle to the next, than in the case of the second. Electrical generators are capable of giving much more exactly periodic sound than the mechanical generators commonly used in musical instruments, where variations in the surface of the bow, or turbulence in the air in front of the mouthpiece, cause differences between one cycle and the next which can be detected readily by the ear.

The musical scale seems to have been laid out especially for the benefit of the engineer. An octave represents a ratio of exactly 2:1, and the musical pitch scale is logarithmic, as nearly as the musician can tell us, in spite of the relatively large change of several semitones in the pitch (subjective) of pure tones with intensity.[1]

The timbre of a musical sound depends to a good approximation on the amplitude of the harmonics, and not upon their phase. Thus it is easier to synthesize a tone with a given timbre than a wave with a given waveform. Consonance, an offshoot of which is harmony, which is very important in our musical system, depends fundamentally upon agreement in the harmonics of the notes which are sounded together. The integral relations found among harmonics in an harmonic series underlie the frequency relations which determine consonance and frequency ratios found in chords and scales.

Although the accuracy to which frequency must be controlled for musical purposes is rather high—nearly one part in a thousand—electrical generators are for the most part more inherently stable than the mechanical ones used in conventional musical instruments.

The response of the ear to changes in sound intensity resulting from changes in amplitude of electrical oscillations is somewhat complicated, but on the other hand, a rather small number of steps—perhaps about ten—are sufficient to describe the variations encountered in the playing range of most musical instruments. Then, compared to electrical attenuators, the means which are used to control the loudness of the sounds in conventional musical instruments are horrifying, to say the least. In the pipe organ, to get a mere 25 decibels or so change in loudness, many ranks of pipes have been enclosed in brick-lined swell-boxes, provided with hollow, evacuated, aluminum shutters. The violinist spends years perfecting the technique of manipulating the bow in such a way that a pleasant sound will be elicited over a moderate dynamic range; the simplest electrical attenuators change sound intensity without disturbing periodicity and other characteristics of the sound.

The history of the progress of electrical music shows one after another of these fortunate relations between music and electronics being made the basis of a new

* Original manuscript received by the IRE, December 28, 1955.
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[1] S. S. Stevens, "The relation of pitch to intensity," *J. Acoust. Soc Amer.*, vol. 6, pp. 150–154; January, 1935.

musical instrument. Cahill, who in 1895 drew up the first comprehensive scheme for a completely electrical instrument,[2] made use of the integral relations between the various members of the harmonic series by generating all harmonics for one tone with alternators having a common shaft. Although to control loudness he used a vast number of power relays and rheostats controlling many hundred watts, his auditors were amazed at the ease of control of loudness which this apparatus afforded the performer, when compared with conventional instruments. Musicians were impressed by the stability of the tuning; in spite of the fact that when the first public demonstrations were made, the "Telechron" was some twenty years in the future, and means for controlling the frequency of alternators were comparatively undeveloped. In spite of the amplitude distortion which must have occurred in the telephone receivers which he used as loudspeakers, musicians praised the extreme "purity" of the tone.

Some thirty years elapsed before it was suspected that electrical tone might be too pure for musical purposes. The musical value of the ever-changing pattern of beats formed in pipe organ music by the inability of the pipes to stay in tune was not realized until technical knowledge had advanced to the point where instruments capable of staying mathematically in tune could be built, and until these instruments had been studied in a musical setting. Similarly, our present understanding of the musical significance of small changes in pitch, loudness, and timbre produced by the performer on a monophonic instrument was not attained until many instruments providing oversimplified control had been designed, built, and used beside older standard instruments. These subjects will be discussed later.

The first of the modern oscillators was perhaps Duddell's negative resistance oscillator, and he duly noted in his paper of 1900[3] the musical possibilities inherent in an oscillator with an entirely new and flexible method of controlling pitch.[4] After World War I, when vacuum tubes were well established, the heterodyne oscillator provided an even more spectacularly simple and flexible pitch control, which became the basis of the "Theremin," an aggressively futuristic instrument, and the "Ondes Martenot," which was heard in a program of the Boston Symphony Orchestra a few years ago. A glance at the pitch-changing mechanisms of the double-bass and the pedal clarinet shows why electronic oscillators provoked so much interest.

In the "Thirties," after the cone speaker and ac-operated tubes had prepared the way for reliable sound-producing equipment, all the known types of mechanical and electronic oscillators were tried as generators in electrical musical instruments. The review of the generator question at that time was so thorough that no new generators have since been discovered, although the marked improvement in the performance of electronic components has been responsible for an increasing preference for electronic generators, rather than mechanical ones. A great variety of monophonic instruments had been experimented on in the "Twenties" and some had actually reached the stage of being manufactured. However, commercial success was first achieved by the electronic organ, which has increased steadily in economic importance.

The Electronic Organ

The commercial success of the electronic organ is based on the fact that while the individual generators are more expensive on the average than the individual pipes in the pipe organ, each electrical generator may be made to take the place of a number of pipes. Organ builders have had little success in using the same pipes to provide different sound intensities, although a single pipe has been used to sound at different ptiches.[5] An electrical generator might, however, appear on an electronic organ as a "diapason," and the same generator, with its voltage attenuated, might be heard as a "dulciana." With a little more elaboration of circuitry, the generator might be made to do duty for a number of different pipes by using appropriate electrical networks to modify the harmonic structure. Cahill had pointed out that the number of generators required to synthesize an harmonic series for each note in the musical scale could be reduced by taking account of the integral ratio required for consonant intervals which occur in the scale. In the Hammond Organ[6] some ninety generators are used to provide fundamentals and harmonics for nearly as many notes in the musical scale. Individual control of amplitude of each harmonic is provided in a way impossible in an acoustic instrument.

Timbre

The hope of early experimenters that electronic instruments would provide organists with completely new timbres of great beauty has not been realized. In the first place, the ability of the ear to recognize variations in harmonic structures, and thus the number of distinguishable timbres, has been greatly over-rated. We have all noticed that musical instruments can be recognized when heard under circumstances which alter harmonic spectra by an amount which greatly exceeds difference between the spectra of the musical instruments. An interesting lecture demonstration by K. A. MacFadyen[7] shows it is difficult to distinguish between

[2] T. Cahill, U. S. Patent 520,667, applied for in 1895.

[3] W. Duddell, "On rapid variations in the current through the direct-current arc," *J. Inst. Elec. Eng.*, vol. 30, pp. 232–267; 1900–1901.

[4] Negative resistance oscillators were to become very popular with electronic instrument designers. An observer in the early "Thirties" noted that performers on the "Trautonium" carried about with them, and carefully protected, their pet neon lamps, much as a clarinettist does his pet reed.

[5] The "Polyphone," manufactured by the John Compton Organ Co., is a single pipe, tuned by pneumatic motors to cover the lowest octave of a 32-foot pedal stop. It has been found satisfactory to sound the four lowest notes, C to E♭, at E♭ pitch (19.65 cps), since at this pitch a listener cannot detect the difference between the pitch sounded and the true pitch.

[6] L. Hammond, U. S. Patent 1,956,350, applied for in 1934.

[7] K. A. MacFadyen, "Acoustical measurements and the organ" (parts I and II), *The Organ*, vol. 34, pp. 155–162; January, 1955, and pp. 200–203; April, 1955.

the steady-state sound of the trumpet and flute under circumstances where inclusion of the starting transient would leave no doubt whatever as to identity of the instrument. Anyone who has access to a tape recorder, a pair of scissors, and a few musical instruments, can do some interesting experiments on effect of starting period on instrument recognition. Attack and decay processes can be made to change places by reversing tape, and artificial starting periods, produced by cutting tape obliquely through the steady-state period, can be substituted for the natural ones.

In the second place, harmonic synthesis, that most thorough-going method of timbre control, is the peculiar property of the pipe organ. For many centuries individual harmonics up to the eighth have been available as mutation and foundation stops. The sprightly fifth harmonic, heard in Ravel's "Bolero," was undoubtedly suggested by the corresponding mutation stop on the pipe organ sometimes referred to as the "tierce." All the harmonic distributions not found in natural spectra[8] have thus become familiar through pipe organ music.

Up to a point, each harmonic exerts a characteristic effect on timbre. The effect of the octave harmonics—the second, fourth, and eighth—is closely related. When the higher octave harmonics predominate, the "octave" effect is brighter than when the lower octave harmonics predominate. Nonoctave harmonics, such as the third and fifth, also have a related effect: they make tone more strident, assertive, and have effect of emphasizing the pitch of the fundamental; whereas octave harmonics tend to create a confusion in pitch between the fundamental and its octaves. High relative intensities of nonoctave harmonics tend to produce a timbre useful for solo work because of assertive quality and definite pitch, but not too satisfactory for playing chords because of the possibility of the prominent harmonic jumping out of the mixture and being classified by the ear as an independent dissonant note.

It is not surprising that the ear will accept inharmonic partials on almost the same basis as true harmonics, provided no unpleasant beating is set up with adjacent harmonics. The writer has used the 5.33th "harmonic" and the 6.66th "harmonic" in steady organ tones, for example. While in our music, purely periodic sounds are given preference, many accepted sounds such as given by tympani, for instance, have continuous distributions, while other sounds such as those of gongs and chimes have inharmonic partials. Unpleasant effect produced when tones containing many inharmonic partials are used to form chords may be due to high probability of unpleasant beating, but is more likely due to failure of ear to resolve the mixture into the right number of tones[9] having intended pitch.

The writer believes that too much emphasis has been placed on the "pleasant" nature of consonant intervals, whereas they would be better described as "recognizable." The steady progression over the years toward intervals of lower and lower consonance is thus seen as due to the increasing familiarity of composer and audience with the domain of pitch, and the ever-changing rules of harmony are seen as a working indication of how complicated a pattern a given audience can be expected to follow. Meyer-Eppler has stated that the essential nature of an interval is not a consequence of the pattern of beating (which depends upon whether a tone has harmonic or inharmonic partials), but is due to the special ability of the ear to recognize periodicity as a special property of sound phenomena.[10]

In connection with the phenomena of consonance, reference should be made to the very special nature of sound formed by harmonic synthesis from sine wave sources tuned to the equitempered scale—a fact which has received surprisingly little discussion in print. Above a certain point on the keyboard a classical discord is impossible, because the separation of adjacent generators in cycles per second is insufficient to cause a sensation of roughness. In the octave below this point only minor seconds could be disagreeable. While one observer[11] has found a characteristic harmoniousness in the sound, and described the use of tempered harmonics as a basic improvement, another[12] states that the tuning accuracy of sine wave generators for additive synthesis must be of the order of one part in 10,000 or better.

After Helmholtz' demonstration of the relation between harmonic spectrum and musical quality, or timbre,[13] it was assumed that the reason the various orchestral instruments are apprehended musically as very distinct entities was due to the fact that each instrument has a characteristic harmonic spectrum. This mistake was due largely to the difficulty of making harmonic analysis. Meyer and Buchmann[14] published in 1931 an extensive investigation of many musical instruments covering a wide range of frequencies. One of these for the oboe is shown in Fig. 1. These analyses, made with the now familiar heterodyne analyzer, showed that the harmonic spectrum of musical instruments varies widely from one end of the range to the other. Apparatus for taking sound spectrograms rapidly, sometimes known as visible speech apparatus, has recently extended the range of sound phenomena which can be studied conveniently by sound spectrograms.[15] Fig. 2 shows a spectrogram of a violin sound made on a new sound spectrograph developed recently at the Murray Hill Laboratory of the Bell Telephone Lab-

[8] It would, for example, be very difficult to produce in a musical instrument a spectrum in which the fifth harmonic was 20 db more intense than the fourth or sixth, except by synthesis.

[9] The effect of recognizing a chord played with bells or tubular chimes is predominantly one of surprise at having performed a difficult feat.

[10] W. Meyer-Eppler, "The mathematical-acoustical fundamentals of electrical sound composition," (see footnote 58).

[11] S. T. Fisher, "An engineer looks at music," *Engrg. Jour.*, vol. 25, pp. 548–553; October, 1942.

[12] A. Douglas, "The electrical synthesis of musical tones, part 3," *Electronic Engrg.*, vol. 25, pp. 370–373; September, 1953.

[13] The French word "timbre" means "stamp" or "character."

[14] E. Meyer and G. Buchmann, "Die klangspektren der musikinstrumente," *Verlag der Akademie der Wissenschaften*, vol. 32, pp. 3–45; December, 1931.

[15] R. K. Potter, G. A. Kopp, and H. C. Green, "Visible Speech," D. Van Nostrand Co., New York, N. Y., 1947.

oratories. A glide extending over a little less than an octave is followed by a steady tone to which vibrato is added gradually. The vibrato at the end is scarcely noticeable on the fundamental, but becomes more so in the higher harmonics because of the linear frequency scale. A separate spectrogram gives a frequency calibration. The time scale is about four inches per second.

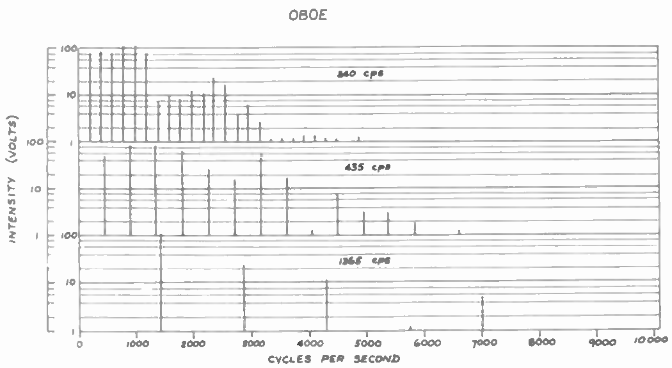

Fig. 1—Spectral analysis of three notes on the oboe obtained by Meyer and Buchmann, 1931. (After Meyer and Buchmann.)

When the spectra for tones of various pitch are looked at on the same frequency scale it is seen that there is a common element at work emphasizing certain frequencies no matter what the pitch of the note sounded may be. The frequency-selective elements are known as "formants." Trautwein[16] was the first to use in an electronic musical instrument an electrical analog for the formants which occur in mechanical musical instruments. Thus if there is a common quality among the timbres of the different tones emitted at different pitches by the same instrument, it may be due partly

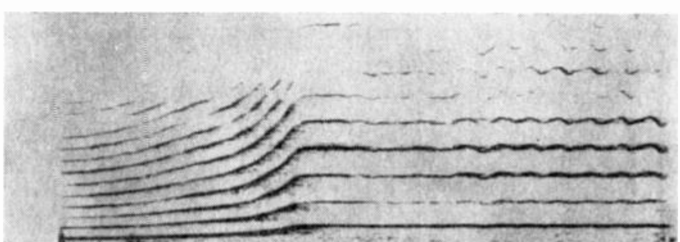

Fig. 2—Spectrogram of a violin sound made on a new spectrograph developed recently at the Murray Hill Lab. of the Bell Telephone Labs. (Spectrogram supplied through courtesy of the Acoustic Research Dept. Bell Telephone Labs.)

to the constancy of the formant position. The ability of the ear to recognize a certain pattern of formant behavior in sounds of widely-differing pitch is shown by the ease with which vowel sounds can be recognized whether sung by a bass voice or a soprano voice. At the same time, as anyone knows who has attempted to imitate orchestral instruments with electrical generators, the most distinctive characteristic of the clarinet is the suppression of the second and fourth harmonics, even though this occurs in the lower register only. Martin[17] points out how little the sound of a flute resembles a laboratory oscillator, even though the harmonic analysis of some flute tones reveals little but fundamental. We are led to conclude that the physics of "familiar" sounds is very complex.

Towards the end of the "Thirties" it was realized[18] that the basic problem in designing timbre-control circuits was to find some means of providing a fairly-detailed control of the distinctive lower harmonics, such as the second, third, fourth, and fifth, while exerting only a modifying influence on a large group of harmonics extending up perhaps to the thirtieth. The difficulty with harmonic synthesis from sine waves is that if enough harmonics are provided, the system is too expensive and provides much more control over the higher harmonics than is required. The formant method, while providing an admirable solution for control of timbre on an electronic instrument patterned after the orchestral instruments, suffers from a limitation when applied to the design of the organ, due to the fact that the pipe organ covers a much greater range than orchestral instruments. An examination of the harmonic spectra of the pipes on the pipe organ which imitate orchestral instruments such as the oboe, shows the presence of a "variable formant," since the resonant structure is different for each pipe. The dependency of the spectrum of organ pipes on pitch is thus midway between the constancy with pitch obtained by harmonic synthesis and the extreme variation found in orchestral instruments. The consequence of this fact for electric organ designers is that formant circuits used over a whole manual must be restricted in their sharpness in order to keep the loudness all over the keyboard within acceptable limits.

While there is thus little hope of finding a practical way of imitating exactly orchestral or even pipe organ sounds, it seems reasonable to expect that the designer of electronic organs should be able to arrive at new solutions to the much more important problem of providing musically useful sounds. Harrison, in "The King of Instruments" series of records,[19] remarks that the test of a really musical sound is whether any note in the chord being held will be missed if dropped out, and heard reentering. An excellent demonstration is given on this record of the unsuitability of the "phonon diapason" (a pipe favored by romantic organ-builders) for polyphonic music. MacFadyen[7] has shown that on the basis of what is known about masking, a soft-toned stop—one having low harmonic development—must have an upward tilt in the loudness characteristics toward the top of the manual in order that the low notes will not make the high notes inaudible through masking. When the harmonic spectrum is more complex, he suggests it is possible to obtain the required clarity of tone with a constant loudness characteristic over the whole manual by emphasizing the higher harmonics of the low notes.

[16] F. Trautwein, "Elektrische Musik," Verlag Weidman, Berlin, Ger., 1930.

[17] D. W. Martin, "Thoughts on the imitation of natural sounds," *J. Acoust. Soc. Amer.*, vol. 25, p. 158; January, 1953.

[18] W. F. Curtis, U. S. Patent 2,227,068, applied for in 1939.

[19] "The American Classic Organ," Record I of "The King of Instruments," Aeolian-Skinner Organ Co., Boston, Mass.

The tendency in present-day electronic organ design seems to be to combine the three possible methods of varying timbre:

1) provide a number of basic waveforms for the same pitch,
2) modify the basic waveforms by electrical networks common to the whole manual, and
3) provide mutation stops derived from the main generators, plus as many octaves as possible.

In the Baldwin Organ, for instance, two different basic waveforms, the square-wave and the sawtooth, are obtained from electronic dividers. Formant circuits operating on these two basic waveforms provide a wide variety of timbres which are available in a number of registers.[20] The Robb Wave Organ, which was manufactured in Canada during the "Thirties," used a number of basic waveforms which were combined in various proportions to form the timbre resources of the instrument. The basic waveforms were generated by rotating steel cylinders carrying milled contours in front of the magnetic pick-up coils. The twelve cylinders may be seen in Fig. 3. The stators carrying the pick-up coils may be seen in the front row, while in the back row, the steel cylinders carrying the contours are visible in front of the stators.

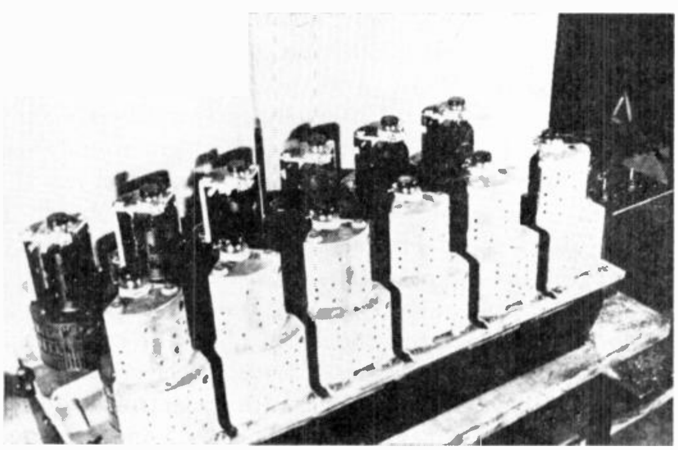

Fig. 3—The generators of the Robb Wave Organ, an electronic organ manufactured in Canada in the "Thirties." (Photograph courtesy of Morse Robb, Montreal, Can.)

In the Compton "Electrone" a number of basic waveforms, each having an harmonic spectrum representative of a class of organ tone, are engraved upon a disc which forms the stator of an electrostatic generator.[21] Fig. 4(a) is a photograph of the disc which carries these waveforms. The waveforms are scanned by a rotor shown in Fig. 4(b). In this design it is economically feasible to provide a variety of attacks.

Speech, Clicks, and Attack

Electronic organs differ from pipe organs in the manner in which the tones start and stop. The ear is conditioned to pay much more attention to the former than to the latter, since the rhythms in music depend almost entirely on the starting of the tones. The essential character of a melody is the same, played either staccato or legato. The cessation of a tone is, moreover, usually disguised by room reverberation. Thus the decay of a note attracts the listener's attention only if it is very rapid and begins as soon as the note has reached peak intensity, as with some percussion instruments.

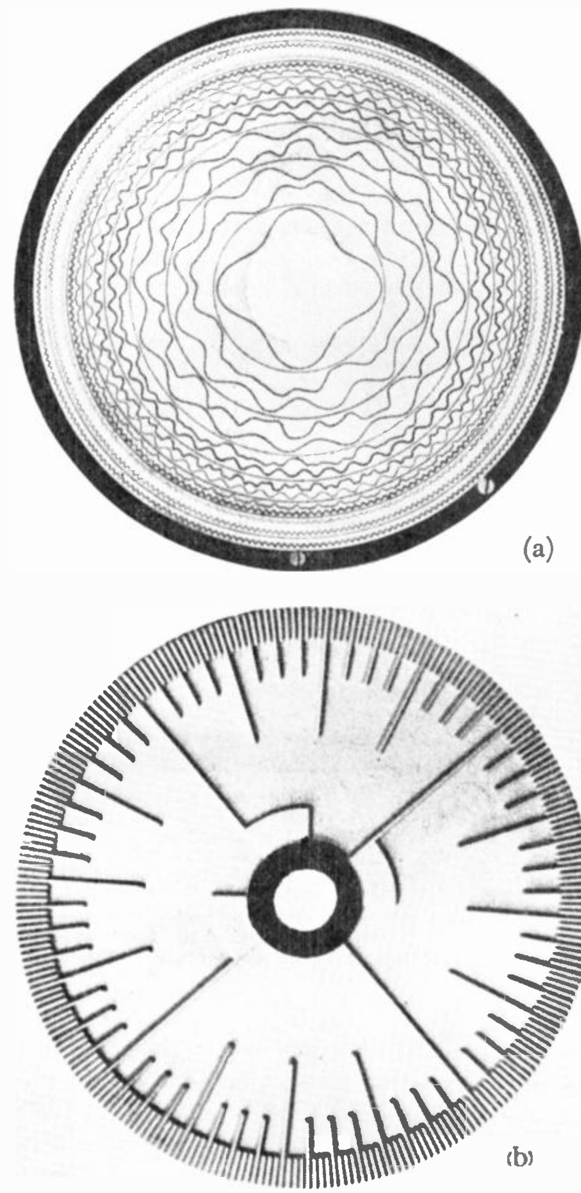

Fig. 4—(a) Stator of an electrostatic generator in the John Compton Organ Co.'s "Electrone"; the waveforms are engraved upon a metallised insulating disc, (b) Rotor of the electrostatic generator. (Photographs courtesy of the John Compton Organ Co., London, Eng.)

It is characteristic of electronic methods of tone formation that very rapid changes in intensity or frequency are possible, and may indeed be troublesome to avoid. A simple tone which is suddenly initiated at time $T=0$, then continues for something like a tenth of a second with undiminished intensity, is shown in Fig. 5(a). Although the manner in which the tone is produced

[20] R. H. Dorf, "Electronic Musical Instruments," Audio Library No. 1, Radio Magazines Inc., Mineola, N. Y., pp. 46–49, 1954.
[21] *Ibid.*, pp. 173–177.

suggests the spectrogram shown in Fig. 5(b), it nevertheless has the spectrum shown in Fig. 5(c). The sudden start and stop of the oscillation brings about a considerable widening in the time-frequency spectrum, which can be heard as a "click." The ear requires a minimum of about 13 milliseconds (depending upon the frequency) to recognize a sound as having a definite pitch. The effect of the broadening of the spectrum, shown in Fig. 5, is to distract the attention from the pitch of the notes being played, and is therefore undesirable. Since the noise occurs at the beginning of the notes, just when the attention of the ear is most required to perceive the melodic pattern, the effect is probably considerably worse than that of a uniform background noise of the same average energy. Where the design of the organ makes a rectangular envelope unavoidable, a noticeable improvement may be obtained by the use of a filter, if the filter is sufficiently narrow. For this purpose, a recent organ design has provided filters grouped in octaves.[22]

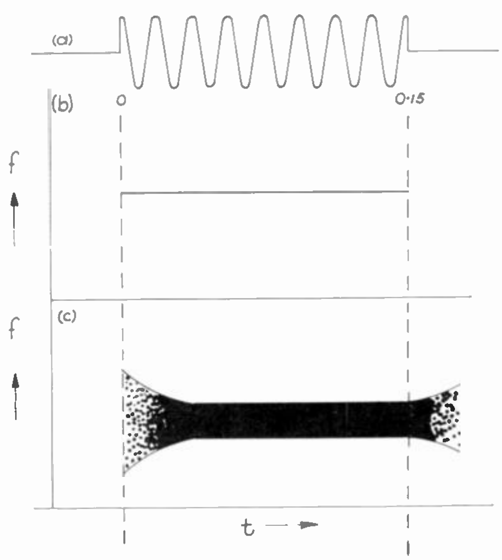

Fig. 5—(a) Oscillogram of a short sinusoidal tone, stopped and started suddenly, (b) Representation of the physical process and apparent spectrogram, (c) True spectrogram (simplified). (After Meyer-Eppler.)

There is great variety in the way the tones of organ pipes and conventional instruments start. The lecture demonstration mentioned earlier[7] shows that the way a tone begins in a trumpet or a flute is more characteristic of the instrument than the harmonic spectrum of the steady-state tone. When the build-up is very rapid, the effect of the build-up on the listener is welded very closely to the other attributes of the tone. It is, in fact, often confused with timbre. If the build-up time is much faster than one-hundredth of a second, the listener begins to perceive the enlargement of the spectrum as a "click." If the build-up time is much slower than one-tenth of a second, the listener perceives a tone steadily increasing in loudness. There is usually a significant change in harmonic spectrum during the build-up of the tone. There is sometimes a characteristic waver in pitch, or a frequency modulation by noise voltage, which is especially prominent during build-up. Sometimes narrow-band noise is present during the build-up, in strengths comparable in intensity to the musical part of the tone. All the foregoing factors determine the "speech" of an organ pipe or a musical instrument. They play a very important part in giving character to the instrument and in making its voice recognizable when heard with other voices in an ensemble. Although marked idiosyncracies of speech give a pipe the pronounced individuality required for solo work, they do not help it to satisfy Harrison's definition, given above, of a musical sound suitable for use in polyphonic organ music, since the peculiarities of speech do not prevent the tone being lost when sustained in an inner part. Thus polyphonic organ music is not always successful when played on the harpsichord, for example, which underlines the entrance of every new note so forcefully that a sustained note is lost long before the intensity has decayed appreciably. In addition to the processes which are closely connected to the pitch of a note, and are therefore useful in drawing attention to a new tone and giving it individuality, conventional musical instruments sometimes produce noises which are unrelated to the pitch, and are therefore as reprehensible as the clicks produced by electronic instruments having rectangular envelopes. Only those deeply attached to the piano in its present form would consider thump obtained on high notes a musical asset.

While it is not impossible to duplicate the speech of organ pipes by electronic means, the physical complexity of the organ sounds makes the apparatus too expensive. Some organ designs, notably those using electrostatic generators, however, allow the economical inclusion of several attacks. Bourne's electronic pipe organ auxiliary, based on a rotating electrostatic generator, was introduced in the "Thirties," and provided the performer with extensive envelope control, including many percussive effects.[23] The post-war version of this instrument is the Compton "Electrone"[21] which has been mentioned in an earlier section. Another approach to the problem of providing a variety of attacks is through the touch-sensitive action, which is discussed in a later section.

Choir Effect, Tremolo, and Vibrato

Although one of the advantages of electronic organs is the fact that one generator may do duty in many places, this economy results in a simplification of the sound texture due to the elimination of beats between mistuned generators. The true "choir effect" obtained

[22] G. H. Hadden, "Some considerations regarding volume production of electronic musical instruments," *J. Audio Engrg. Soc.*, vol. 1, pp. 29–36; January, 1953.

[23] "Electrical organ tones," editorial, *Wireless World*, vol. 36, pp. 514–515; May, 1935.

in a large pipe organ makes it possible for the full organ sound to be extremely bright without sounding "hard" or "screaming," and gives it instead a scintillating quality. The very satisfying feeling produced by the choir effect is undoubtedly connected with the inability of the ear to function properly under continuous stimulation.

The choir effect is vaguely related to the tremolo, an amplitude modulation at 5 or 6 cycles, and to the vibrato, a frequency modulation at 6 or 7 c. The universal adoption of intelligently applied vibrato has been responsible for a good deal of the improvement in the sound of the electronic organ which has taken place since the "Thirties." Thanks to the efforts of engineers during that period, we now hear rectangular envelopes disguised by synthetic reverberation, and the hard quality of sustained tones which is caused by their excessive purity—that is, constancy of frequency and amplitude of the electronic tones—is blurred by the application of vibrato. It is also interesting to recall how little the popular appeal of vibrato on organ tones was appreciated in the "Thirties." As far as the writer knows, Bourne's electronic pipe organ auxiliary[23] was the first commercially available organ to be provided with a true vibrato. When we recall how difficult it is to drive rotary equipment without periodic variations in driving speed, it is all the more surprising that the organs of the "Thirties" using rotating generators (with the one exception noted) did not provide vibrato effects.

Vibrato is a very effective device when correctly applied, the musical usefulness of which far exceeds the simple function of disguising the excessive purity of electronic tones. Vibrato is related to biological tremors, and when used by a player of an expressive instrument, is clearly connected with the fluctuations of the emotional temperature of the music. An extensive analysis of the vibrato has been made by Seashore[24,25] at the University of Iowa. A novel method for the production of vibrato based on the Doppler effect in an electrical line was patented by Hanert[26] in 1943, and is now used in the Hammond Organ.

A mechanical vibrato which is not under the direct control of the player—that is, which can only be turned "on" or "off"—while tolerable in background music, cannot be used more than a small percentage of the time in music which is to occupy the attention of a serious audience. The reason for the unmusical effect produced by prolonged use of a mechanical vibrato seems to lie in the fact that, unlike the choir effect which modifies the hard effect of pure tones without making the listener aware of a periodicity, the mechanical vibrato has a tyrannical and overriding effect which forces itself on the attention of the listener and breaks up the delicate and more rapidly moving patterns of the music. Although players of expressive instruments use vibrato a large part of the time, their vibrato is carefully related to the pattern of the music. The very mechanics of the application of vibrato to stringed instruments materially assists the player in this regard.

Unfortunately, to produce a successful choir effect, a high degree of complication and unpredictability is essential. The choir effect achieved by including just one extra set of generators is unsatisfactory, since after some minutes have elapsed the listener begins to perceive and then be distracted by the regularly recurring periodicities which are produced. The widespread use of electronic generators increases the complexity of the frequency and amplitude variations which may be introduced economically. The application of frequency modulation by a low-frequency noise source to the top oscillator of a divider chain can be carried out quite simply, and adds a pleasant "waver," but is useful to a rather limited degree in alleviating the "hardness" of tones of constant pitch and amplitude.

Several commercial instruments have provided an extra "vibrato" running at the beating rate to which the "unda maris" and "celeste" stops are usually tuned; that is, between one and two cps. When this is used together with the normal vibrato, an effect slightly suggestive of the choir effect is obtained. A novel musical effect was suggested by Hanert.[27] In this arrangement the beating rate between the auxiliary generators and the main generators varied from 6 to 8 c at 400 c, to 9 to 13 c at 2,400 c. The high beating rate at the low frequencies produced what was thought to be the maximum usable choir effect in bright tones.

Another arrangement suggested by Hanert[28] involves three oscillators for each note, each of which could be frequency modulated. Slightly different frequencies such as 5, 6, and 7 cps were used, the beat rates being thus confused.

The choir effect now used in the Hammond Organ, also due to Hanert[29] while not as musically satisfactory as the two just described, is much easier to manufacture in conjunction with the Doppler-effect vibrato.[26] In this scheme some of the unaltered signal is combined with the signal to which vibrato has been added. Addition of the unmodulated signal produces an amplitude modulation in addition to the phase modulation. The pattern of beats is complicated by the fact that the phase shift between the signals which are combined depends upon frequency, hence the nature of the amplitude modulation is different for each harmonic. Thus the periodicity is less obtrusive and less objectionable than in the case of a simple vibrato. Since for any note or harmonic the cycle repeats at the scanning frequency, the effect of a periodicity is by no means entirely absent, and the result is not as satisfactory as a true choir ef-

[24] C. E. Seashore, "The Vibrato," Studies in the Psychology of Music Series, vol. 1, The University of Iowa, Iowa City, Iowa; 1932.

[25] C. E. Seashore, "Psychology of the Vibrato in Voice and Instrument," Studies in the Psychology of Music Series, vol. 3, The University of Iowa, Iowa City, Iowa; 1936.

[26] J. M. Hanert, U. S. Patent, 2,382,413, applied for in 1943.

[27] J. M. Hanert, U. S. Patent 2,498,367, applied for in 1944.

[28] J. M. Hanert, U. S. Patent 2,500,820, applied for in 1945.

[29] J. M. Hanert, U. S. Patent 2,509,923, applied for in 1946.

fect. The problem of the electrical simulation of the choir effect obtained in a pipe organ is thus essentially an economic one, and an entirely satisfactory solution has not yet been obtained.

Reverberation

The writer believes that tones which end rather suddenly, even though no undesirable click can be noticed, are less useful in music than those which die away slowly, since the slow decay has little effect on the ease with which intricate and rapidly moving rhythms can be detected, and it assists the listener in fixing the pitch of the note. The most suitable decay time for well known music is probably that found for optimum reverberation time, that is, about 1.6 seconds. However this may be, the long reverberation which is characteristic of the structures in which pipe organs are usually heard, is felt to be an essential part of organ music. Builders of pipe organs have used special reverberant chambers with good effect when building conditions were unfavorable. Much ingenuity has been shown by designers of electric organs since the "Thirties" in providing synthetic reverberation effects which are suited to organ music.[30] In a recently developed type[31] reverberation coils of steel wire are coupled mechanically to a cone which is smaller than the loudspeaker cone and mounted concentrically in front of it. Vibrational energy absorbed by the reverberation coils is transmitted to the auxiliary cone and radiated as reverberant sound at a later time.

The Touch-Sensitive Organ

The electronic organ makes possible a touch-sensitive action which is virtually impossible with the pipe organ. The removal from the performer of intimate control of the loudness of the sound produced by the pipe organ is the result of its mechanical complexity. A glance at the structure of the most common types of music suggests that the loss of control of the loudness is a very serious one. The rapidity with which the piano replaced the harpsichord some 250 years ago confirms this impression. In fact, if we were unacquainted with the artistic excellence of preromantic organ music, it would be difficult to believe that a performer could produce effective music simply by timing accurately the entrance and exit of the sounds.

When considering a touch-sensitive action for the organ, one immediately wonders whether or not the organist could exert a satisfactory control over the loudness of each of the notes. The attention which the performer brings to bear over controlling the loudness is, of necessity, diverted from the control of the entrance and exit of the note, which therefore must become less precise. It also becomes possible for a performer to turn out a bad performance because some of the notes are incorrectly accented; whereas, with the standard organ, incorrect timing is the only possible fault. A similar comparison might be made between the piano and the harpsichord. In a piano transcription of harpsichord music, ornaments, and rapid runs lose some of the crispness and beautiful clarity which they would have on a harpsichord. It seems likely, however, that once the touch-sensitive keyboard has been mastered, the texture of contrapuntal passages and the independence of the parts will be considerably enhanced by the ability to accent certain notes slightly, and to obtain a crescendo or decrescendo of parts independently. The ability of the performer to control the attack and decay of each note independently gives rise to interesting possibilities for maintaining the independent character of parts. In imitating the stops of the baroque organ, a considerably more satisfactory effect is obtained with a given timbre when a touch-sensitive keyboard is used in place of the conventional kind. The direct control of the air valves used on baroque organs led to a certain degree of touch-sensitivity, although it was not very satisfactory.

The touch-sensitive issue may then be stated as follows: a fine structure in the loudness pattern and individuality of parts may be obtained with the touch-sensitive action, while the "expression" pedal is more suited to production of general fluctuations in loudness covering a wide dynamic range, and accompanied by the inevitable merging of the individual parts into the general mass of sound.

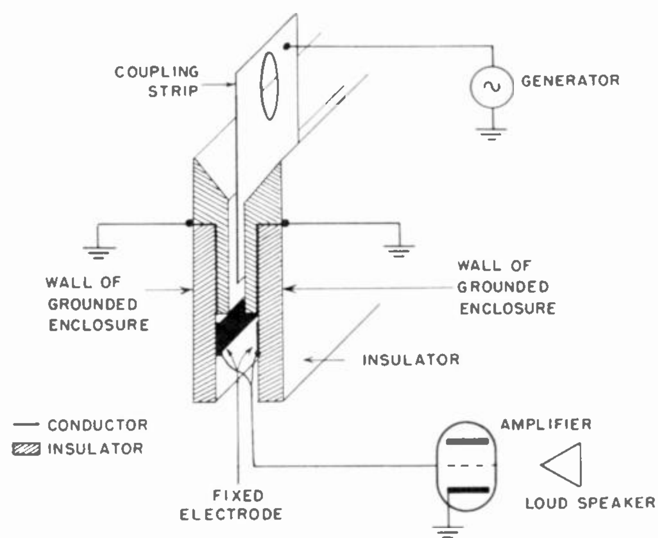

Fig. 6—Electrostatic coupling device suitable for use in a touch-sensitive organ keyboard (from *J. Acoust. Soc. Amer.*).

While there are many ways of realizing touch-sensitivity in an electronic organ, the electrostatic coupling device shown in Fig. 6 is notable for its low cost and simplicity. The attenuation vs displacement curve (the zero of the attenuation curve is arbitrary) is shown in Fig. 7. To obtain the rapid attenuation rate

[30] L. Hammond, U. S. Patent 2,230,826, applied for in 1939.
[31] D. W. Martin and A. F. Knoblaugh, "Loudspeaker accessory for the production of reverberant sound," *J. Acoust. Soc. Amer.*, vol. 26, pp. 676–678; September, 1954.

with a reasonably wide spacing of grounded surfaces, it is necessary to use a grounded channel as shown in Fig. 6, rather than a slit in a thin grounded shield. The depth of the channel must be sufficiently great that the required attenuation is obtained before the coupling strip leaves the channel. The device is described in a recent issue of the *Journal of the Acoustical Society of America*.[32]

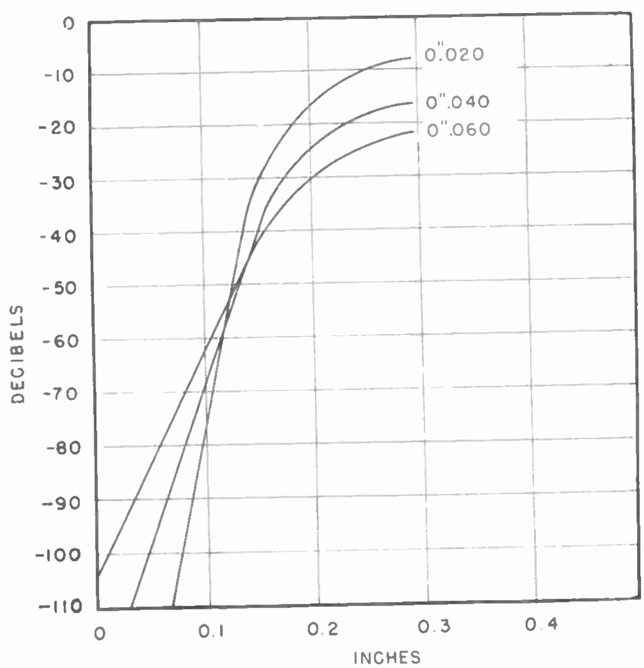

Fig. 7—Coupling in decibels versus position in inches of the coupling strip for various widths of grounded slot, obtained for the coupling device shown in Fig. 6 (from *J. Acoust. Soc. Amer.*).

From five to ten of these are normally used on each key for coupling harmonically related generators. When the highest harmonic and the dissonant harmonic coupling strips are adjusted to be a short distance farther out of the slot than the remaining coupling strips, a variation of timbre between the loud and soft sounds on the same manual may be produced. The difference between the various attacks which may be produced by the manner of operating the key is also accentuated, the slow attacks having an almost vocal quality due to the pronounced shift in timbre produced by the building up of the dissonant harmonics. An additional element of variety is introduced by the fact that the nature of these effects depends upon the timbre in use.

The writer's experience in playing a touch-sensitive organ has so far been limited to a single manual model which was built at the National Research Council of Canada in 1954. Oscillograms of two attacks produced on this model are shown in Fig. 8. A model now being constructed is designed to make the best possible use of piano-playing technique. The key linkage has been modified in favor of the production of fast attacks and comparatively slow decays. A holding pedal has been provided.

The Monophonic Instrument

Careful listening to a single musical part discloses that the performer on a standard monophonic instrument such as the saxophone, cello, or voice, has continuous and detailed control of the three musical parameters: pitch, loudness, and timbre.

While the ear can recognize several thousand distinct pitches, less than one hundred can be written on a musical score, and the development of significant musical ideas takes place in a considerably smaller range than this. Although our musical notation is not capable of providing the performer with more than a bare indication of the composer's desire to depart from the set of preferred pitches, players of expressive instruments do regularly make these departures and they constitute an important part of the effect produced by a monophonic instrument. The essential continuity of the pitch line in a monophonic instrument is an important point of departure from the polyphonic instrument of the organ type, and certain types of pitch flexibility are impossible on such a polyphonic instrument, the basic feature of which is the separation of each musical part into discrete tones which are supplied from separate generators. The monophonic instrument is thus the most important musical instrument, and the starting point of all musical thinking; and the polyphonic or organ-like instrument is simply an expedient necessary in view of the difficulty of assembling the required number of monophonic instruments and performers.

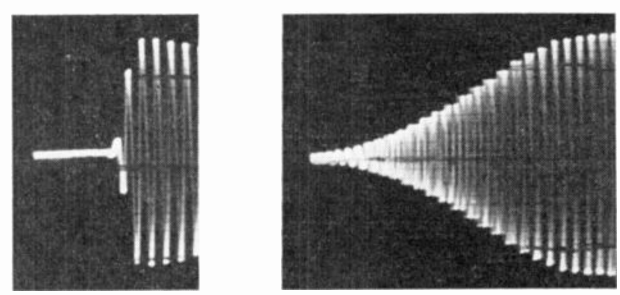

Fig. 8—Oscillograms of two attacks produced by the manner in which the performer operates the key of a touch-sensitive organ (from *J. Acoust. Soc. Amer.*).

Fig. 9 is a graph of pitch vs time showing how a performer on an expressive instrument might approach and perform a sustained note. The section shown at 1 is a slide or a portamento; the intensity of the note, which is not shown, rises from a low value so that the listener gradually becomes aware of the slide as the true or written pitch of the note is approached. Part of the effect of such a slide on the listener is one of smoothness. When the slide is prolonged, however, the effect produced is not related to that produced by a stationary note. The section shown at 2 is an off-pitch

[32] H. Le Caine, "Touch-sensitive organ based on an electrostatic coupling device," *J. Acout. Soc. Amer.*, vol. 27, pp. 781–786; July, 1955.

note which is purposely performed, perhaps about one-third of a semitone below the written or true pitch, and produces in the listener a feeling of unrest, possibly because the true pitch is anticipated. The section shown at 3 is performed on true pitch, but without vibrato. Finally, in the section shown at 4 a vibrato is added gradually, modifying the shrill effect of the section without vibrato. Thus the performance of the note comes to a close. While the listener is not usually aware of these effects as such, he would describe the above performance as having "warmth," or being "alive"; while a note begun and ended on the written pitch, situated among notes begun and ended in similar fashion, all having a constant vibrato, would be described as "mechanical."

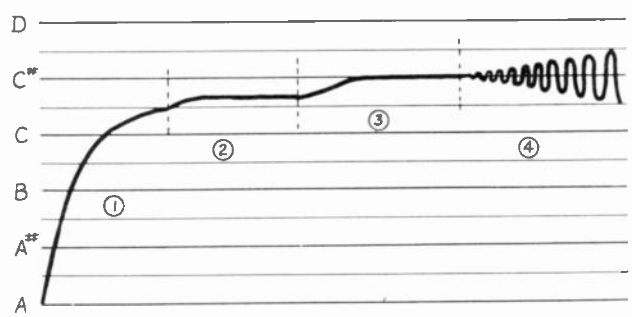

Fig. 9—Graph of pitch vs time, showing how the performer on an expressive monophonic instrument might approach and perform a sustained note.

The variation in intensity of the note which constitutes the attack and decay has already been discussed. The manner in which the intensity of the sound builds up in a musical instrument, together with the other processes which occur at the beginning of a note, are characteristic of the various musical instruments. The attack and decay of most of the expressive instruments can, however, be modified by the performer, although the nature of such control is far from ideal. In addition to these changes of intensity, there is a period during a sustained note when the intensity may be increased or decreased at the will of the performer, the time constant of such a process being considerably greater than one-tenth of a second.

A performer on some conventional instruments, such as the human voice, can vary the harmonic spectrum independently of frequency and intensity. Some instruments, such as the cello, have only a small range of variation: in others, such as the saxophone, the harmonic spectrum is closely related to the intensity. In all expressive instruments, including the concert harmonica, the variation of harmonic spectrum, even though it may in some cases be rather small, plays an important role in setting up the sound patterns with which we are familiar. This fact may be appreciated by noting the difference in effect produced between a line of notes played on the expressive instrument in question, and the same line of notes played on an electronic organ, even though the average harmonic spectrum may be imitated quite closely. In some cases, notably the human voice, the ability of the instrument to change its harmonic spectrum clearly enlarges its expressive power. In other cases, it may be only the lack of the accustomed diversity of effect in the electronic counterpart which causes the listener to raise the derisive cry, "electronic tone!"

In addition to being able to vary the sound in the ways which have just been indicated, our musical system demands that the performer be capable of executing notes at a top speed of from ten to fifteen notes per second. Naturally the sound structure at these high speeds is not required to have the complicated form which has been outlined, and is extremely simple, one of the chief prescriptions being against the introduction of spurious frequencies due to undesirably sudden changes of pitch or amplitude—a way in which electronic instruments are likely to offend.

The early experimenters on electronic monophonic instruments made daring departures from conventional control means. In the "Theremin" the player controlled the sound without making any physical contact with the instrument. By this arrangement the inventor proposed to avoid the "resistance" to being played which conventional instruments offer. A study of the instruments of the "Twenties" suggests that every conceivable means of controlling pitch and intensity was tried.[33] Leading composers wrote for the new instruments, and a well-known conductor predicted in 1929 that within a few years the conventional instruments, with their limitations and unnecessary playing difficulties, would have disappeared completely. Several monophonic instruments reached the production stage in Europe, and the "Theremin" was produced in the United States in the early "Thirties." The commercial success of these instruments was limited by the unsatisfactory state of electronic devices (tubes with ac-operated filaments were still a novelty). The inventors, in their eagerness to explore new musical effects, gave little thought to the prevention of clicks and other unpleasant sounds, or to making their instruments playable. When the intellectual exuberance of the "Twenties" gave place to the more conservative atmosphere of the "Thirties," development of the monophonic instrument was concentrated on the problem of producing an instrument which would stay in tune, sound pleasant, and be easy to play. The commercial instruments of the "Forties" succeeded admirably in these regards; nevertheless, the important orchestral combinations, both large and small, will still be found to consist almost exclusively of instruments invented more than seventy-five years ago, and we are forced to conclude that the problem of producing an electronic monophonic instrument to supplement or replace the conventional monophonic instruments is still unsolved.

[33] "Electronic Musical Instruments, A Bibliography," 2nd ed., Tottenham Public Libraries and Museum, London, Eng.; 1952.

The Electronic "Sackbut"[34]

In beginning work on a monophonic instrument in 1945, the writer started with the idea that the three coordinates of space should correspond to the three musical attributes: pitch, loudness, and timbre. The vertical direction was to correspond to loudness by analogy to the touch-sensitive keyboard instrument; the horizontal direction was to correspond to pitch by analogy to the arrangement of the musical scale on the keys of keyboard instruments, and the remaining direction was to correspond to one chosen subparameter of timbre. Instead of using as a control means a motion in the given direction, the control surface was made very stiff, and the performer concentrated his attention upon exerting appropriate components of force in the three directions. A number of control surfaces were provided; these control surfaces, in fact, constituting the keys of a standard keyboard. Thus it was intended that continuous variations of pitch, loudness, and timbre should be set up by a pattern of varying forces applied by the performer's finger on an individual key, while transformations in the pitch coordinate, in steps corresponding to the musical scale, would be made when the performer moved to a new key.

Much the most interesting of these control means is the association of continuously variable pitch with horizontal force applied to the key. The first record of its application is contained in a patent by Martenot.[35] Here a small horizontal displacement of the key varies the capacity in one of the radio-frequency oscillators of a heterodyne audio oscillator. When applied in this manner, the range of continuous variation of oscillator frequency must be kept small, otherwise playing becomes difficult at the low end of the keyboard. In order to use this control for producing a pitch variations of the type shown in Fig. 9, it is necessary first to arrange the pitch control so that the application of the same force or displacement on any key causes the same pitch change. The operation may then be greatly improved by making the relation between pitch and horizontal force as shown in Fig. 10. The sensitivity to horizontal force is seen to increase as one gets farther from the normal pitch associated with the key. In the normal playing range, that is, about half a semitone on either side of the standard pitch, the sensitivity should not be greater than 0.02 semitones per ounce, in order to permit rapid operation of the keys. The hoirzontal force need not then be controlled to better than one or two ounces except on long sustained notes, and an average vibrato requires an oscillating force of about eight ounces. About half a semitone on either side of standard pitch the sensitivity should increase to a value sufficiently great to reach the maximum pitch excursion at less than five pounds force. When the range is an octave on either side of standard pitch, the sensitivity must be increased about ten times, or to 0.2 semitone per ounce. With this relation between pitch and horizontal force, the performer feels that the sensitivity is approximately constant, as long as he confines himself to the type of pitch variation found in conventional music, since the wide deviations from standard pitch are done rapidly and usually at low intensity, and little accuracy is required. When applying large horizontal forces, it is helpful to depress keys below the key being sounded. A slip-proof surface on the key is also useful.

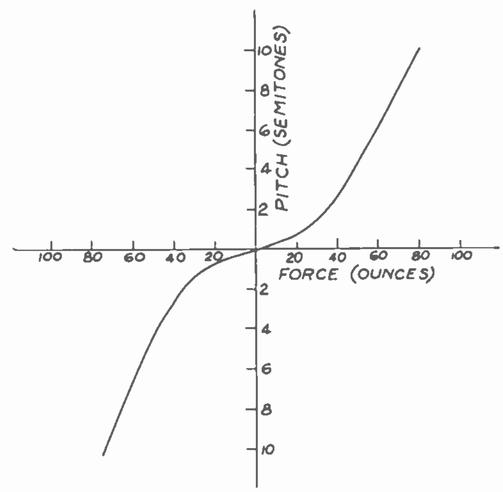

Fig. 10—Desired relation between pitch and horizontal force on the key in the electronic "sackbut."

To start a note off-pitch and gradually bring it toward the correct pitch, it is a simple matter to approach the keyboard at a slight angle to the vertical, and, during the production of the note, gradually remove the horizontal component of force. A portamento is done in much the same manner as that used by the player of a stringed instrument. The key being left is first forced in the direction of the new key; while the force is increasing, the key is dropped and the new key is depressed with no horizontal force applied. Thus the pitch moves in the direction of the new note, transition becoming ever more rapid, so that the listener is aware principally of the smoothness of the transition. For a satisfactory portamento the keyboard time constant must be very short; that is, the springs which return the keyboard to its normal position must be as stiff as possible. As constructed, the maximum horizontal displacement of the keyboard was about 0.001 of an inch, and the return-spring stiffness was 5,000 pounds per inch. It will be seen from the circuit diagram, Fig. 11, that a smoothing filter is also included. This helps to avoid irregular variations of pitch which are not intended by the performer, but which are due to improper execution of the movements described. Needless to say, the constants of the filter are such that no perceptible glide effect is caused by the filter itself. Vibrato

[34] The original sackbut is a thoroughly obsolete instrument. This choice of name was thought to afford the designer a certain measure of immunity from criticism.

[35] M. Martenot, U. S. Patent 1,853,630, applied for in 1930. (French patent application filed in 1928.)

is produced directly by the performer. The limitations inherent in mechanically-produced vibrato, which have been discussed in the section on electronic organ design, are thus avoided.

A simple circuit which gives uniform pitch sensitivity over the whole keyboard is shown in Fig. 11. Force exerted on any key is applied to a stiff beam which forms the grounded plate of the tuning condenser of a radio-frequency oscillator. The frequency change caused by the very small motion of the beam is converted to a dc voltage by a discriminator in which the separation of the tuned circuits has been made considerably greater than that used for fm detection. By controlling the separation of the tuned circuits and the relative coupling of each to the oscillator, a relation between force on the beam and pitch of the final oscillator, similar to that shown in Fig. 10, may be achieved.

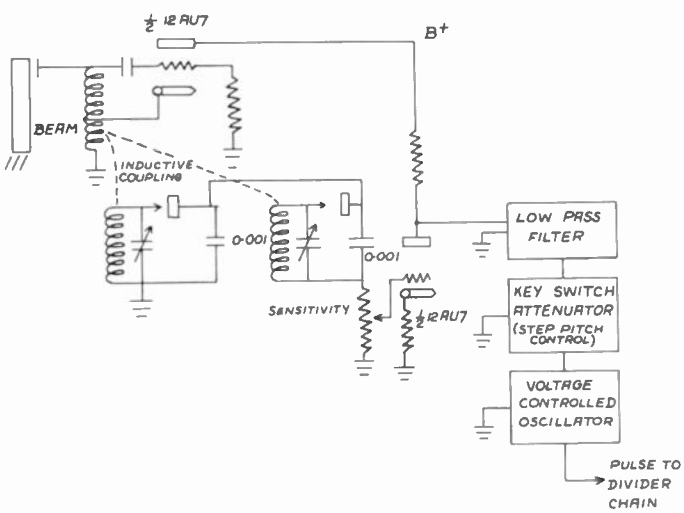

Fig. 11—Pitch control circuit which provides a relation between pitch and horizontal force on the key, of the type shown in Fig. 10.

The vertical force controls the loudness, thus enabling the player to determine the attack and decay of the note in addition to producing crescendo and diminuendo effects. Since there is only one control device for the whole keyboard, no engineering problem is involved. The musical considerations have been discussed earlier under "The Touch-Sensitive Organ."

Control of the timbre was removed from the key and put in charge of the left hand. The controls are arranged so as to stay where they are left; thus the left hand may be conscripted for use on the keyboard during rapid passages. Emphasis was placed on plastic control of the timbre, rather than imitation of special orchestral instruments or the provision of all conceivable timbres. Thus no stops were used, and the main controls are continuous, and divided up as nearly as possible so that a given change of position corresponds to a constant change of sensation. They are light, and the motion is restricted to not much over an inch. The controls may thus be moved fast enough to change the timbre during the attack of a note. A timbre vibrato[24] may be used. Timbre patterns of the same nature as those described earlier as characterizing existing instruments may be set up. The possible variation of timbre is of course much greater than that found in the familiar musical instruments. In addition to the timbre variations produced under the direct control of the player, an increase of brightness with intensity is provided by the use of nonlinear circuits. The only justification for this for this procedure is that it makes the tones sound more "natural."

The timbre controls are organized along the lines discussed in an earlier section. One control is a basic waveform control, which provides the basic waveforms which are found in musical instruments, such as square-wave, sawtooth, pulse, in addition to others. This is a two-dimensional control operated by moving a knob in a plane with the index finger. One dimension corresponds to brightness, and the other to octaveness. While this division is by no means an entirely logical one, it has been found of considerable help to the performer in arranging the timbres in his mind and realizing them on the instrument.

Control of the basic waveform has been obtained by the motion of a conducting felt pad over the surface of a disc of resistor card material. A recently revised form, shown in Fig. 12, consists of a circular plate divided into a number of separate conducting sections which are connected to various waveforms derived from the divider chain, which itself supplies square-waves at octavely related frequencies (f, $2f$, $4f$, $8f$ on the diagram). A conducting disc, which is shown by the dotted circle, is moved over the conducting sections, and is capacitively coupled to them. While the points corresponding to the basic waveforms of the various classes of instrument have been marked on the diagram, it has been pointed out earlier that timbre plays only a limited part in instrument recognition. Thus, if the basic waveform and formant controls are set to produce an

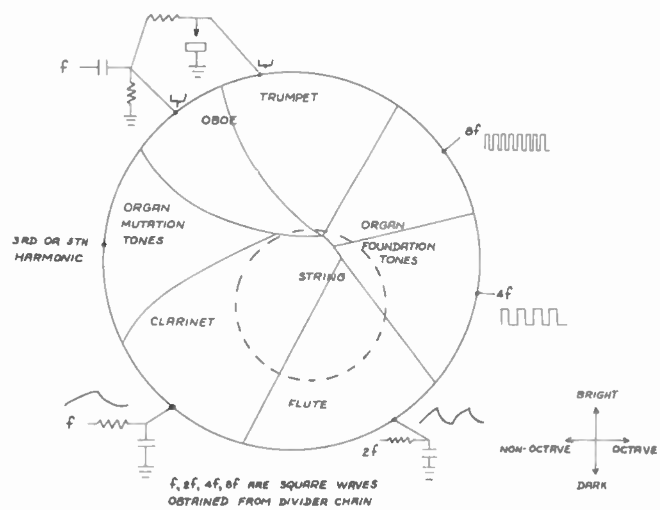

Fig. 12—Basic waveform control in the electronic "sackbut." A circular plate is divided into a number of conducting sections insulated from each other, shown by the solid lines. A conducting disc is capacitively coupled to the conducting sections. The dotted circle shows the relative size of the disc and the position where the basic waveform is approximately a sawtooth.

harmonic spectrum appropriate for a cello-like effect, the sound does not remotely suggest a cello if a fast attack is used.

The other timbre control, shown in Fig. 13, is the position of two formants which may be controlled together or singly by the thumb. Continuous control by the inductors covers a range of two octaves which may be located anywhere in the musical spectrum by auxiliary controls. Other auxiliary controls also control the Q of the formant circuits.

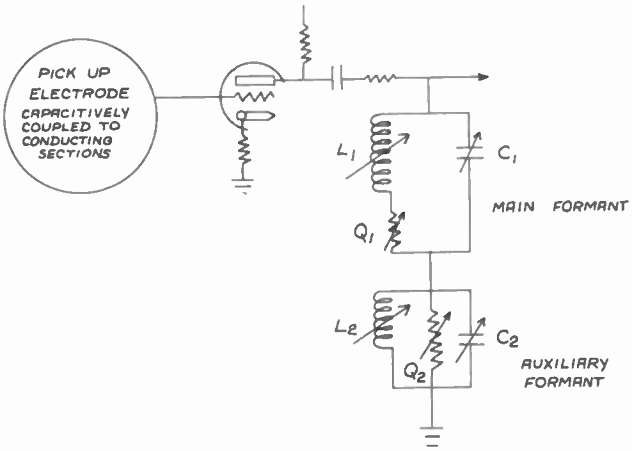

Fig. 13—Formant control in the electronic "sackbut."

The left hand normally occupies the position shown in Fig. 14 where the thumb controls the formant circuits, the index finger controls the basic waveform, and the three remaining fingers are available to operate keys which control departures of the waveform from simple periodicity, such as by frequency modulation of the pitch generator with a 20-c square-wave, or with narrow-band low frequency, or with wide-band noise.

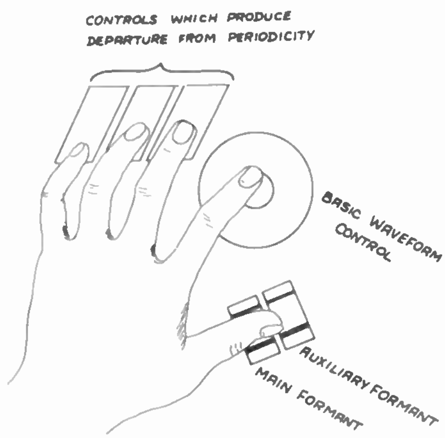

Fig. 14—Arrangement of timbre controls in the electronic "sackbut."

These controls are continuous, increasing force resulting in increasing depth of modulation, and are of the spring-return type, since they are seldom used for long periods.

In order to facilitate moving the range of the instrument up and down by octaves, all the waveforms for the timbre control circuits are derived from a chain of scale-of-two counters which are driven by the voltage-controlled oscillator. Changing the range of the instrument by octaves then involves simply moving all the take-off points together up or down the chain the required number of units. Since accelerated playback performances are now so frequently required by composers and arrangers, it is well worth adding several octaves to the lowest range normally required for direct performance so that the performer can record at a speed of one-quarter of the playback speed and have low notes on the record. A "monitor" output one or two octaves higher in pitch than the recorded output is also useful. The lowest note which can be used in direct performance is lower than one might expect. While no standard reproducers are capable of making a pure tone of 15 c loud enough to be musically useful, a musical tone having a pitch corresponding to 15 c may be quite effective when heard through a standard reproducer.

Development of this instrument stopped in 1948; however, further development has been undertaken recently at the National Research Council of Canada. The pitch flexibility of the instrument is being extended by combining a continuous pitch control with the keyboard. The combination is made in such a way that motion of the finger backward along any key transfers control from the key control to the continuous control without introducing any discontinuity in the sound. Thus it is possible to terminate a figure started on the continuous control, on a key on the keyboard, and vice versa. It is hoped that the integration of the two control means will be sufficiently good that the pitch flexibility obtained by the use of a continuous control may be combined with the ease of playing rapid passages which is characteristic of the keyboard.

A helical-spring reverberation device used in the earlier instrument will be replaced with one based on magnetic recording techniques. Using a magnetic tape feedback loop, it is possible to add to a musical instrument a reverberation pedal, shown in Fig. 15, which functions very much like the sustaining pedal on a piano, in that when the pedal is down, notes are prolonged after release of the key, and when the pedal is released the stored energy disappears rapidly. The decay constant of the sound may, however, be controlled accurately by the displacement of the pedal—a type of operation which is very difficult to achieve consistently with a damper pedal. Control of the reverberation by the signal itself offers new possibilities in suppressing undesired transient effects without introducing undesirably long decay times.

The Monophonic Instrument as an Organ Auxiliary

The Hammond Organ has a pedal auxiliary unit which consists of a monophonic oscillator.[36] Monophonic oscillators are sometimes found on the manuals of a small organ, primarily as a means of increasing the

[36] R. L. Eby, "Electronic Organs," Van Kampen Press, Inc., Wheaton, Ill., pp. 97–140, 1953.

resources at comparatively small expense.[37,38] Since the monophonic oscillator is not synchronized with the main organ generators, such devices also produce musically useful beats. Within limits, they also permit solo and accompaniment to be played with one hand. Similar devices have been provided on the pipe organ to achieve this last result,[39] but were rather infrequently used by organists.

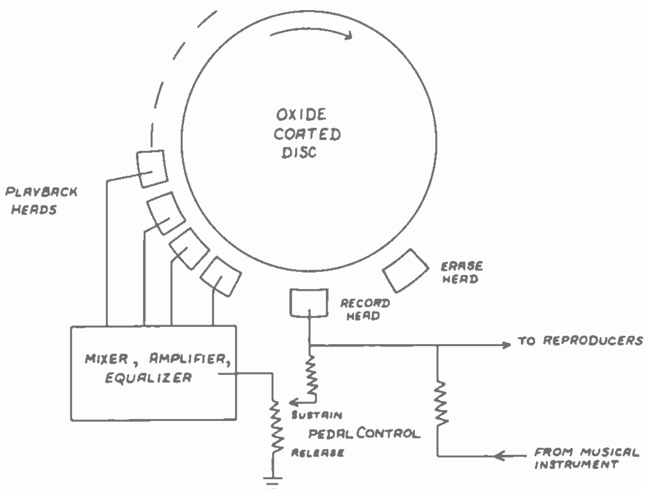

Fig. 15—A "reverberation pedal" which may be fitted to an electronic musical instrument and which performs in a manner similar to the sustaining pedal of the piano.

European Electronic Monophonic Instruments

An electronic monophonic instrument manufactured in Germany, the Hohner "Electronium,"[37] is superficially similar to the Hammond "Solovox," and is mounted in an accordion case. Control of the loudness is obtained through operation of the bellows.

The "Ondes Martenot"[40] appeared in its earliest form in France in the late "Twenties." Some of France's leading composers have written for it. The most celebrated instance was its use in "Joan of Arc at the Stake" (Honneger, 1938). It was heard in Boston in 1950 in "A Concerto for Ondes Martenot and Orchestra" by Andre Joilvet.[41]

A new monophonic instrument, the "Melochord,"[42] shown in Fig. 16, which has become an indispensable tool in the creation of electronic music in the Studio for Electronic Music at Cologne, is an electronic keyed instrument with two independent monophonic playing ranges, and can function as a normal musical instrument, as well as produce the effects required for electronic sound composition. Normally each playing range of the "Melochord" contains a separate tonal generator and sound filters of various kinds belonging to each generator. Each playing range also has a control device which functions to produce known attacks and decays, such as the sound produced by wind and plucked-string instruments. There is also a vibrato generator. Each playing range is provided with a swell pedal for controlling the loudness. Naturally, the studio is provided with echo chambers, tape loops, and many other devices both well-known and new, and the policy followed in developing the instrument is to include on it only those devices directly connected with the sound generation.

Fig. 16—"Melochord" of Studio for Electronic Music at Cologne. (Photograph courtesy of Nordwestdeutscher Rundfunk.)

A particularly interesting feature of the "Melochord" is the sound filter which can be tuned by the operation of the keys in step with the pitch of the generators. The "traveling formant" is obtained by means of a special set of normally closed contacts on the keyboard. By means of these contacts all of the condensers which happen to be on the left-hand side of the key touched, are separated from the series of condensers connected in parallel. The sum of condensers thus remaining on the right-hand side serves for tuning the parallel resonant circuit formed by these condensers and an inductance, which latter is also variable.

In addition to adding the outputs from the two keyboards, as is usually done, they may be connected to a ring-modulator and mixed in multiplication. A few examples of the effects which may be obtained with this instrument are given in the reference quoted above.[42] By turning off the second generator, and including the step-by-step filter in Channel 1, it is possible to determine the pitch and the nonsteady processes by playing the keys of Channel 1; while playing the keys of Channel 2 influences the timbre. By turning off Channel 2 and connecting a noise generator, it is possible to produce a noise sound having pitch, which is reminiscent of sound produced by the wind. By connecting impulse

[37] See *Instrumentenbau Zeitschrift,* (special electronic instruments number), particularly, "A technical survey of electronic instruments," pp. 170–172; April, 1954.
[38] Eby, *op. cit.*, pp. 123–129.
[39] J. I. Wedgewood, "A Dictionary of Organ Stops," Vincent Music Co., London, Eng., p. 100, 1905.
[40] Dorf, *op. cit.*, pp. 180–181.
[41] E. Vuillermot, "Odd concerto for Ondes Martenot and orchestra," *Christian Science Monitor, Magazine Section;* March 11, 1950.
[42] H. Bode, "The Melochord of the Cologne Studio for Electronic Music," (see footnote 58).

noises to the input, various "plucked" effects having various timbres may be obtained. Again, the music of an orchestra may be connected at the generator input; then the music appears "modulated" with vocal colors.

Another instrument developed for the Cologne Studio is the "Electronic Monochord,"[43] shown in Fig. 17. This instrument was developed from the "Trautonium" which appeared in the late "Twenties." The new version may be used either in the direct performance of music, or for the composition of electronic music. The new instrument contains two monophonic electronic sound producers in one cabinet. While in some versions several oscillators having harmonically related frequencies have been included, this was not considered necessary for an instrument designed for electronic sound composition. In describing the instrument, Trautwein emphasizes the necessity for the player to have a freely variable frequency range at his disposal. Some earlier models used thyratrons as the frequency determining element, and potentiometers with a linear taper as the control element. By proper choice of the voltage across

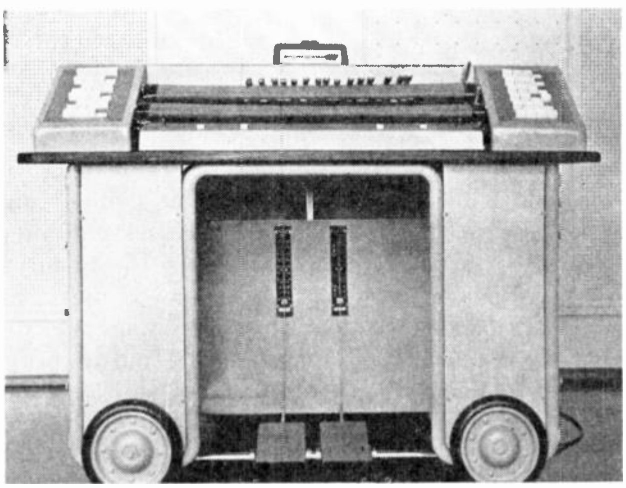

Fig. 17—The "Electronic Monochord" of the Studio for Electronic Music at Cologne. (Photograph courtesy of Nordwestdeutscher Rundfunk.)

the potentiometer and the fixed grid bias, it was possible to get three octaves which were approximately logarithmic. In the "Electronic Monochord" at the Cologne Studio, the grid resistor is the control element. It is coiled around a drum of elliptic cross-section; the return connection is provided by a piece of braided material placed a few millimeters distance from the drum. The spacing of the winding in this design can be so arranged that a semitone step is the same length anywhere on the keyboard. Elastic keys have been placed above the playing device so that desired frequencies may be selected easily. There are five keys over each octave which give a pattern similar to that of the black keys of the standard keyboard. The keys are, however, tuned to the pitches required in the composition.

[43] F. Trautwein, "The Electronic Monochord," (see footnote 58).

The basic waveform of the device is a sawtooth with an exponential curve of amplitude. The low notes are strongly exponential in character, and thus are richer in overtones than the high notes, with their almost linear curves. Pulses may also be taken from a cathode resistor. Formant filters serve for the formation of tone colors. These filters consist of band filters, acceptor circuits, and suppression filters, which permit selection of a widely variable frequency response.

For the concert hall a loudspeaker output of 150 w, producing a level of 110 db is considered necessary. On each manual two devices have been provided for controlling the signal energy: a pedal, and a control consisting of a pressure-dependent resistance which is placed underneath the elastically supported playing device.

The Coded-Performance Instrument

The distinguishing feature of coded or mechanically performed music is that it is completely written down in advance, and the performance consists simply of carrying out the written instructions. The difference between this kind of music and conventional music may not seem very great, but the complete writing down of a musical performance has been attempted only in certain special cases. In the early days of the player piano, the rolls were punched directly from the written music, but these were soon replaced by "hand-played" rolls. In view of the mechanical sound of the "hand-played" rolls when played on the average player piano, it may seem surprising that a workable solution to the problem of writing down the nuances of tempo was never obtained. Seashore's work with the piano-camera has revealed, however, that time differences of one-hundredth of a second may be significant in piano performances.[44] His work also shows that the distortion of the tempo in rubato passages may amount to one hundred per cent of the bar length, and that such distortions are organized in complicated patterns. These patterns have never been analyzed directly, but are learned by the musician by a subconscious process, as canons of musical or artistic taste. When preparing for a performance, a performer may choose between a number of well-known alternatives, or he may invent new patterns, both processes producing a variation in the performance of different artists which is known as "interpretation." Nevertheless, there is a great deal of information in the final performance which is known to both performer and composer, yet does not appear on the written music. The same may be said of the pitch, loudness, and timbre patterns which appear in the final performance. The codification of these patterns constitutes a formidable task to be completed before the direct production of familiar music can become routine. The production of novel effects, and the per-

[44] C. E. Seashore, "Objective Analysis of Musical Performance," Studies in the Psychology of Music Series, vol. 4, The University of Iowa, Iowa City, Iowa; 1936.

formance of *avant-garde* compositions are likely to be much more successful, since for the presentation of these works the task of translating the very large body of musical conventions into terms appropriate to the apparatus need not be completed so thoroughly.

In spite of the lack of knowledge concerning the precise nature of the unwritten parts of music, composers have always shown an interest in writing for mechanical instruments. Handel wrote a considerable amount of music for the barrel organ, giving instructions for the placing of the pins. Stravinsky and others have written music to be punched directly onto player piano rolls. Modern mechanical or "synthetic" music began perhaps about 1000 A.D. with the invention of clockwork, although mechanically-operated sound-producing devices were used earlier than this in connection with religious services. The continuous invention of mechanical instruments from the advent of clockwork to the present day forms a fascinating story.[45] One of King Henry VIII's prized possessions was a player-virginals. Maelzel constructed a mechanical orchestra of over forty instruments, in which, in addition to the sound-producing mechanism, there were clockwork figures which actually went through the motions of playing their instruments.

Electronic Coded-Performance Instruments

The first electronic "synthetic music" instruments were naturally perforated-roll-operated, following the example of perforated-roll-operated pianos and organs. There is, of course, no reason why a performance on a pipe organ with a properly-programmed perforated roll should be inferior in any way to a performance by a human being, since all the human performer can do is to close the same switches.[46] It is worthy of note, however, that the vast number of organ records now in print are hand-played, in spite of the obvious superiority of the roll-operated organ as a performer.

The first public demonstration of "synthetic music" made by electronic devices was at the Paris Exposition of 1929, where a roll-operated device consisting of four monophonic electronic oscillators was shown with great success. Following the basic patent covering this device,[47] there are other similar French patents. In one of these,[48] a number of different devices are described, that allow the composer or arranger to draw by hand the sound envelope. In one form of the invention, the arranger engraves a groove in a suitable support which varies either in depth or in position at right angles to the time axis. When the music is reproduced, a needle following the groove operates an optical wedge to control the light passing through a sound-on-film recording to a photocell. In another form of the invention, the arranger draws by hand in conductive ink, a mark of varying width or position which is read by a series of brushes to set up the sound envelope. As a sound source, the inventor uses a "sound library" (sonothèque) consisting of suitable supports on which are recorded by any known method the various notes of the various instruments, in addition to vocal sounds and other noises. As an alternative, synthesis from pure tones is mentioned.

The first United States patent on an electronic synthetic music device was applied for by Hanert in 1945,[49] and assigned to the makers of the Hammond Organ. Hanert described monophonic oscillators operated by a punched-card system which overcomes some of the limitations of the perforated roll. He paid special attention to the problem of making it as easy as possible to alter the performance until a satisfactory result was obtained. The operation of Hanert's electronic music synthesizer was based upon the breakdown of a tone into characteristics such as frequency, intensity, growth, duration, decay, portamento, timbre, and vibrato.

Animated Sound

While the use of electronic generators to replace the prior art generators such as violin strings, pipes, chimes, and so forth, evidently increases enormously the scope of a mechanical instrument, the first really new development in the art of mechanical music since 1000 A.D. was the idea of abandoning the time-honored parameters: pitch, amplitude, and timbre, for one single parameter, that associated with the sound wave. Thus any sound whatever, including the sound made by a 100-piece symphony orchestra[50] may be considered simply as a particular function of time. Once this function is specified, the sound may be heard by the use of suitable apparatus. While this revolutionary new point of view was implicit in the theory of the propagation of sound waves, and later pointed up by the invention of the phonograph, really convenient apparatus did not appear until the sound film.

McLaren of the National Film Board of Canada, has, since 1939, made a special study of this method of producing sound.[51] His beautiful and imaginative films with hand-drawn sound were responsible for a renewed interest in all methods for the production of sound without direct performance upon a musical instrument. McLaren made no attempt to imitate the conventional instruments, but used sound waves having simple and easily drawn shapes. A diagram of radiating lines placed

[45] "Oxford Companion to Music," Second American Edition, Oxford University Press, New York, N. Y., pp. 549–556; 1943.

[46] The automatic player manufactured by the Moller Organ Co. is capable of playing six independent tonal lines at the same time. (See W. H. Barnes, "The Contemporary Organ," J. Fisher Bros. New York, N. Y., p. 315, 1948.) Thus this mechanical performer has six "hands."

[47] E. E. Coupleux and J. A. Givelet (France), U. S. Patent 1,957,392, applied for in 1930.

[48] L. Lavallee, French Patent 806,076, applied for in 1936.

[49] J. M. Hanert, U. S. Patent, 2,541,051, applied for in 1945.

[50] Because of the enormous amount of information involved, the hand drawing of such a sound is of course not practicable.

[51] R. E. Lewis and N. McLaren, "Synthetic sound on film," *J. Soc. Mot. Pict. Engrs.*, vol. 50, pp. 233–247; March, 1948.

under the film was used to obtain the basic spacing required to give the desired pitch. Fig. 18 shows a somewhat different method used by McLaren for putting a sound track together. A series of variable density patterns, seen in a card file at the right of the figure, were photographed through an adjustable mask seen at the left. In McLaren's most recent film, simple marks made with a needle point directly on 35-millimeter black emulsion film were used. While the marks are easy to draw, the sounds so produced are very complex from the point of view of pitch, loudness, and timbre, and are well suited to the development of the complex rhythms which have interested composers since the turn of the Century.

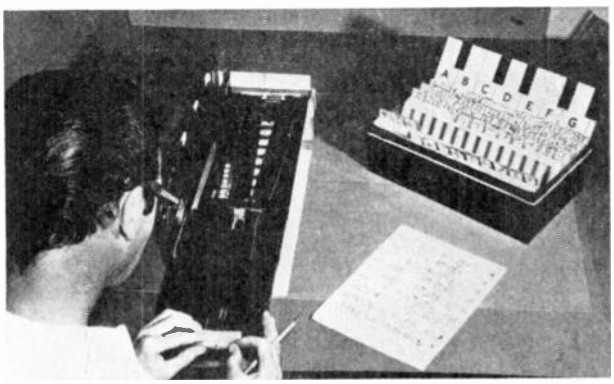

Fig. 18—One of the animated sound techniques developed by Norman McLaren of The Natl. Film Board of Canada. A series of variable density patterns seen in the card file at the right of the figure are photographed through an adjustable mask seen at the left. (Photograph courtesy of The Natl. Film Board of Canada.)

Recent Electronic Coded-Performance Devices

An electronic method of direct sound production based on McLaren's work is due to Kendall, also of the National Film Board of Canada. In Kendall's method[52] the adjustable mask shown in Fig. 18 is replaced by a hand-drawn sound envelope which is applied to the output of an electronic generator by means of a cathode-ray tube curve reader. This instrument is being developed by the Canadian Marconi Company.

A great deal of interest was aroused earlier this year by the Radio Corporation of America's disclosure of a perforated-roll-operated device, designed by Olson at the RCA Laboratories, Princeton, N. J. Olson's apparatus uses a pair of monophonic oscillators controlled by a coded record on a perforated roll. The holes control the pitch, timbre, growth, duration, intensity, and decay of the note sounded by the monophonic oscillators. The circuits and mechanical devices used to achieve this control are described in detail in a recent paper.[53] The oscillators are used on alternate notes so that the decay portion of one note may be continued during the attack portion of the next note, the combination thus producing what might be regarded loosely as a single monophonic part or line of notes. A sufficient number of individual parts to form the musical composition are combined by multiple recording.

The versatility of the device is shown by the fact that intelligible speech can be produced from a very simple code (525 bits per second). It may be noted in passing, however, that to synthesize with this apparatus the familiar sound produced by a glissando on the keys of a piano with the sustaining pedal down, would require forty-four individual record tracks. This difficulty arises when an effect similar to glissandos, arpeggios, and rapid figures on polyphonic instruments is required, and represents a sacrifice of convenience which must be made to gain the independence of parts and pitch flexibility which can be obtained with a monophonic instrument.

New Musical Horizons Through Electronics

Up to this point we have considered the problem of fitting electronic musical instruments into the existing musical scheme. It is here, of course, that the significant commercial developments of electronic musical instruments are to be expected. The nature of our musical system, however, is dependent to a large extent upon the musical instruments for which composers have been writing. If electronic instruments have disappointed us by not immediately replacing archaic instruments constructed of such nonmodern components as catgut and horsehair, this is due in part to the fact that electronic techniques of sound production belong to a musical system which has not yet been evolved. The experiments directed toward the discovery of this musical system made by workers in the "Twenties" were enthusiastic and promising, but were slowed down by the economic hardships of the "Thirties" and World War II. In the rebirth of interest in electronic music which took place after the war, sound storage devices have played a large part.

Before a musical performance can be presented it must be conceived or planned by the composer or the arranger. To aid in building a mental impression of what the final performance will be like, composers have traditionally used three methods: (a) the parts are played over separately on an instrument, a mental impression of the total effect being thus built up; (b) one or more parts are played over slowly and speeded up in the imagination, and (c) short sections are played over on an instrument and joined together in the imagination. Modern sound-storage devices allow a musical composition to be put together by three well-known recording techniques which carry out physically the three processes traditionally used by the composer in the first stage of conceiving the composition. They are: multiple recording, accelerated playback, and splicing of the record. The sound-composition techniques up around these recording processes combine

[52] A. Phillips, "Osmond Kendall's marvelous music machine," *Maclean's Magazine*, pp. 22–23, 52–56; June 11, 1955.
[53] H. F. Olson and H. Belar, "Electronic music synthesizer," *J. Acoust. Soc. Amer.*, vol. 27, pp. 595–612; May, 1955.

direct performance of parts of the composition with mechanical assembling of other parts. In the United States, experiments on sound-composition processes are being conducted by the "Music for Tape Recorder" group (V. A. Ussachevsky, Otto Luening, John Cage, Christian Wolff, Earl Brown, and Louis and Bebe Barron). The work of this group, including Cage's "Music for Prepared Piano" combines natural and electronic sounds with sounds which have been altered by the recording process. In Europe, the Musique Concrète group (Pierre Schaeffer, Pierre Henry, and their associates) limit themselves to natural or "concrete" sounds picked up by a microphone.[54]

The Musique Concrète group has found that the arbitrary cutting of a piece of recorded material which can be reproduced as often as desired leads to considering this piece as a "sound object," either an elementary one, or one capable of being broken down further. Because it is stored in a permanent form, it can be analyzed physically and musically and a judgment may be made of its value when taken by itself, dissociated from the group of phenomena from which it has been taken. It can be used as it is, or it may be transformed by various electroacoustic processes to give birth to new sound objects.

The assembling and evaluation of sound objects is part of the work of the Musique Concrète group. Their extensive library includes such sound objects as the falling of a drop of water, the sound of a gong without the impact by which the sound is produced, and the click of a Chinese block. This sound library recalls the "sonothèque" of Lavallee, and Respighi's use of the recorded voice of nightingales stored in an earlier form—the phonograph disc—and used as a complex sound object in "The Pines of Rome" (1924). The transformation of sound objects may be compared to Milhaud's creative use of a change of playback speed in the late "Twenties" to produce a chorus of voices singing above the pitch range of a normal voice. Schaffer has given in his book,[55] a fascinating account of the development of the ideas of "musique concrète" and of the processes which occur during the conception and putting together of a piece of "musique concrète."

Two pieces of apparatus used by the Musique Concrète group are the "Phonogène" which transposes the spectrum from one pitch to another by variation of tape drive speed, and the "Morphophone" which controls the sound envelope. Multiple recording technique and splicing of the tape record are used to assemble the sound elements so formed. The Phonogène, which is shown in Fig. 19 is a 24-speed tape recorder which uses a closed tape-loop. Twelve drive spindles may be seen, each having its own pressure wheel. The speeds correspond to the twelve semitones in the equitempered chromatic scale. The drive spindles may be engaged by depressing one key of a 12-key piano keyboard which does not appear in the figure. The motor speed may be changed by a factor of two to extend the number of tape-drive speeds to twenty-four. Rerecording to a constant speed tape recorder and back to the Phonogène allows further extension of the speed range. The Phonogène carries out a rigorous translation of the spectrum, while transposition on the well known musical instruments is not exact at all, since the harmonic spectrum varies within wide limits over the range of the instrument. The attack and decay are also varied with the transposition on the Phonogène, transposition to a lower pitch bringing about an increase in the time of build-up and decay, and an increase in the reverberation which accompanies the note. Conversely,

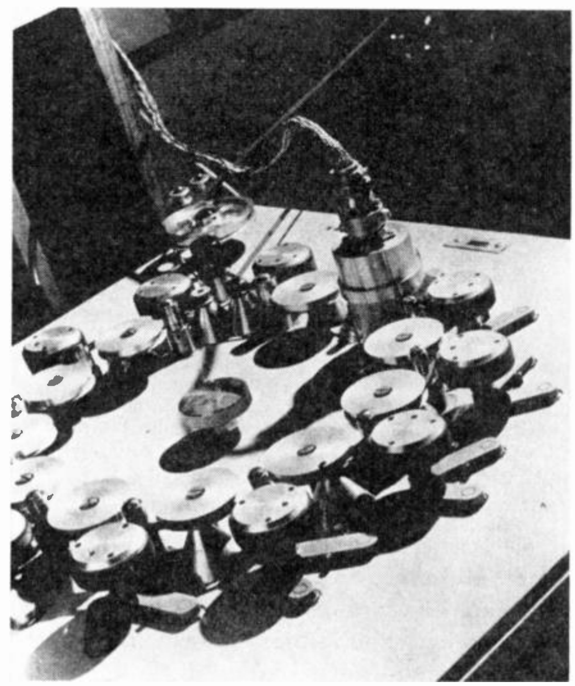

Fig. 19—The "Phonogène" of the Musique Concrète Group of the French Radio and Television System in Paris. A tape loop may be driven at twenty-four speeds using a keyboard control which is not shown. (Photograph courtesy of UNESCO.)

a note transposed to higher pitch will have a faster build-up and decay, and less reverberation. The term "complex note" is used to designate a superposition of simple notes such as a chord, or a sound obtained from sampling a sound complex. These sounds, as well as noises, vocal sounds, or rhythmic patterns formed by groups of notes, may also be transposed by the Phonogène. In the slide-Phonogène the tape may be run at continuously variable speeds.

The Morphophone is a closed-loop tape playback device possessing ten playback heads, each head connected to a preamplifier, the gain and response curve of which can be modified with the aid of simple filters. The outputs of the preamplifiers are mixed to feed the output amplifier. A primary use of the apparatus is to provide artificial reverberation of variable form and

[54] J. Poullin, "L'apport des techniques d'enregistrement dans la fabrication de matières et de formes musicales nouvelles. Applications à la Musique Concrète," *Onde Elect.*, vol. 34, pp. 282–291; March, 1954.

[55] P. Schaffer, "A la Recherche d'Une Musique Concrète," Editions du Seuil, Paris, France, 1952.

color, since the decay time of every sound phenomenon depends upon the preamplifier playback controls. Another device permits the superposition of an envelope belonging to one sound upon another sound; thus a plucked string attack may be imposed upon a normally steady sound.

Musicians have shown considerable interest in a mechanism for moving the source of sound about in space. It is a piece of apparatus which resembles that used for stereophonic projection, except that the distribution of sound output is determined by the composer or the conductor. Four loud-speakers are used: two placed in front of the audience, one above, and one behind. These are connected to four channels on synchronized tapes. When used in this way the space projection is known as "relief statique," "statique" meaning that the distribution of sound among the speakers is prearranged. Alternatively, a conductor or "executant," by moving a small magnetic unit through the air during a performance, can control the distribution of loudspeaker outputs by the relation between the movable magnetic unit and four stationary loops connected to the four loudspeakers. In a ten-day series of lectures and concerts given in 1953, organized by the Groupe de Recherches de Musique Concrète de la Radio-Television Française, American and French workers in the field of sound composition on magnetic tape "projected" their own recorded compositions. Alternatively, a five-track tape recording may be used, the conductor controlling during the performance the distribution of the sound on the fifth track, while the distribution of the sound on the other four is prearranged.

The music produced by the Musique Concrète group recalls Debussy's interest in Javanese music and the rhythmic experiments of Varèse in the "Twenties." Speaking in 1932, Stokowski,[56] who also found himself attracted by Javanese rhythms, predicted that one of the contributions made to music by scientific techniques would be the exploration of new and complicated rhythmic patterns which are too difficult for present-day performers and instruments. He mentioned in the same lecture the possibility of recording a particularly favorable note in the range of a singer and reproducing it at all pitches by sound-on-film recording technique. Here he probably had in mind Radio Station WCAU's photoelectric organ on the design of which he collaborated.[57]

One example of the work of the Musique Concrète group presents bursts of sound from a number of unrelated sources in a complicated rhythmic pattern. One of the sources is a single chord sustained on an organ. In another example, after striking a key on the piano, by variation of drive-speed, the pitch of the note is raised several octaves and returned to normal pitch, the time for this maneuver being about one second. This bizarre sound is combined with other previously recorded piano sounds which have also been "denatured" in a variety of ways. Again, the rhythmic patterns are complex. Poullin,[54] speaking of the numerous works of the Musique Concrète group, says, "Their first hearing may excite a lively interest, astonishment, indifference, or severe censure." (The writer believes that the second last reaction is rather rare.)

The approach of the Cologne Studio for Electronic Music is to dispense entirely, not only with natural or concrete sounds picked up by a microphone, but with preformed sounds of any kind. An entire issue of the technical magazine published by Nordwestdeutschen Rundfunks has been devoted to their work. The acoustical and musical bases for their viewpoint of music are explained, and the apparatus and musical work of the North-West German Broadcasting Station at Cologne are described. Only a very brief outline of this important development will be given here, and the reader is urged to consult the original publication.[58] This group has been concerned principally with discovering the essential nature of music which is "electronic" in the restricted sense of the word. Eimert[59] believes that the point of departure for electronic music is the rationalization of musical elements because of which music is no longer reducible to manual performance. Busoni regarded the world of sound as a continuum, of which the world of our traditional music was a very small part. Eimert sees in electronic music a means of permitting the composer to work in the whole continuum. The limits of playability are abolished, or replaced with the limits of audibility.

Eimert places electronic-acoustic phenomena in five classes:[60]

1) A simple tone, or a pure sinusoidal tone without overtones. These are not found in traditional music but are the basis of all musical sound processes. Sinusoidal tones cannot be used to build up a system of tones in the traditional sense since they have no "tonal" character.

[56] L. Stokowski, "New horizons in music," *J. Acout. Soc. Amer.*, vol. 4, pp. 11–19; 1932–33.

[57] "WCAU's photocell organ," *Electronics*, vol. 7, p. 157; May, 1934.

[58] "Technische Hausmitteilungen des Nordwestdeutscher Rundfunk," vol. 6, 1954. The papers from this volume which are listed below, have been translated by the National Research Council of Canada. Copies of the translations are available at a nominal fee from the Library, National Research Council, Ottawa 2, Can.
H. Eimert, "Electronic music"
W. Meyer-Eppler, "Terminology of electronic music"
F. Enkel, "The technical facilities of the electronic music studio of the Cologne Broadcasting Station"
F. Enkel and H. Schütz, "Magnetic-tape technique"
F. Trautwein, "The Electronic Monochord"
H. Bode, "The Melochord of the Cologne Studio for Electronic Music"
W. Meyer-Eppler, "The mathematical-acoustical fundamentals of electrical sound composition"
F. Enkel and H. Schütz, "The production of sound effects for radio dramas"
H. Eimert, "The musical situation"
K. Stockhausen, "Composition 1953, no. 2"
H. Eimert, F. Enkel, and K. Stockhausen, "Problems of electronic music notation."

[59] H. Eimert, "Electronic music," (see footnote 58).

[60] H. Eimert, "The musical situation," (see footnote 58).

2) A complex tone with harmonic partials, which is composed of a series of simple tones whose frequencies form an harmonic series. The "tone" of an instrument is a complex sound determined by the components of the series. In normal instruments the timbre can be varied over a comparatively small range. Electronics, now, for the first time, makes these components variable.
3) A tone mixture, in which the frequencies of the partials are not harmonically related to the fundamentals and cannot be expressed in terms of integral ratios. Mixtures of sinusoidal tones are not to be confused with chords: they have a higher binding level and can turn into "musical sound" much more readily than instrumental chords. In electronic music steady tone mixtures can be realized without effort.
4) A noise, which is determined by its specific character and pitch. The pitch of colored noise has approximate value only; the so-called "white noise," which extends over the entire auditory range, has no pitch at all.
5) A combination of two or more different complex tones sounded simultaneously; that is, an interval or a chord. In instrumental music "complex tone" and "chord" are clearly distinguished from one another. In electronic music, however, the tone mixture, with its new binding levels becomes a bridge between the two. Complex tones and mixtures can be composed electronically according to a prescribed composition arrangement.

The Electronic Music Studio of the Cologne Broadcasting Station contains three classes of equipment:[61] 1) electronic sound and noise producing sources, to provide the raw material for further processing; 2) electroacoustic shaping means, for the purpose of influencing the frequency spectra and the transient processes of sound phenomena—the methods used here are taken from communications techniques and yield sound phenomena which cannot be obtained by mechanical methods; 3) mechanical sound-recording apparatus for further processing of material obtained with 1) and 2).

1) One of the characteristics of the electronic sound-producers is that timbre is freely variable and is thus accessible as a new shaping element. The various available apparatuses for the purpose of generating the original sounds are: an Electronic Monochord, a Melochord (both of which have been described earlier in this paper), a generator of "white noise," and a number of standard laboratory audio-frequency generators. The possibility afforded by the Monochord of producing any desired musical scale conveniently and independly at fixed tone intervals is regarded as particularly valuable. A special noise source is used for the production of white noise. A beat-frequency oscillator is very useful for covering the entire audio-frequency range. For obtaining musical intervals accurately, audio-frequency generators with a comparatively large constancy are required. Commercially sold bridge-stabilized arrangements with decade frequency adjustments have a frequency accuracy of one part in a thousand, and have proven very satisfactory for this purpose.
2) A ring modulator is used to produce a multiplicative mixture of sounds which are quite different from the original sounds. An example of a multiplicative mixture is given later. Sounds are altered by playing back at speeds other than the recording speed. While the intervals remain unchanged, the transient processes are subjected to a far-reaching transformation. By playback at speeds greater than the recording speed, especially fine effects can be produced, which, owing to the size and weight of the acoustic apparatus which would be needed, could not possibly be attained by purely mechanical means. It is also possible to create echo times of extreme length by playing the record more slowly than it is recorded. Vibrato effects may be added by rhythmic tape-speed variation. For the production of restricted frequency bands from broad spectra, low-pass, high-pass, and band-pass filters are used. Eight octave filters having consecutive pass-bands have been used, in addition to high-pass, band-pass, and low-pass filters.

   Rhythm may be imposed easily on musical sound structures by means of ring modulators and tape loops. The sequence of sounds and the audio-frequency voltages picked up from a scanned tape loop are fed to the input terminals of a ring modulator. The sound is passed only when the pulses on the tape loop occur. There is also an arrangement which permits the timbre structure to be varied rhythmically with the aid of a set of octave filters. For this purpose, eight control frequencies are recorded on tape.

   Means are provided for hand-drawing the sound envelope. A gain regulating device is controlled by a photo-resistor, which is controlled in turn by the varying density of motion picture film carrying a hand-drawn pattern. The pattern is obtained by applying quick-drying varnish. For short reverberation times a real reverberation chamber is used. For discrete echoes a magnetic-tape echo is used.
3) Four synchronized tracks are provided by two tape layers side-by-side, with a synchronized drive driving perforated magnetic tape. The material on three recorded tracks is transferred to the fourth, the three tracks then being erased so that an arbitrary number of layers can be combined. Eighteen loudspeakers along the sides and

[61] Enkel, *loc. cit.*

ends of the studio are divided up into three separate groups, each being connected to one reproducing channel of the four-track tape recorder. In this way an arrangement of three sound sources in space is obtained. A considerable increase in the emotional response evoked by the sound patterns can be attained by changing the apparent distance and volume of the sound sources at the recording stage on the tape by electroacoustic means. The fourth track is used for control functions similar to those performed by the tape loop used for obtaining rhythmic effects.

The presentation and reproduction of sounds and noises is greatly facilitated by their clear characterization. A ring modulator in combination with a reflecting galvanometer having a build-up time of one second, and hence a resolving power of one cps, proved very useful. The arrangement yields high selectivity up to a fraction of one cps.

The music of the Cologne group demonstrates in a striking way the tremendous range of new sound possibilities provided by electronic techniques. Multiplicative mixture of sounds is an example.[10] When two sinusoidal voltages are mixed in suitable apparatus such as a ring modulator, the original frequencies are replaced by the sum and difference frequencies. If a complex tone is kept constantly at a fundamental frequency of, say, 300 cps, while a sinusoidal tone traverses all the frequencies starting from zero in glissando, the time-frequency spectrum of the product of the two oscillations shows the structure of Fig. 20. The

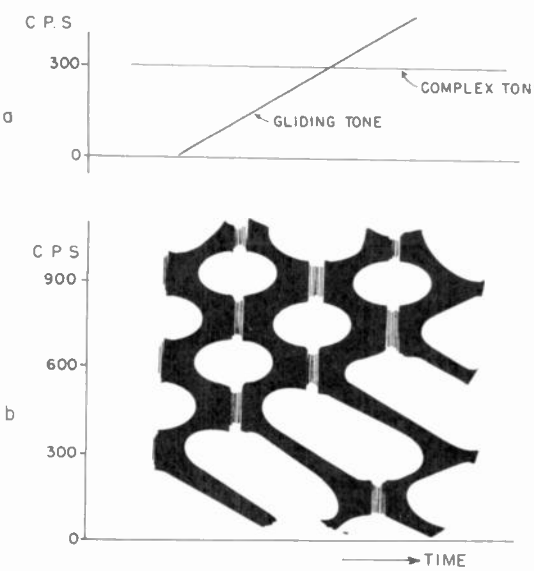

Fig. 20—*b* is a sound spectrogram of a complex tone of 300 cps. mixed multiplicatively with a sinusoidal gliding tone, starting from zero and varying with time, as shown at *a*. (After Meyer-Eppler.)

more rapidly the glissando is performed, the more closely will the result approach a noise. By mixing noise with a sliding tone multiplicatively, a sliding or howling noise with two spectral regions is obtained.

Schoenberg's idea of relating different timbres in a manner similar to that in which the notes of a melody are related, and Webern's permutations of sounds, lead directly to the shaping of sounds by the grouping of sinusoidal tones which is being investigated by the Cologne group. Stockhausen[62] described the first composition for sinusoidal tones which was produced at the Electronic Music Studio at Cologne. No ready-composed spectra such as produced by the Melochord or Trautonium were used at all. Simple frequency and amplitude conditions of the musical construction were set up, the nature of the sound following as a result.

The control of the composer or arranger over the sound is so much greater in the case of electronic music, that conventional musical notation is inadequate.[63] The logical way to describe electronic sound might be in the form of an acoustic representation; that is, the complete description of all the sound processes. A glance at the history of our present musical notation, however, suggests the importance of description by means of symbols rather than the complete representation. In setting up a new notation, care must of course be taken to make it as broad and as scientific as possible, rather than rational according to some previously known musical system, as is our conventional notation.

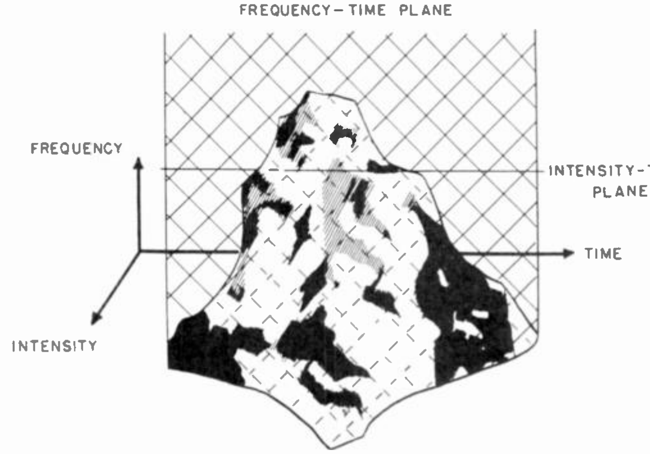

Fig. 21—Representation of an arbitrary sound. (After Eimert, Enkel and Stockhausen.)

Acoustic phenomena can be represented by a three-dimensional diagram employing the coordinates: frequency, intensity, and time. Sound processes which have time limitations are depicted as solids. It is possible to draw these solids using the rules of analytical geometry, by projecting them on various planes. For this purpose the solid representing the sound is cut into sections in such a way that all the details required for such a process are recognizable (see Fig. 21). In almost all cases it is sufficient for this purpose to have sections which depict the frequency-time plane and the gain-time plane. The number of sectional areas depends

[62] Stockhausen, *loc. cit.*
[63] Eimert, Enkel, and Stockhausen, *loc. cit.*

upon the structure of the sounds and noises, and must be great enough so that a reconstruction of the sound event is possible on the basis of the diagrammatic elements. It has been found useful for this purpose in each case to set up a frequency-time plane and the corresponding gain-time plane. The sound processes occurring in these planes can be described with the aid of a few simple symbols.

The Electronic Music Studio at Cologne characterizes acoustic properties in the following way.

Pitch and duration are plotted in the frequency-time plane, the exact amount of the frequency and the gain to be written above the line representing the partial (see Fig. 22). So that the entire auditory range can be covered by a single scale, the frequency range involved in each case is indicated by a factor placed in front.

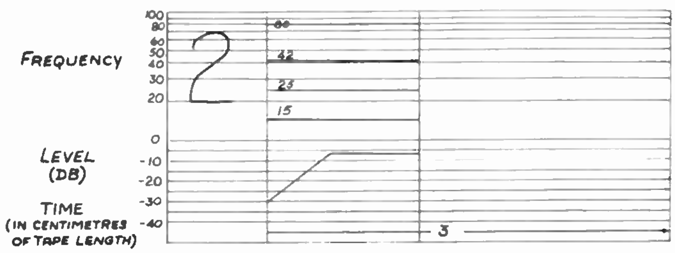

Fig. 22—Notation used by the Studio for Electronic Music at Cologne for characterizing sounds to be produced by electronic techniques. (After Eimert, Enkel and Stockhausen.)

The dynamic character of the sound event is obtained from the gain diagram. As seen in the example, the fade-out conditions are presented by a falling dashed line. For reasons which derive directly from the method of production, the time axis is graduated directly into lengths of magnetic tape referred to a tape speed of 66.2 centimeters per second.

If two sounds are to be modulated one by the other, the partial distribution of the original sounds are set down as above. This, of course, is done in two systems arranged one above the other. The kind of modulator to be employed is indicated in writing between the two systems. If a colored noise band of given width is required, this can be represented by two frequency lines joined at the beginning and at the end, showing the band limits. The dynamic characteristic, as usual, is indicated on the gain plane. If the frequency band has been compressed or expanded, the partial distribution of the sound to be compressed or expanded, as described earlier, is entered in the usual way and the expansion or compression value is indicated on the time scale; for example, the number "3" indicates a threefold expansion of the original sounds, while "$\frac{1}{4}$" means compression down to one-fourth of the original sound. By combining the above examples, and from the composition in a given case, a sufficient number of symbols may be assembled easily.

A score which uses only sinusoidal tones can be realized from the graphical instructions in the way that a manufactured article is produced from a drawing; that is, it can be executed by the technician. In other cases—for the complex sounds—the composer must specify the type of tone desired.

The notation used by the Cologne Studio may be compared with a system shown in Fig. 23, which is being experimented with by the Musique Concrète group.[54] Two staffs resembling those used in conventional music notation are used. Distance from the origin represents time. The melodic staff represents pitch in the conventional way. The character of the attack, the sustained portion, and the decay, are represented conventionally by appropriate signs. The dynamic staff may be effected by a clef sign of which the role is analogous to the clef sign in the melodic staff, and which refers to the vertical scale of the symbolic representation. In Fig. 23 the figure "30 db" indicates the range, and "0 db" indicates the level. Timbre is defined by a code where the large families are identified by letters placed at the start of the notation of the sound object being considered, for example, the letters "KZ" in Fig. 23.

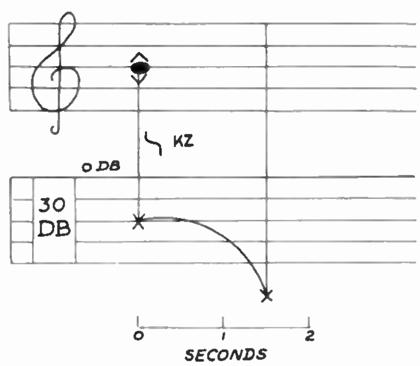

Fig. 23—Experimental notation used by the Musique Concrète Group. (After Poullin.)

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Transistors versus Vacuum Tubes*

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IT IS NOT the intention of this review to emphasize competition between transistors and vacuum tubes, as the title might suggest, although it must certainly be admitted that there is a contest between these devices in the important matters of technical manpower and funds supporting their development. Rather, the purposes in contrasting these devices are: first, to compare the status of their development and, second, to define their respective spheres of utility in the hope that such a survey may assist in guiding their future development along constructive lines.

The triode vacuum tube is forty-eight years old, the transistor seven, going on eight. Based on their relative maturity, one would expect sharply different scales of activity, and this is indeed the case. For example, one index of activity is the type numbers which have been assigned to transistors and vacuum tubes. The International Radio Tube Encyclopedia, the *Vade Mecum* published in Europe, lists all the type numbers of all the vacuum tubes ever produced anywhere in the world. The present total is 18,500 type numbers, many assigned to identical structures to be sure, but a very impressive figure nonetheless. Similar figures for transistors are harder to come by, but the latest compilation of type numbers produced in the United States contains approximately 235 types; and if we add foreign numbers, the list probably does not exceed 350.

Another index of activity is production. Total transistor production in seven years to date certainly does not exceed 4,000,000 units. We make that many receiving-type vacuum tubes in *two working days* in this country alone, as many as 42 million of them in one month, 450,000,000 in the last 12 months. Moreover, at a conservative estimate, the all-time worldwide production of vacuum tubes exceeds seven billion—three tubes produced in a generation for every man, woman and child on the globe.

A third, most significant, index of technical accomplishment is the variety of practical applications to which the two devices have been put. It is feasible only to indicate the range of application in vacuum tubes by comparing two types which illustrate extremes of power-handling capacity. The grandfather of all vacuum tubes in this respect is the super-power beam triode, type 5831, a water-cooled, three-foot-high electronic engine which will absorb the astonishing plate input power of 650,000 watts. A pair of these tubes, in broadcast-band or medium-shortwave service, will deliver an average cw power of one million watts.

At the other extreme of power-handling capacity in tubes, consider the miniature rf amplifier tube used in the tuners of a majority of the 20,000,000 television sets produced since 1952, that is, since the adoption of the gold-plated grid in such tubes. This tube will amplify usefully at 200 megacycles a fringe-area television signal of 10 microvolts across 300 ohms, representing a power available from the antenna of a mere one-sixth of a micromicrowatt. We thus find that the type 5831 super-power transmitting triode customarily handles a signal power which exceeds that handled by the 6BQ7 miniature rf pentrode by *nineteen orders of magnitude*. This is, perhaps, a fair indication of the power range covered by vacuum tubes. Transistors cover a much smaller range of power, about thirteen orders of magnitude, reaching into the watts, rather than megawatts.

In one field, the transistor has completely displaced the vacuum tube. In 1954 the hearing aid industry produced some 360,000 instruments, only 25,000 of which used vacuum tubes. In 1956, no hearing aid in production uses any vacuum tubes at all.

The technical reason for this complete victory is the fact that the transistor is the most efficient amplifier of low-level audio-frequency energy known to science. It needs no A battery and it makes extraordinarily good use of the current provided by a low-voltage B battery. The net result is that the transistor hearing aid pays for itself, in lower cost of battery power, in a year of regular use. No wonder the transistor has taken over in this department. A similar trend, having less economic urgency behind it, is showing up in portable radio sets. And in battery-operated and airborne military electronic equipment the transistor definitely has its foot in the door, for similar reasons.

The growing strength of the transistor is even more impressive in another field, the electronic digital computer. Here economy of power consumption is important, but not so important as utter reliability. The transistor designers have been arguing for many months that transistors are more reliable than vacuum tubes in computing machinery, but those in charge of the funds which support computer developments have had understandable doubts. Now it appears that the doubts are being resolved and that there will be general agreement, before this year is out, that the transistor is indeed the preferred active element in computing machines.

The emphasis on reliability here comes from the fact that digital computers capable of taking on the problems of modern warfare and modern industry must contain thousands, or tens of thousands, of tubes or transistors, and the computers must operate without error or interruption for many days or weeks at a time. Since the

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† Philco Corp., Philadelphia, Pa.

failure of even one tube or transistor from among the thousands in use can produce large errors in the computation, and since failure occurs in a statistical distribution, it is necessary that the life expectancy of the individual tube or transistor shall be many thousands of times as long as the period of uninterrupted service demanded of the computer as a whole. Thus a week of tube life is not sufficient for a week of computer service; rather thousands of weeks of life are needed in the individual units. The life expectancy of the transistor, for reasons to be described later, is far greater than that of a vacuum tube of similar properties and cost; hence, its commanding position in this field.

On another front, however, the transistor does not have the commanding position. This is the field of extremely high-frequency operation. To achieve high-frequency operation in any electronic device, the charge carriers (electrons or holes) must move through the active control region of the device in an extremely short period of time. This means that the active region must be of small dimension, normal to the charge flow, and that the motion of the charge carriers must be rapid.

On the first score, small dimensions, the transistor is already ahead. For example, the smallest active region in a production-type transistor, the base thickness of the 2N128 surface-barrier transistor, is one ten-thousandth of an inch. The smallest corresponding region of a triode vacuum tube, the grid-cathode spacing in the microwave triode used in the transcontinental television network, is six times as great.

The high-frequency trouble with transistors is not dimensions. Rather, it is the principle underlying the motion of the charges within the transistor. In vacuum tubes, the charge carriers (electrons) move through a *vacuum* within which the applied electric field is fully effective in accelerating them to high speed. Moreover, since a good vacuum is an excellent insulator, high voltages may be applied to achieve phenomenally rapid transits of the electrons.

In present-day transistors, however, the useful charge carriers are interspersed among a vast multitude of non-useful charges of the opposite sign, and the presence of the latter charges prevents the applied voltage from producing an electric field within the body of the transistor. Consequently, the useful carriers, instead of being whisked across the transistor by electric attraction, move in the relatively aimless motion of diffusion, like people drifting away from an overcrowded room.

This relatively slow motion is a block on the road to higher-frequency performance of transistors, and it far outweighs the effect of small dimensions. For example, the surface-barrier transistor mentioned is a 50-megacycle unit; the microwave triode with a sixfold *disadvantage* in dimensions, achieves an 80-fold *advantage* in frequency. It works nicely as an rf amplifier at 4,000 megacycles, higher than any transistor has been even rumored to operate to date.

We know the general path around this slow-motion roadblock. This is to remove, partially or wholly, the nonuseful majority carriers which dissipate the applied electric field, which can be done by changing the distribution of impurities in the semiconductor in a known manner. This is not to say that the work on transistors using intrinsic materials or graded distributions of impurities has gone very far in a production sense. It merely points the way to be followed if the transistor is to imitate the charge-motion patterns of vacuum tubes.

It must be emphasized that the diffusion mode of motion, typical of transistors available today, is a real advantage, when the frequency of application lies in the range up to 100 megacycles which such transistors now cover. This advantage lies in the fact that *low voltage* sources suffice to operate these transistors. And low voltage, particularly in battery-operated equipment, is an important practical consideration.

On still another front, tubes and transistors are different breeds of cat. This is high-power operation. High power in an electronic device means high current, which requires a large cathode and a large plate; and it means high voltage, which requires good insulation. Neither requirement is any particular hindrance to the vacuum tube designer. In fact, vacuum-tube cathodes can be designed larger and larger almost without limit; the filament current of the 5831 super-power beam triode, previously mentioned, is 2,220 amperes; the filament heating power is 13 kilowatts, and the typical plate current is 41 amperes.

To get rid of the heat generated in a high-power device, it must be big enough to present a large surface to the cooling medium. Large size is also dictated by the need to get sufficient insulation to permit high voltages to be applied. Large dimensions work against high-frequency operation. For many years the size of high-power tubes limited the amount of power that could be generated at such frequencies as, say, 1,000 megacycles. But the technique of compartmentalizing the electrons into groups, as in the klystron, cavity magnetron, and traveling wave tube, has changed all this. Today, advanced versions of the klystron will generate 10,000 watts continuously at 1,000 megacycles, and cavity magnetrons will generate brief peaks in the order of hundreds of kilowatts in the range of 10,000 megacycles and above.

Where does the transistor stand in this situation? The part of a transistor which corresponds to the vacuum tube cathode is its emitter. The semiconductor emitter has the important advantage, relative to a cathode, of being able to produce more carriers per square centimeter.

But as of the moment, the transistor designer is having a very difficult time designing large emitters, even though his brother in the high-power semiconductor diode department is making some startling breakthroughs in this region. The semiconductor surface on which the transistor emitter is formed must be perfection itself. If any crystal defect or other inhomogeneity is included within the emitter, the number of useful minority carriers emitted falls away all out of proportion.

The way out here is also clear in principle but fraught with practical difficulties. We must seek a wholly new order of control in the metallurgical preparation of semiconducting materials, which is already the most skilled art that metallurgy has to offer. We must know, also, a lot more about the surface properties of the semiconductors under the emitter electrode and around its edges to prevent a fatal loss of useful carriers, even when the metal itself has been urged to perfection.

Sometime we will, no doubt, know how to design and produce an emitter of any required size. If, by then, we also know how to apply high voltages to the material without burning it out at hot spots or fracturing it from electric stress, we will have the basis of transistor design worth the name "high power" as it is used in the tube laboratories.

Such structures will have to be large to get rid of heat losses, and this means that the charge-carrier grouping principle will have to be introduced to the transistor if it is to achieve both high power *and* high frequency. At present there is no analog to electron grouping in transistors. But if we arrange that the useful carriers are in the majority, rather than in the minority as at present, there seems no fundamental bar to charge grouping in transistors. At this stage, the really high-power, high-frequency transistor becomes a possibility.

All this will, of course, take a lot of effort and it may well be asked, "Why bother?" After all, these high-power, high-frequency jobs are being very well handled by vacuum tubes right now, Why insist on a transistorized version? Transistors in the kilowatt class won't be battery operated. Moreover, they are likely to contain a large amount of highly refined silicon, and, hence, are likely to be rather expensive.

Should we take on the massive labor of improving transistors until they compete on every front with tubes? From a long-range point of view, perhaps so; after all, we can't stop, and we don't want to stop, the steady extension of the transistor principle, nor the steady extension of the vacuum tube principle, to frontiers not now in sight. But there is a great deal to be said for another shorter-range point of view, extending perhaps only over the next five years. During that period, it would appear prudent to emphasize that we have, in tubes and transistors, two quite different electronic citizens, capable of holding down quite different jobs, and we might do very well to concentrate on putting them to the most effective uses, not in competition, but as members of a team.

It is appropriate, therefore, to recommend that we pay particular attention to the special abilities of transistors in the areas of *immediate application*. The justification for this goes back to the matter of life expectancy previously noted for its importance in large assemblies of electronic devices, such as digital computers· To this we must add the importance of long life in smaller assemblies, where safety of life and effectiveness in military operations depend on completely *dependable* operation, unfettered by a statistical expectation of failure.

On this question of life expectancy we have some important new evidence. In at least two laboratories life tests of certain types of transistors over periods in excess of 10,000 hours have shown such a phenomenally low failure rate that the extrapolated life expectancy, using statistics applicable to vacuum tubes, exceeds *one million hours*, a period extending from now to 2069 A.D.! It must be admitted that the exponential failure rates assumed in these statistics may not (in fact, probably do not) apply to transistors, and that much more life testing must be completed before the million-hour figure can be accepted with the same assurance that now surrounds the life expectancy of the 40-year (350,000-hour) repeater tubes, now being laid in cable repeaters at the bottom of the Atlantic. But the evidence is so startling that even conservative forecasters are willing to concede that transistor life is on a new plateau, far above that of vacuum tubes. In fact, with every passing day, the conviction grows that transistors, properly made and properly applied, can outlast just about any garden-variety component used in present-day electronic equipment, except possibly the chassis base.

We have become used to the fact that the vacuum tube is the weak sister in just about every assemblage of electronic components. We accept the fact that tubes habitually fail in service, at times which can be predicted only statistically and which give no assurance, for example, that the autopilot *definitely* will be working during the next hour. We go along with the fact that tubes are the most delicate item on the chassis deck. And we forget that it's an unusual tube whose technical capabilities, in power or frequency, fully match up to those of the components with which it is connected. We excuse all this by saying, "The tube is the active element; it has the toughest job; to make it an equal partner with every resistor, condenser and coil is just not economical."

But, before the electronic industry will have reached maturity we must accept, as a categorical imperative, that electron devices, tubes and transistors, must be equal partners with other components, equal in life, equal in ruggedness, equal in sharing the load.

In this department, the transistor is taking the lead. Transistors are rugged, naturally so. And long life in transistors, given a certain amount of essential know-how, comes relatively easy. Tubes are not naturally rugged; ruggedness must be developed. And they are not naturally long-lived.

The fundamental limit to vacuum tube life, assuming it doesn't meet an accident or suffer a mechanical failure, is exhaustion of electron emission at the cathode.

Electron emission depends not only on a steady flow of electrons to the cathode from the external circuit, but also on the presence of very special conditions at the cathode surface, including impurity elements which lower the work function and allow the electrons to escape. The surface condition appears, by every evidence gathered over nearly fifty years, to be subject to inevitable deterioration as the impurity elements boil off or otherwise escape. And there is no sure mechanism for putting them back. So, sooner or later, the emission falls off and the tube goes dead. Much can be done to arrest the deterioration, as in the case of the 40-year submarine cable repeater tube. But the life expectancy of this tube (about 0.02 per cent failures at 1,000 hours) was achieved *in spite of* the fact that it uses cathode emission, not *because* of it. As a result, the tube is used so far under normal ratings, and is so painstakingly put together that its use can be justified only in very special circumstances.

Contrast this with the "million-hour transistor" postulated above. It achieves long life expectancy without overdesign or costly production methods. This is possible because the charge emission process in a transistor is fundamentally different from that in a vacuum tube. The transistor emitter is self-replenishing, indefinitely. Transistors do burn out, of course, for a large number of other reasons; and they do lose their amplifying function if they are overheated. But they do not fail due to exhaustion. Tubes do.

At the moment this difference appears to be fundamentally rooted in the principle of operation of the two devices. If further investigation confirms this view, we may then be sure that long life (which means unvarying ability to amplify) will always be easier to get in a transistor than in a vacuum tube. The conclusion then is evident. We should use transistors, now, where long life and ruggedness is important. We should use tubes in the many areas where transistors, for the present anyway, can't handle the job.

Perhaps, then, we can conclude with the observation that the two devices have a lot to learn, one from the other. In such a situation, fast and easy communication from one group of technical workers to another is essential to rapid progress. The Professional Group on Electron Devices, a single group having cognizance of both devices, is in an ideal position to foster this communication through its conferences, its TRANSACTIONS, and through the PROCEEDINGS OF THE IRE.

The Cryotron—A Superconductive Computer Component*

D. A. BUCK†

Summary—**The study of nonlinearities in nature suitable for computer use has led to the cryotron, a device based on the destruction of superconductivity by a magnetic field. The cryotron, in its simplest form, consists of a straight piece of wire about one inch long with a single-layer control winding wound over it. Current in the control winding creates a magnetic field which causes the central wire to change from its superconducting state to its normal state. The device has current gain, that is, a small current can control a larger current; it has power gain so that cryotrons can be interconnected in logical networks as active elements. The device is also small, light, easily fabricated, and dissipates very little power.**

The Cryotron Principle

Before describing the cryotron as a circuit element and potential computer component, the basic physical phenomena underlying its operation will be described.

Superconductivity

Superconductivity was discovered in 1911 by H. Kammerlingh Onnes at Leiden, three years after he succeeded in liquifying helium. While extending electrical resistance measurements to this new low-temperature region he found that the resistance of mercury drops suddenly to zero at 4.12°K. Soon many other materials were shown to display this same unusual behavior. Niobium becomes a superconductor at 8°K, lead at 7.2°K, vanadium at 5.1°K, tantalum at 4.4°K, tin at 3.7°K, aluminum at 1.2°K, and titanium at 0.5°K. In addition to 21 elements, many alloys and compounds are superconductors with transition temperatures ranging between 0 and 17°K.[1,2]

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[1] D. Schoenberg, "Superconductivity," Cambridge University Press, Cambridge, England; 1952.

[2] F. London, "Superfluids," John Wiley & Sons, Inc., New York, N. Y., vol. 1; 1950.

The resistivity of many superconductive materials is relatively high at room temperature, especially those which have high transition temperatures, such as niobium, lead, tantalum, etc. It is interesting that relatively poor conductors become superconductors at low temperatures whereas good conductors such as gold, silver, and copper do not. The resistivity of superconductive materials drops as they are cooled. Just above their superconductive transition, the resistivity is between 10^{-1} and 10^{-3} of their room temperature resistivity, depending on the purity and mechanical strain in a particular sample.

Below the superconductive transition the resistivity is exactly zero. That it is truly zero is vividly demonstrated by an experiment now in progress by Professor S. C. Collins at M.I.T. wherein a lead ring has been carrying an induced current of several hundred amperes since March 16, 1954, without any observable change in the magnitude of the current.

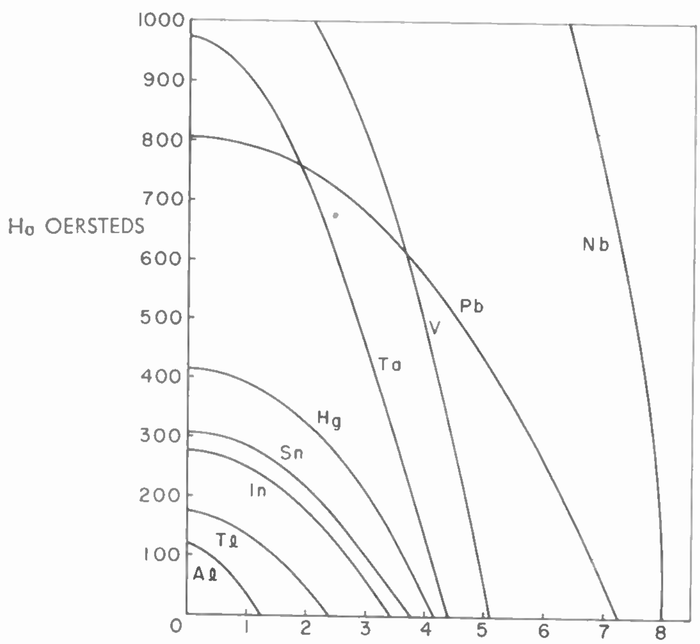


Fig. 1—Threshold magnetic field vs temperature for several common superconductors.

Destruction of Superconductivity by a Magnetic Field

The foregoing discussion of the superconductive transition is valid only in zero magnetic field. With a magnetic field applied, the onset of superconductivity occurs at a lower temperature. If the intensity of the magnetic field is increased, the transition temperature is still lower. A plot of the transition temperature as a function of the applied magnetic field is more or less parabolic in shape, levelling out as absolute zero is approached. Such a plot for several common elements is given in Fig. 1.

If the temperature is held below the transition temperature for one of these materials, the resistance of that material is zero. Its resistance will remain zero as a magnetic field is applied until that magnetic field reaches a critical value. Above this value the normal resistance returns. If the field is lowered, the resistance disappears.

Raising and lowering the magnetic field thus controls the resistance of the material in the magnetic field by causing it to shift from its superconducting state to its normal state and back without changing the temperature. In Fig. 2, this operation corresponds to moving up and down on a vertical (constant-temperature) line which has its lower end in the superconducting region and its upper end in the normal region. If the operating line is moved to a lower temperature, the magnetic field required to reach the normal region is greater. For each of the materials which becomes superconducting, there is a temperature about 0.2°K below the zero-field transition which allows operation with rather small magnetic fields—between 50 and 100 oersteds. For lead, this temperature is about 7.0°K, for tantalum about 4.2°K, for tin about 3.5°K, for aluminum about 1.0°K.

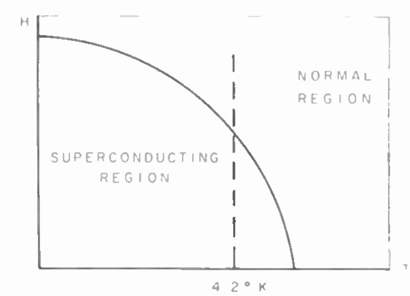


Fig. 2—Threshold magnetic field as a function of temperature for a superconductor.

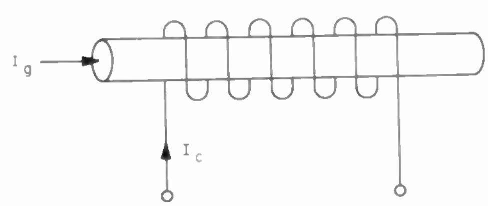
Tantalum has been used in many of the early experiments at M.I.T. because 4.2°K is the boiling point of helium at a pressure of 1 atmosphere and therefore the temperature of most storage tanks for liquid helium. Higher temperatures (up to 5.2°K) involve raising the pressure on the liquid helium bath; lower temperatures (down to about 1.0°K) involve lowering the pressure. At 4.2°K, then, experiments do not involve sealing of the lead-in wires.

In a typical cryotron, the resistance being controlled is in the form of a straight piece of wire about 1 inch in length. The magnetic control field is generated by current in a single-layer winding which is wound over the central wire (Fig. 3, top). The central wire is analogous to the plate circuit of a vacuum tube and the control winding is analogous to the control grid. In this case, the plate resistance is zero in the cutoff region and rises rapidly as grid-current cutoff is reached.

Superconducting Control Winding

The control winding is made of a superconducting wire which has a relatively high transition temperature. Niobium (formerly called columbium) is used because it has a very high transition temperature and can be drawn into fine wire which is strong. Lead or lead-plated wire is a second possible control-winding material.

At the temperature used, the control winding remains a superconductor at all times, and would remain so even in magnetic fields much higher than those being used to control the central wire. Therefore, there is no resistance in the control winding. A magnetic field, once established, needs no further energy for its support; the control current is maintained against zero back voltage. Similarly, all interconnecting wire is also superconducting.



Single Cryotron

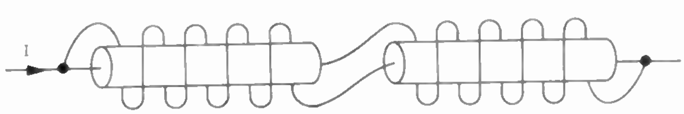


Fig. 3—Cryotron bistable element (flip-flop).

The Cryotron as a Device

Static Characteristics

The resistance of the central wire of a typical cryotron is plotted as a function of current in the control winding in Fig. 4. The central wire is called the gate circuit. This particular cryotron is made by winding a single layer of 0.003-inch insulated niobium wire over 0.009-inch bare tantalum wire. The insulation on the niobium is heavy Formvar. The finished winding has 250 turns per inch. In the midportion of the winding, the magnetic field due to a current is 124 oersteds per ampere. When the control current of Fig. 4 is translated into magnetic field intensity, the highest transition field is seen to be about 40 oersteds.

As current in the tantalum gate circuit is increased, the transition control current becomes lower. This effect is due to the additional magnetic field at the surface of the tantalum wire created by the gate current. This field, commonly called the self-field of the wire, limits the amount of current which can be carried by a superconductor. The effect was, in fact, discovered shortly after the discovery of superconductivity when Onnes and his coworkers (1913) tried to make a powerful electromagnet out of their newly discovered zero-resistance materials. When the current in their superconducting solenoid reached a certain critical value, its resistance suddenly reappeared. When the discovery was made that magnetic fields cause restoration of resistance, it was quickly seen by Silsbee (1916) that the limit on the current that can be carried by a superconductor is due to the magnetic field created by that current. The magnetic field, H, at the surface is given by:

$$H = \frac{I}{\pi d}$$

where

H is in ampere-turns per meter
I is in amperes
d is in meters.

If H is given in oersteds, I in amperes, and d in mils (thousandths of an inch) this becomes

$$\underset{\text{oersteds}}{H} = \frac{175.5}{d \text{ mils}} I.$$

It will be noted that the transition characteristics are very sharp for high gate currents. The additional sharpness is a peculiarity of the measuring technique wherein a current is passed through the gate circuit and the voltage across the gate circuit is measured. When resistance suddenly appears, I^2R loss in the gate circuit causes heating which lowers the critical field and speeds switching.

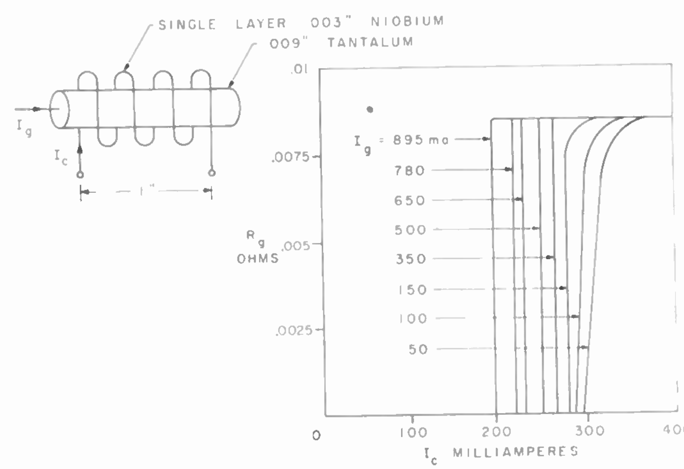


Fig. 4—Cryotron gate resistance vs control current.

The magnetic field due to the control winding is along the axis of the central wire while the self-field of the wire due to its own current is tangential to the wire. The two fields thus add in quadrature and the resulting net field is the vector sum of two fields. Results indicate that the superconducting central wire reaches its critical field when the net field reaches a certain value, regardless of which way the net field points in relation to the center line of the wire. In one experiment, the curves of Fig. 4 were reproduced exactly for all four combinations of positive and negative control and gate current. There is no reason to suspect that there would be anisotropy in the critical field for different orientations of the net field with respect to the wire axis as long as the control field is longitudinal, especially since the wire is polycrystalline. Thus the cryotron has an interesting property as a circuit element. Control is independent of the sign of the control current—it depends only on the magnitude. Furthermore, when the gate circuit is ON, that is, in its superconducting state, current can flow in either direction, unlike a vacuum tube which can pass current only in one direction.

Current Gain

Because the two fields add in quadrature, the self-field of the wire has less effect on the threshold control current at low gate current than it does at high gate current. The locus of threshold control current points as a function of gate current is an ellipse. The ratio of major axis to minor axis of the ellipse is the ratio of the magnetic field produced by a current in the control winding to that produced by the same current in the gate circuit. This ratio is also called the current gain of the cryotron. If the current gain were less than unity, it would not be possible to control one cryotron with an identical cryotron because more current would be required to bring the second cryotron to its control-current threshold than the first cryotron could handle through its superconducting gate circuit.

The control field is related to the control current by the number of turns per inch in the control winding, and the self-field is related to the gate current by the diameter of wire used in the gate circuit. Current gain, K, is simply given by:

$$K = \pi d \frac{N}{L}.$$

For a given pitch control winding and a given gate wire diameter, the current gain is specified. Fig. 5 is a plot of lines of constant K as a function of winding pitch and gate wire diameter. For the cryotron whose characteristics are plotted in Fig. 4, $K=7$. The current gain actually observed for a given cryotron is often less than calculated, presumably due to the constriction of supercurrents by small normal regions which nucleate about flaws in the wire surface. Control-current threshold points thus form a locus in the gate current-control current plane which lies on an ellipse of smaller major-to-minor axis ratio.

Power Gain and L/R Time Constant

The input power to a cryotron, exclusive of eddy current and relaxation losses, is the product of the energy stored in the magnetic field of the control winding and the frequency at which the control winding is energized:

$$P_{in} = \frac{fL_cI_c^2}{2}.$$

The input power is reactive. In an oscillator circuit the input inductance can be resonated with a linear capacitor to minimize input losses. In computer pulse circuitry, however, the control windings are untuned. The entire amount of power is therefore dissipated.

The output power of a cryotron can be approximated as follows: Consider a cryotron amplifier delivering square waves of equal on and off periods to a resistive load. The gate circuit shunts the current when superconducting and allows part of it to flow through the load when normal. Maximum power transfer occurs when the load resistance, R_L, is made equal to the normal resistance of the gate circuit R_g. Under this condition, average load power is given by:

$$P_{ave} = \frac{I_g^2R_L}{2}.$$

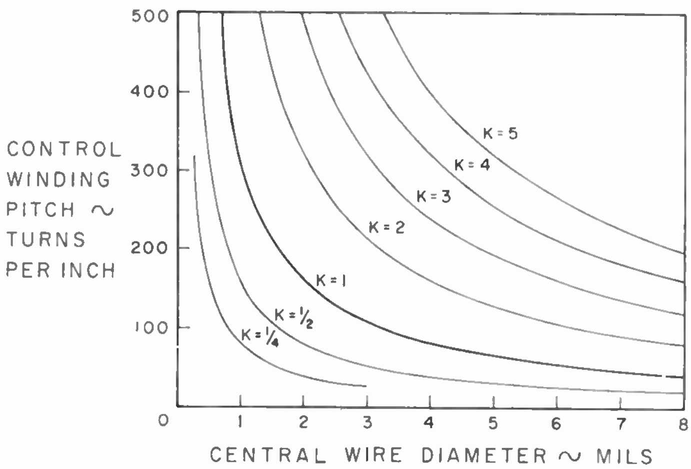


Fig. 5—Current gain vs control winding pitch and central wire diameter.

Power gain, G, can be approximated by:

$$G = \frac{\text{power out}}{\text{power in}} = \frac{I_g^2R_L}{fL_cI_c^2} = \frac{1}{f}\left(\frac{I_g}{I_c}\right)^2 \frac{R_g}{L_c}.$$

In the pulse circuits of the following section on cryotron computer circuitry, the gate current of one cryotron becomes the control current of another; $I_g = I_c$. For this condition, the frequency at which the power gain becomes unity is:

$$f_{max} = \frac{R_g}{L_c}$$

which is the reciprocal of the L/R time constant of the circuit. The L and R are on different cryotrons, but since large numbers of identical cryotrons are involved, one can speak of the L/R time constant of a given cryotron as being the fundamental time constant of the circuitry.

If a given cryotron is made longer while holding the pitch of the control winding constant, the resistance and inductance increase together such that the L/R time constant is not affected. The L/R time constant is thus independent of cryotron length.

If the diameter of a given cryotron is made smaller while holding the pitch of the control winding constant, the resistance increases inversely as the diameter squared, while the inductance decreases directly as the diameter squared. The L/R time constant thus de-decreases as the fourth power of the diameter.

The current gain of the cryotron drops if the diameter is made smaller while holding the pitch of the control winding constant because the current-carrying capacity of the gate circuit decreases directly with the diameter. If the current gain is to be held constant by increasing the pitch of the control winding proportionately as the diameter is made smaller, the inductance remains constant and the L/R time constant decreases as the square of the diameter. One thus pays rather dearly for current gain. Cryotron computer circuitry, discussed below, is operated with a minimum of excess current gain.

The resistivity of the normal state varies over several powers of ten among the various superconductors. The L/R time constant varies inversely as the resistivity. An increase in speed of circuit operation can therefore be achieved by alloying superconductors to increase resistivity.[3]

Circuit speed can also be increased by using a hollow central wire. Superconductivity is a skin effect, penetrating but a few hundred atom layers, and therefore the core of a wire can be removed and the wire will still have zero resistance in its superconducting state. The resistance in the normal state, however, will be higher by the ratio of the original cross sectional area to the new cross sectional area. The core need not actually be removed, provided it is made to have a relatively high resistivity. Wire with a high resistivity core and a superconducting shell can be fabricated by vapor plating.

Eddy Currents

It has been shown by Faber[4] that the delay τ, due to eddy currents in the destruction of superconductivity of a wire by a longitudinal magnetic field is:

$$\tau_e = \text{const}\,\frac{\mu d^2 H}{\rho(H - H_c)}$$

where H is the external magnetic field, H_c is the threshold magnetic field and ρ is the resistivity. The switching time varies directly as the square of the diameter and inversely as the resistivity, and is a function of the amount by which the threshold magnetic field is exceeded.

As the circuits of the following section are speeded up by making cryotron diameters smaller, there will be a speed range where eddy currents become important. Lowering the diameter still further and increasing the pitch proportionally should then increase the speed as the inverse square of the diameter, since both circuit L/R time constants and eddy current time constants decrease proportionally. The observed time constants of the free-running multivibrator of the next section are of the same order of magnitude as the calculated L/R circuit time constants. Eddy current effects should become important during the next order of magnitude increase in speed.

The transition from normal to superconductor also involves delays and a somewhat different switching mechanism.[5,6] A supercooling effect is important. A nucleus of superconducting material forms at one spot on the wire surface, sweeps around the wire, and then grows along the wire. Extrapolation of slow velocity data on tin rods in fields just barely below the threshold field indicate that in cryotron operation, velocities of the order of tens of centimeters per microsecond ought to be encountered with a current gain of two. As soon as a superconducting path is established over the surface of the wire, the cryotron is in its superconducting state —even if the center of the wire requires additional time to become superconducting. While it is not anticipated that this transition will be a major source of delay, it is interesting to note that this delay is one which depends on the length of the cryotron.

As circuit speeds are increased by increasing the resistance of the central wire, thereby shortening L/R circuit time constants and minimizing eddy current effects, a fundamental limit to the ultimate speed exists in the form of relaxation losses. The exact frequency region in which these losses will become predominant is not known, but from experiments with superconducting coaxial cable and waveguide resonators, an estimate is available which places the limit between 100 and 1,000 megacycles.

Cryotron Computer Circuitry

The low impedance level of cryotron circuitry dictates a high-impedance power supply (current source) with circuit elements connected in series. Each element allows the current a choice among two or more paths only one of which is superconducting; all of the current flows through the superconducting path. The current encounters zero back voltage except when the paths are changing. The standby power is therefore zero. Several circuits, representative of those found in digital computers, are described below.

Flip-Flop

A bistable element, one of the most common in a digital computer, can be made using two cryotrons. The two gate circuits are each in series with the control winding on the other and the two paths are in parallel (Fig. 3,

[3] B. Serin, "The Magnetic Threshold Curve of Superconductors," ch. VII, "Progress in Low-Temperature Physics," edited by C. J. Gorter, Interscience Publishers, New York, N. Y.; 1955.

[4] T. E. Faber, "The phase transition in superconductors II. Phase propagation above the critical field," *Proc. Roy. Soc. A*, vol. 219, pp. 75–88; 1953.

[5] A. B. Pippard, "Kinetics of the phase transition in superconductors," *Phil. Mag.*, vol. 41, p. 243; 1950.

[6] T. E. Faber, "The phase transition in superconductors III. Phase propagation below the critical field," *Proc. Roy. Soc. A*, vol. 223, pp. 174–194; 1954

bottom). If the current is established in one of the two paths, that current makes the alternate path resistive. Current in one path, once established, will therefore continue to flow indefinitely in that path.

Two additional cryotrons can be added to the circuit, one in series with each branch, in order to place the flip-flop in the desired state. A pulse on one of the two input cryotrons momentarily places a resistance in that side. Both sides are then resistive, and the current divides between them. If the power supply current is not larger than twice the critical current of the cryotrons, both sides of the flip-flop will become superconducting. One side of the flip-flop has a resistance inserted by the input cryotron, however, and the current thus chooses the other side. Once the current builds up in the other side, it makes the side on which the input cryotron is being pulsed resistive, and therefore the pulse in the control winding of the input cryotron can be removed; the current will continue in the new path.

Two more cryotrons can be added to the circuit for sensing the state of the flip-flop. Placed with their control windings each in series with one of the two sides of the flip-flop, one of the read-out cryotrons is resistive and the other is superconductive. The gate circuits are joined and a read-out current pulse is applied at the junction. The read-out pulse will choose one path or the other, depending on the state of the flip-flop. The flip-flop with read-in and read-out cryotrons is shown in Fig. 6.

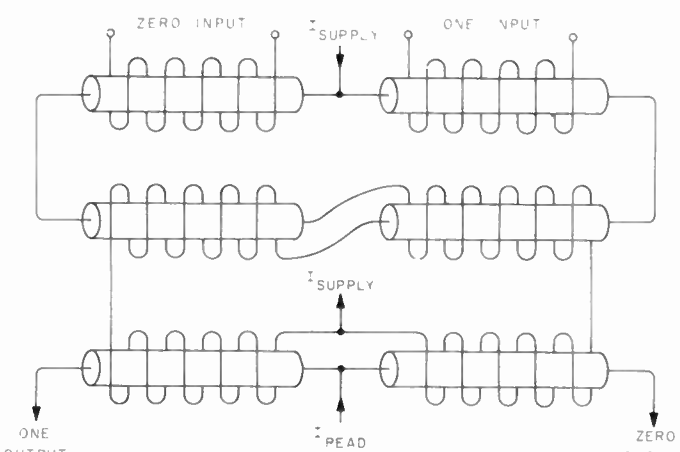


Fig. 6—Cryotron flip-flop with read-in cryotrons and read-out cryotrons.

Any number of input cryotrons can be added in series with those already described (Fig. 7) to set the flip-flop to one state or the other. Connected as such, they are OR gates; any one of them acting alone can set the flip-flop. Similarly, additional cryotrons can be added with their gate circuits in parallel with the control winding of the input cryotron already described, behaving as AND gates (Fig. 8). The flip-flop set current is bypassed through one or more of these parallel gates unless all of them are resistive. This latter connection involves superconductors in parallel, in which case the current divides inversely as the inductance of the parallel paths.

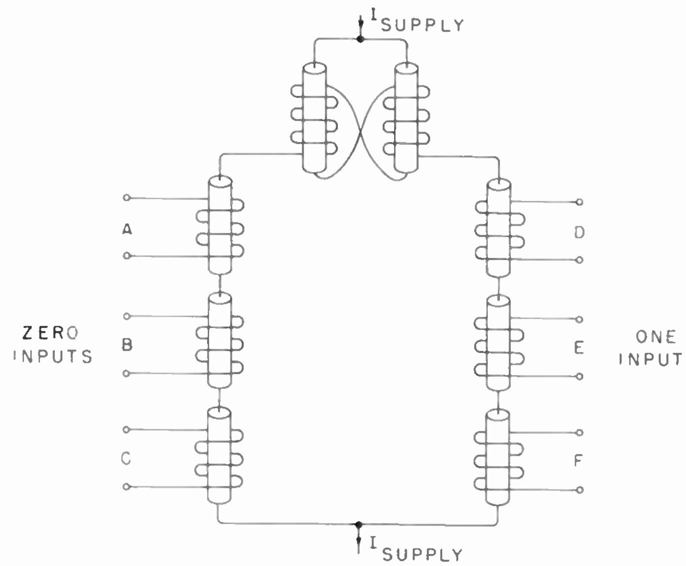


Fig. 7—Cryotron flip-flop with OR gates in both sides.

Additional read-out cryotrons can be added in series with those already described. Since their control windings are superconducting, the additional cryotrons do not add any resistance to the flip-flop. The additional inductance increases the L/R time constant of the circuit, however, lengthening the transition time between states.

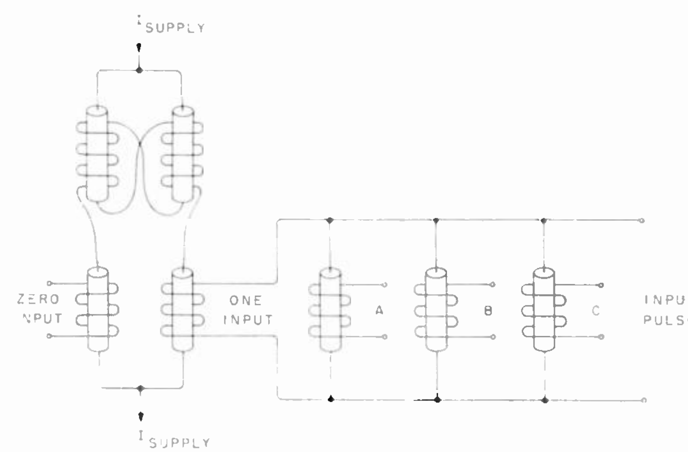


Fig. 8—Cryotron flip-flop with AND gates in one side.

Multivibrator

Three flip-flops made of one-inch pieces of the cryotron stock whose characteristics are given by Fig. 4 have been studied in a multi-vibrating circuit (Fig. 9). The read-out cryotrons of flip-flop *A* are connected in

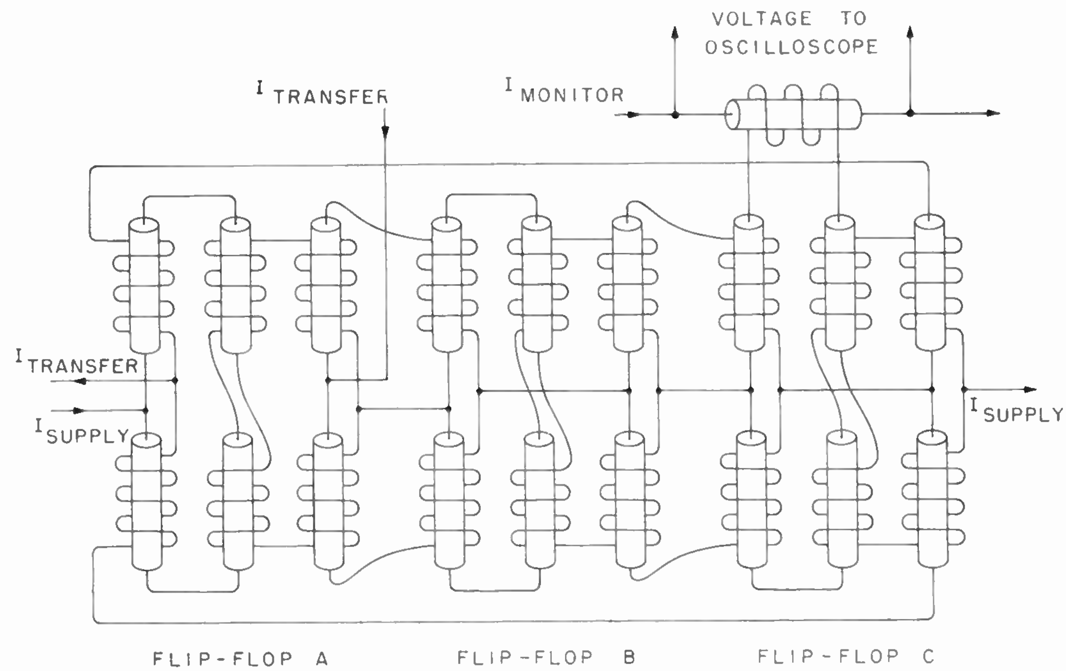


Fig. 9—3-flip-flop cryotron multivibrator.

TABLE I

SEQUENCE OF MULTIVIBRATOR FLIP-FLOP STATES

	Time Period								
Flip-flop	0	1	2	3	4	5	0	1	2
A	0	0	0	1	1	1	0	0	etc.
B	0	1	1	1	0	0	0	1	etc.
C	1	1	0	0	0	1	1	1	etc.

such a way as to set flip-flop *B* to the state opposite that of *A*. A similar connection between *B* and *C* causes *C* to assume the state opposite to that of *B*, and a similar connection between *C* and *A* causes *A* to assume a state opposite to that of *C*. Since there are an odd number of stages, the ensemble free-runs through the sequence given in Table I.

ZERO is defined as conduction through the upper cryotron of the flip-flop pair and ONE is defined as conduction through the lower.

The time taken for transition from one time period to the next is a function of the transfer current. If transition occurs at a fixed threshold current value, the final value of the rising current in a given control winding determines the fraction of the L/R time constant required to reach that threshold value. If the final value is (a) times the threshold value, the time required to reach the threshold is given by: $t = L/R \ln (a/a-1)$. The particular multivibrator circuit described completes the round-trip through its six time periods at the rate of 100 to 1,000 times per second depending on transfer current. The higher frequency gives individual time periods of 167 microseconds duration.

To monitor the transitions of one of the flip-flops, an additional cryotron gate is added with its control winding in series with one side of the flip-flop. A current source is connected to its gate circuit. When the control current is zero, the gate circuit is a superconductor and the voltage is zero. When the control current reaches the threshold value, the gate circuit becomes resistive and develops a voltage which is amplified and displayed. Typical values are: $R = 0.01$ ohm, $I = 100$ ma; $V = 1$ millivolt. The true current waveform is not preserved by the monitoring gate due to its inherent nonlinearity plus the sharpening of its transition due to I^2R heating as it becomes resistive.

Multiterminal Switch

Distributing a pulse among several wires can be accomplished by a cryotron switch (Fig. 10). Information is fed into the switch from cryotron flip-flops, here represented by toggle switches. One flip-flop causes the odd or even rows of the switch to be resistive, a second flip-flop causes odd or even pairs to be resistive, a third flip-flop causes odd or even fours to be resistive, and so on. A single path survives as a superconductor, and all of the read current follows that path and thence to the load. With the flip-flops set as shown with binary input 101, row 5 is selected. This particular switch can thus be used as a binary-to-octal converter.

Binary Adder

The principles embodied in the flip-flop and switch can be used to design the stages of a binary adder. The task to be done by each stage is represented by Table II.

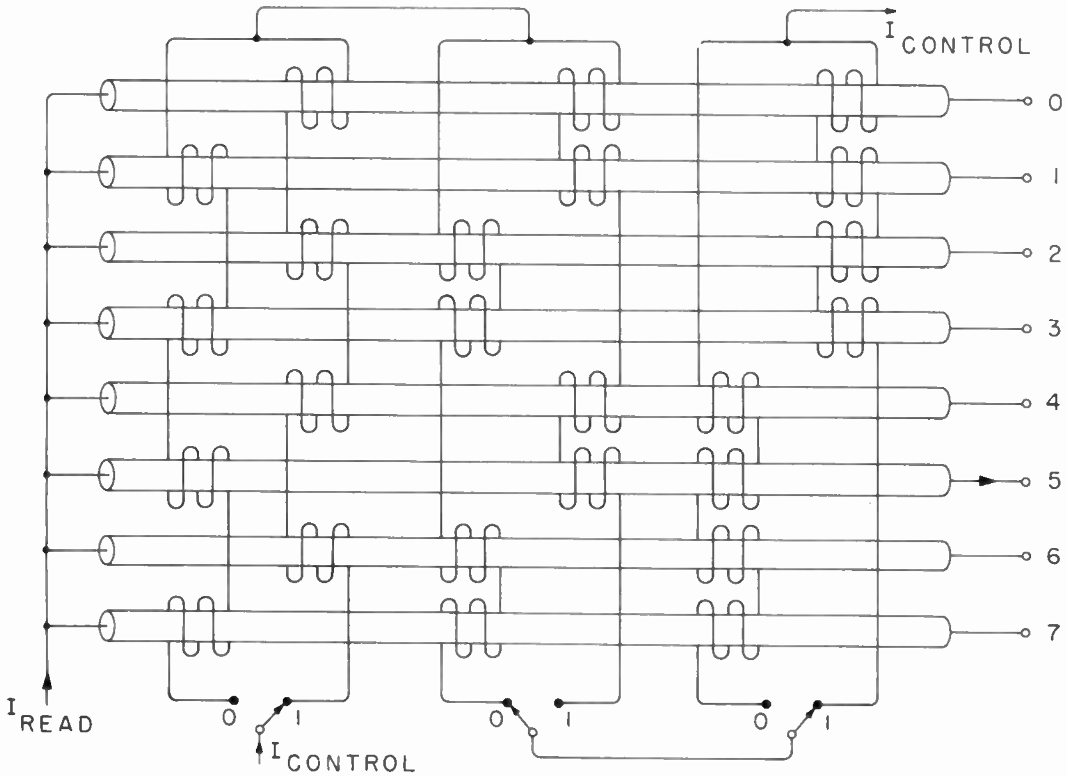


Fig. 10—8-position cryotron switch.

TABLE II
BINARY ADDITION TABLE

Input A	Input B	Carry In	Sum	Carry Out
1	1	1	1	1
0	1	1	0	1
1	0	1	0	1
0	0	1	1	0
1	1	0	0	1
0	1	0	1	0
1	0	0	1	0
0	0	0	0	0

The (n)th digits of the two numbers to be added are combined with the carry from the $(n-1)$th stage to form the (n)th digit of the sum and the carry to the $(n+1)$th stage. Since there are eight possible combinations of the three inputs, an accumulator design can center about the eight-position switch already described. The three inputs operate the control windings and the eight output leads, one of which carries current, can actuate cryotron gate circuits which set up paths to determine sum and carry digits. Fig. 11 shows such a stage. The carry input actuates either the upper four rows or lower four rows of the switch, thus eliminating one of the control-winding pairs. The eight gates which operate the sum flip-flop are connected with four in series in each of two parallel paths. The element which is caused by the switch to be resistive diverts the current to the path opposite itself setting the sum flip-flop to its proper state. A similar group of gates develops the carry for the following stage. Note that all circuits are in series from a current source power supply.

The foregoing binary adder design is described to illustrate the way in which switches and gates can be interconnected. A design having fewer cryotrons per stage is available wherein the carry is handled by a lattice network shown in Fig. 12. The label beside each of the six control windings indicates when it is to be energized. The $A=B=0$ and $A=B=1$ windings can each be made of two cryotrons in a parallel AND circuit and then directly excited from the A and B flip-flops, or the current necessary to excite them can be derived from a four-position cryotron switch. The latter method has an advantage since $A=B$ and $A\neq B$ currents are useful in forming the sum digit. After the current has passed through the $A=B=1$ and $A=B=0$ coils in the carry network, the two coil ends can be combined to provide a current $A=B$. This involves the outputs of two of the four output terminals of the 4-position switch. The other two output terminals can be tied together directly to provide a current if $A\neq B$. The sum digit can then be simply formed in the following way.

Note that the sum is ONE if $A=B$ and the carry in is ONE; ZERO if $A=B$ and the carry in is ZERO; ONE if $A\neq B$ and the carry in is ZERO; ZERO if $A\neq B$ and the carry in is ONE. The $A=B$ and $A\neq B$ currents can therefore be used to route the carry input currents to the proper side of the sum flip-flop. Fig. 13 is a schematic

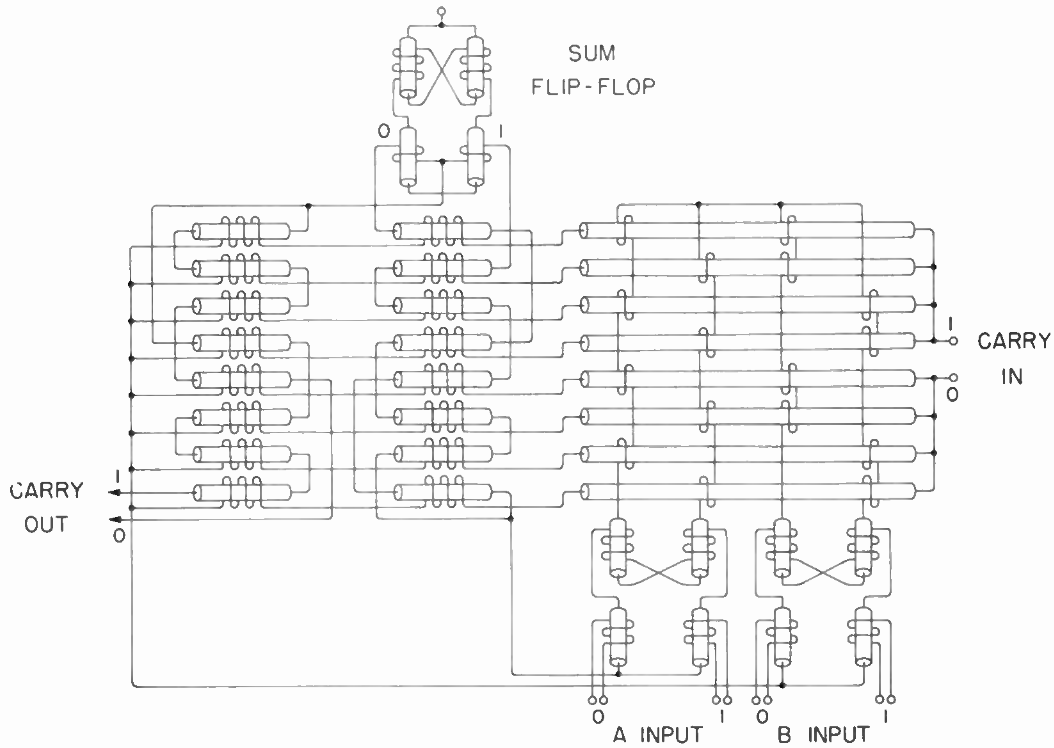


Fig. 11—One stage of a binary adder.

drawing of such a stage of an accumulator, abbreviated in that the A and B flip-flops are not shown, nor is the transfer link from the sum flip-flop back to the A flip-flop shown (used during accumulative addition and subtraction). In this design one notices the convenience of interconnecting cryotrons without regard to dc levels, very much as relay contacts are placed in relay computer circuitry.

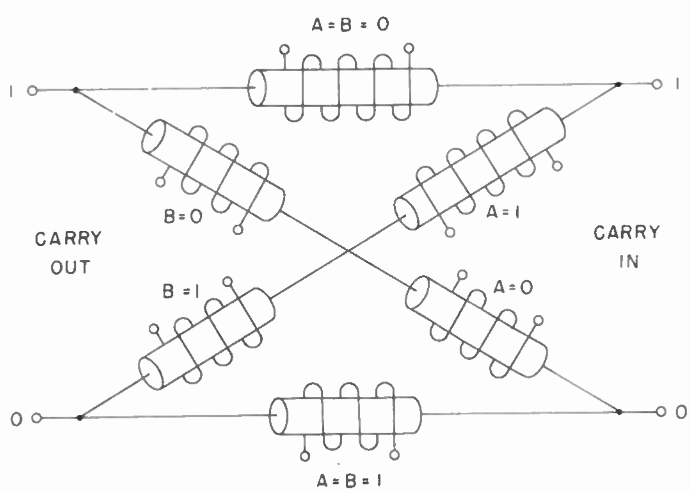


Fig. 12—Carry network.

Stepping Register

Stepping registers are commonly used for receiving digital information in serial form at one pulse repetition frequency and after a predetermined number of binary bits have been stored, shifting the information out at a different frequency. A second common use for shifting registers is to accomplish the conversion between digital information in serial and parallel form. The stepping registers in common use are made of vacuum tubes, transistors, or magnetic cores. Cryotrons can also be used in the same service. Each stage of the shift register consists of two cryotron flip-flops with read-in and read-out cryotrons. One transfer circuit sets the second of the two flip-flops of each stage to correspond to the state opposite that of the first. The coupling link to accomplish this is similar to the one described in the above section discussing the multivibrator, which interconnects stages of the multivibrator. A second transfer circuit sets the first flip-flop of each stage to correspond to the state opposite that of the second flip-flop of the stage to its left. A line of such stages serves as a shifting register, capable of shifting digital information to the right. Information (ONE's or ZERO's) fed into the first flip-flop in synchronism with the second of the two transfer pulses (called ADVANCE B pulse), will advance through the stepping register one stage for each pair of transfer pulses, ADVANCE A and ADVANCE B, which are displaced in time. Fig. 14 shows two stages of a cryotron stepping register. Parallel output gates are not shown.

Coincident-Current Circuits

Many interesting circuits can be made of cryotrons with two or more control windings wound over each other in such a way that the net magnetic field affecting

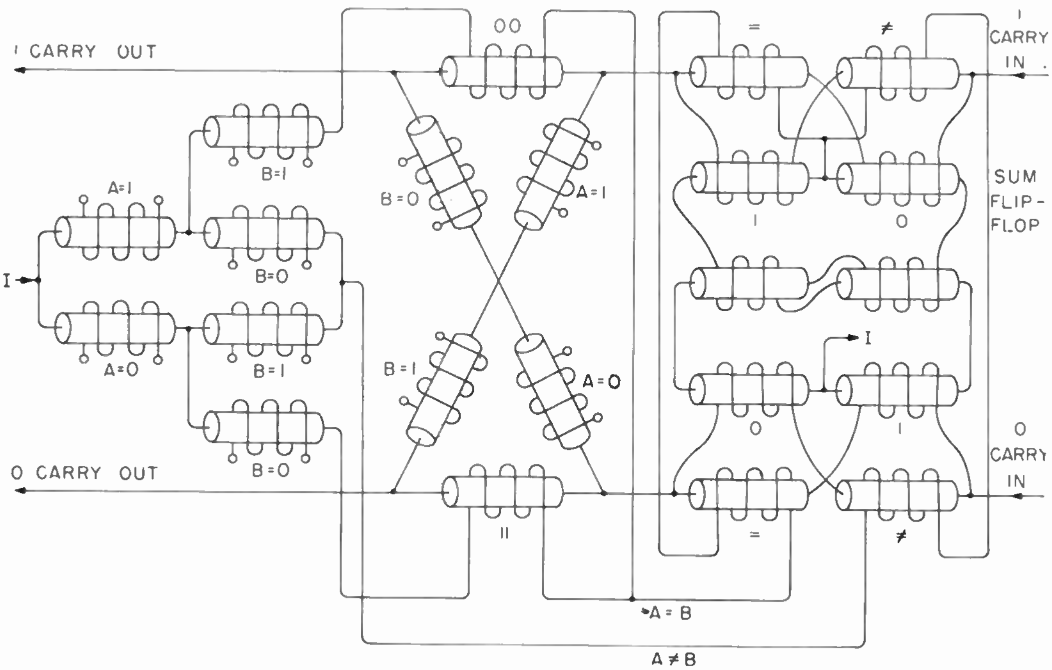


Fig. 13—Binary accumulator stage.

the central wire is due to the sum of the magnetic fields of the individual windings. The dc cryotron characteristics of Fig. 4 are sufficiently sharp in their transition between superconducting and normal states to allow the transition to result from the sum of two half-amplitude fields or even three one-third-amplitude fields. A coincident-current circuit of this type is useful for the selection of cryotron flip-flops placed at the intersection of the rows and columns of a matrix. A one-half-amplitude pulse is applied to the flip-flop control windings along a row, and a similar pulse to the flip-flop control windings along a column. The flip-flop at the intersection of that row and column can thus be placed in one of its two states; all other flip-flops in the matrix are unaffected.

If two such control windings are operated in opposition in such a way that the magnetic field of one subtracts from that of the other, a gate circuit of the "exclusive OR" type is available, wherein a flip-flop is set if either pulse A or pulse B occurs, but not if they both occur. Operation in this manner takes advantage of the fact that control depends on the magnitude of the controlling field and not on its polarity.

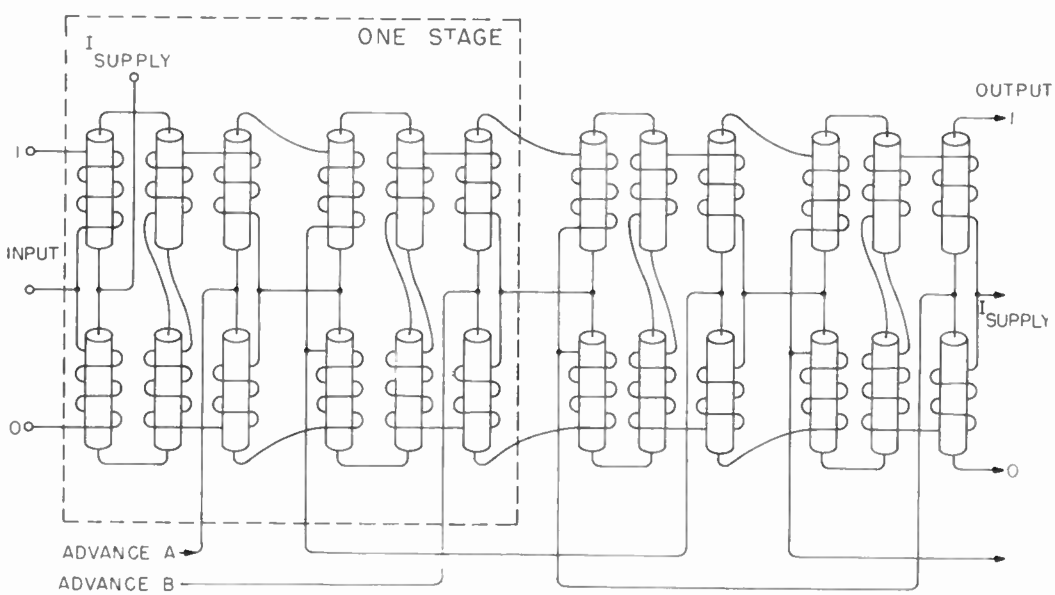


Fig. 14—Two stages of a cryotron stepping register.

Engineering a Cryotron System

Low-Temperature Environment

The most unusual requirement of a cryotron system is that it operates at a temperature near the absolute zero. Ten years ago this requirement would have precluded serious thought of such a system. Today, however, such an operating temperature is relatively easy to achieve.[7] This change is mainly due to the work of Samuel C. Collins whose helium liquefiers revolutionized the field of low-temperature physics. Arthur D. Little, Inc. of Cambridge, Mass., has built seventy Collins helium liquefiers of a 4-liter-per-hour capacity. The liquefier at M.I.T. liquefies 27 liters per hour. Storage of liquid helium has also improved. Commercially available double Dewars which use liquid nitrogen in the outer Dewar lose less than one per cent of their liquid helium per day.

The heat dissipated by a cryotron system causes evaporation of the helium. If the average power dissipated per cryotron is 10^{-4} watt, an estimate based on present experimental units, a 5,000-cryotron computer would dissipate one-half watt. The latent heat of vaporization of liquid helium at 4.2°K is 5 calories per gram,[8] its density is 0.1257, and therefore one-half watt corresponds to an evaporation rate of 0.93 liter per hour. A continuous system which recycles helium would be most economical for a stationary installation; a ten- or twenty-liter charge at the time of launching would suffice for portable systems.

The temperature of a liquid helium bath can be controlled by controlling the pressure of the bath. Table III gives the boiling point of helium at various pressures.

TABLE III
Boiling Point of Helium

Pressure mm Hg	Temperature °K	Pressure mm Hg	Temperature °K
0.001	0.657	720.0	4.156
0.01	0.791	730.0	4.170
0.1	0.982	740.0	4.184
1.0	1.269	750.0	4.198
10.0	1.743	760.0	4.211
100.0	2.638	770.0	4.225
200.0	3.067	780.0	4.239
300.0	3.368	790.0	4.252
400.0	3.605	800.0	4.266
500.0	3.803	900.0	4.40
600.0	3.975	1000.0	4.52
700.0	4.127	1500.0	5.03
710.0	4.141	1720.0	5.20

Below 2.19°K, the so-called *lambda-point*, liquid helium exhibits unusual properties which may prove useful in a cryotron system. A second phase of liquid helium appears which acts as a second fluid free to move through the first fluid with no friction. This zero-viscosity component is able to conduct heat with zero temperature gradient. It thus flows intimately in and around any structure immersed in it and allows rapid conduction of heat away from the structure. If heating is a problem in a cryotron system, operation in this temperature region should provide a solution. It is interesting, incidentally, to watch a liquid helium bath being pumped down. It may boil rather vigorously until the temperature drops below 2.19°K at which point the surface becomes perfectly still; heat is conducted through the liquid and liberated at the surface rather than on the container walls which causes boiling.

[7] C. A. Swenson and A. G. Emslie, "Low-temperature electronics," Proc. IRE, vol. 42, pp 408–413; February, 1954.

[8] W. H. Keeson, "Helium," The Elsevier Pess, Inc., New York, N Y.; 1942.

Physical Construction

Fig. 15 shows some experimental cryotron circuits. They are mounted at the ends of three-foot cupro-nickel tubes for immersion in a liquid helium storage vessel. Power supply and signal wires come up through the center of the tube. The experiments read chronologically from the large probes on the right which were used for dc characteristic measurements to the three-flip-flop multivibrator circuit on the left which contains nineteen active elements. A closeup of the latter experiment is shown in Fig. 16. The individual elements are those whose dc cryotron characteristics are given by Fig. 4. Spotwelding has been used to interconnect niobium and tantalum wires. Nickel lugs, while not superconducting, have proven useful for mounting. They both spotweld and solder nicely and careful design minimizes the resistance they introduce. The feasibility of using superconductive etched-wiring boards is under study. In these, lead would form the superconductive paths.

Many materials are used in the construction of circuits to operate in liquid helium. Ordinary wire insulation (enamel, silk, glass, Formex, Formvar, etc.) shows no sign of failure after repeated immersion. One experiment using wooden coil forms glued together with Duco cement was successful. Scotch electrical tape, while it freezes, seems to hold well. Commercially available feed-through and standoff insulators have been used without any sign of cracking. Metals in general are much stronger at extremely low temperatures. Some are relatively good thermal insulators (stainless steel and cupro-nickel) and may be used for mechanical support. There is no basis for the common impression that everything falls apart just below JAN specifications (−85°C).

Input, Output, and Power Supply

Input pulses to cryotron circuits involve current amplitudes which are easily achieved in the terminal equipments commonly associated with digital computers. Since the voltage level is low, input of information to a cryotron system involves no unusual problems.

Connecting the output pulses of a cryotron system to terminal equipment, on the other hand, is difficult due to the low power level of the cryotron circuitry. Power cryotrons can be designed to increase the power level,

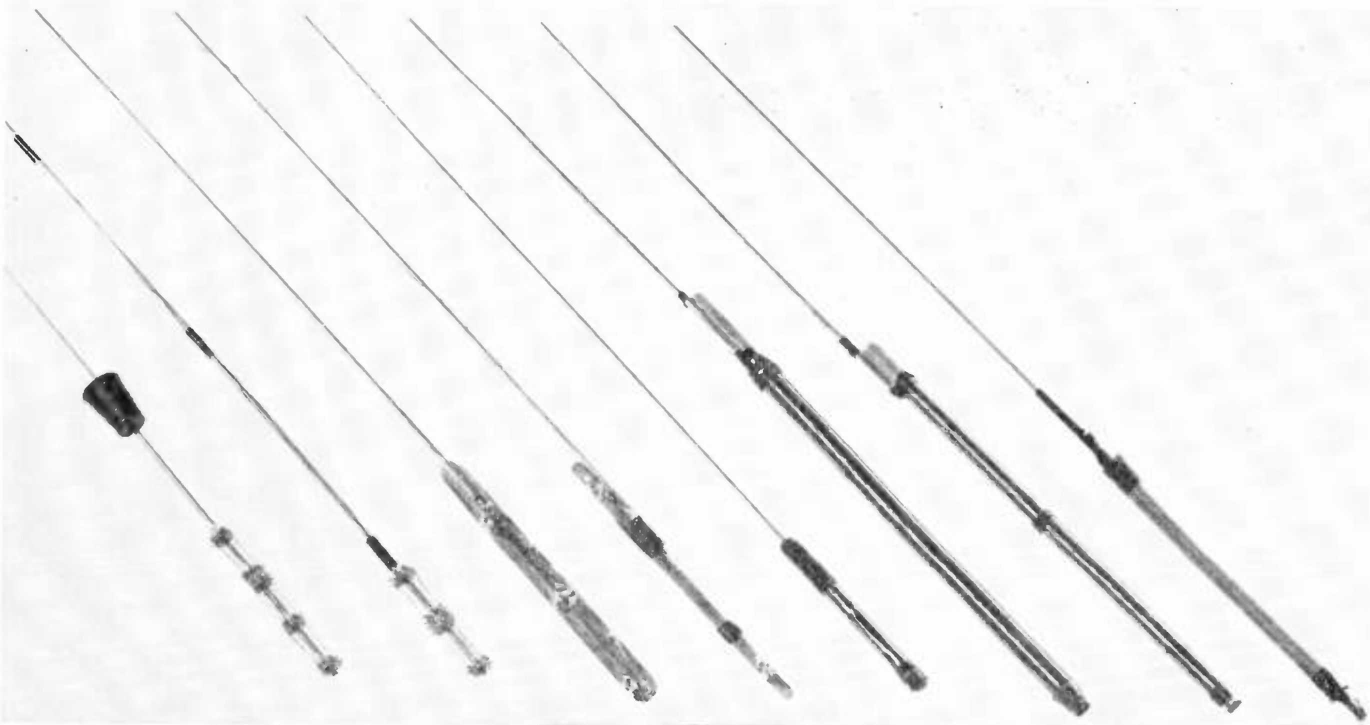


Fig. 15—Experimental cryotron circuits.

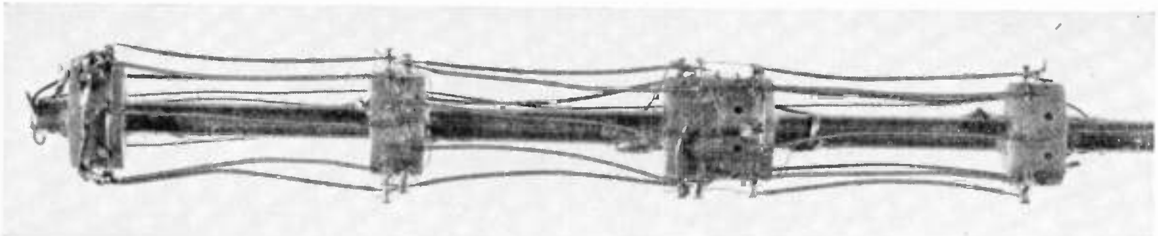


Fig 16.—3-cryotron-flip-flop multivibrator.

but it appears that vacuum-tube or transistor amplifiers are necessary to bring the level up to that of most output equipments. Magnetic amplifiers with superconductive control windings are an interesting possibility for power amplification.

Power supplies for cryotron systems are easy to achieve. The low impedance of the circuitry dictates a current-source power supply. A battery with a series resistance is adequate.

Conclusion

The cryotron in its present state of development is a new circuit component having power gain and current gain so that it can be used as an active element in logical circuits. It is easily and inexpensively fabricated from commercially available materials and its size is small. Extrapolating the volume occupied by the present experimental circuits to larger numbers of components indicates that a large-scale digital computer can be made to occupy one cubic foot, exclusive of refrigeration and terminal equipment. The power required by such a machine extrapolates to about one-half watt, once again excluding refrigeration and terminal equipment. The reliability of cryotron circuitry is not known, but it is anticipated that operation in an inert helium atmosphere at a temperature near to absolute zero where chemical activity and diffusion processes are essentially stopped promises a high degree of reliability. The circuit noise level is similarly not known, but due to the low temperature, very little thermal fluctuation noise is anticipated. The device is at present somewhat faster than electromechanical relays, but far slower than vacuum tubes and transistors. A program is under way to increase the speed.

Factors Affecting Reliability of Alloy Junction Transistors*

A. J. WAHL†, MEMBER IRE, AND J. J. KLEIMACK†

Summary—**Oxygen and water vapor, when individually in contact with the transistor surface, have been found to cause substantial and generally opposite changes in the characteristics of germanium alloy junction transistors. These changes, however, are reversible: by means of vacuum baking a reproducible set of characteristics can be repeatedly reestablished after water vapor or oxygen has caused a large change in the characteristics. Very pure forms of other ordinary gases, such as hydrogen, nitrogen, and helium, are found to have no effect on these transistors.**

Very nearly ideal time stability of characteristics can be obtained, even under severe aging conditions, when water vapor and oxygen are completely removed and permanently excluded from the transistor surface, although under some circumstances a very pure atmosphere of oxygen (or air) may be desirable for the *p-n-p* transistor.

INTRODUCTION

As A SOLID-state competitor to the vacuum tube, the transistor offers the promise of an extremely high order of stability and reliability, inasmuch as the primary electronic processes occur inside a continuous solid, with no hot cathode or vacuum or delicate mechanical structure to go awry. The expected stability of the interior of the transistor has indeed thus far been borne out by experience, but the extent to which surface effects contribute to over-all behavior was not foreseen at first.

In early experience with transistors excessive water vapor was found to cause transistor deterioration by increasing the collector junction reverse current.[1] A solution to this difficulty appeared to lie in dry room temperature hermetic sealing, a practice which was widely adopted and is still being credited with insuring against transistor deterioration. However, when such hermetically-sealed transistors still exhibited serious deterioration of characteristics during severe aging tests, a more careful and detailed study was undertaken to determine the cause of the deterioration.

In particular, the study was made on freshly-etched germanium alloy or "fused" junction transistors of both the *n-p-n* and the *p-n-p* types, similar in basic structure to those now widely marketed by various manufacturers (particularly of the *p-n-p* type). This basic structure is shown in Fig. 1.

The results of the investigation have disclosed that water vapor and oxygen individually have rather drastic effects on the transistor characteristics, such as junction breakdown voltage (V_B), junction reverse current (I_s), and alpha, whereas the effects of very pure forms of other ordinary gases, such as hydrogen, nitrogen, and helium are essentially nil.

NECESSITY FOR CAREFUL TECHNIQUES

Careful experimental techniques are necessary to separate the effects of oxygen and water vapor: 1) because of the prevalence of these two substances as normally encountered ambients, 2) because of the relative difficulty in completely removing them, 3) because of the large changes which each of them causes on transistor characteristics, and 4) because the effects of the two are generally counteracting. Therefore, to observe what are now believed to be the true individual effects of either water vapor or oxygen, the transistors used in this study were initially cleaned by baking in a high vacuum at as high a temperature as the particular transistor structure would permit.

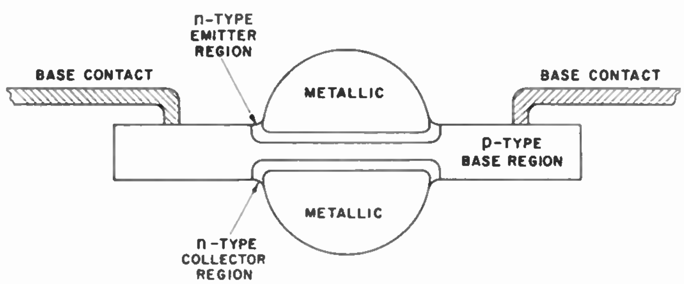


Fig. 1—Basic structure of alloy or "fused" junction transistor.

EXPERIMENTAL TECHNIQUES

Since the *p-n-p* transistor uses indium, whose melting temperature is about 150°C, the baking temperature was held to 135°C. On the other hand, since the *n-p-n* transistor uses arsenic-doped lead, it can be vacuum baked to at least 100°C higher temperature. A significant point of interest on the *n-p-n* is the fact that its characteristics are essentially the same whether the transistor has been vacuum baked at 135°C or 235°C, indicating that essentially all uncombined oxygen and water vapor is removed at the lower temperature insofar as effects on device parameters are concerned.

The vacuum in all cases was about 2×10^{-6} mm Hg. Spectroscopically pure reagent grade oxygen was used from glass bottles sealed to the vacuum station and ad-

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† Bell Telephone Labs., Inc., Murray Hill, N. J.

[1] R. M. Ryder and W. R. Sittner, "Transistor reliability studies," PROC. IRE, vol. 42, pp. 414–419; February, 1954.

mitted in measured quantities starting at a pressure of 2×10^{-2} mm Hg. In the case of water vapor, pure deionized distilled water was distilled once in the vacuum station under forepump vacuum into the cold trap and held at liquid nitrogen temperature during the vacuum baking process, then later released to give various vapor pressures by holding the cold trap at various temperatures starting at 0°C. (Before admitting oxygen or releasing water vapor, it was observed that the vacuum station could hold a high vacuum over long periods of time when shut off from the pumps.) When water vapor was used, pumping continued for a short time immediately after removal of the liquid nitrogen in order to remove any gases trapped with the water vapor. Such removal was evidenced by a rapid change in pressure from 2×10^{-6} mm Hg to about 5×10^{-5} mm Hg and back to about 4×10^{-6} mm Hg within about a minute after the liquid nitrogen was removed and before the beginning of the subsequent slow, steady rise in pressure due to release of water vapor. When the latter slow rise in pressure had begun, the system was shut off from the pumps and the temperature of the cold trap was brought up to 0°C by means of an ice water bath. The transistors did not change as a result of the first rapid pressure excursion into the 10^{-5} mm Hg pressure range. Since pumping was always started with a nitrogen atmosphere inside the vacuum station, most of this pressure was quite probably due to nitrogen, which has no effect on these transistors.

The results presented here were obtained with both *n-p-n* and *p-n-p* transistors on the vacuum station at the same time. The *n-p-n* transistors were designed with low α for a particular application. Since they were symmetrical units (emitters and collectors of same area), the α for each transistor was approximately the same in either direction, and the junction reverse currents were in the same range of values for emitters and collectors. On the other hand the *p-n-p* transistors had collector diameter twice the emitter diameter, hence the reverse currents of the collectors were correspondingly higher and the alpha in the normal direction much higher than that in the inverted direction (when emitter and collector are interchanged).

The experimental procedure consisted of taking measurements of the transistors in the sequence of steps shown in Figs. 3 to 8. The "vacuum bake" readings were taken after the transistors had cooled to room temperature but were still in high vacuum. The other readings were taken also at room temperature with the transistors in the various pure ambient atmospheres as shown.

Definition of Measurements

The junction breakdown voltage (V_B) is defined as the voltage at which the reverse current is 20 microamperes higher than the low voltage saturation current. In most cases this measurement was quite well defined, particularly for the *n-p-n* transistors. The reverse junction impedances were high (>10 megohms) and then with increasing voltage broke quite sharply to very low impedances. In general, the effect of the ambient (oxygen or water vapor) was to cause a shift of the characteristic parallel to itself to a different saturation current and different breakdown voltage, leaving the junction impedance essentially unchanged, as shown in Fig. 2.

The reverse current (I_s) was measured at 18 volts reverse bias on the junction being measured and no bias on the other junction.

The α is the dc alpha measured at one milliampere emitter current and very nearly zero collector voltage.

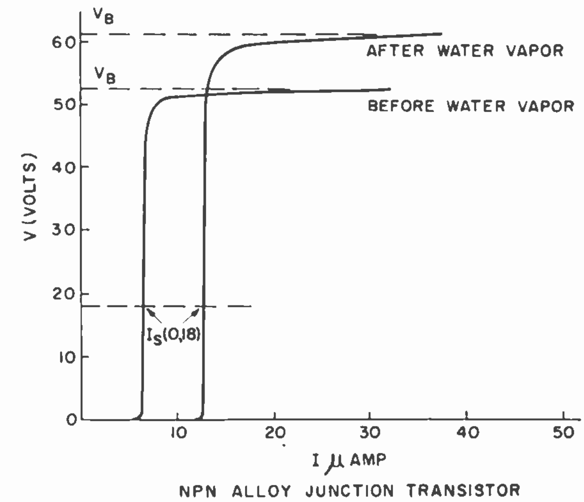

Fig. 2—Definition of I_s and V_B

Experimental Results

Figs. 3 to 8 (pp. 496–497) show how the junction breakdown voltage (V_B), junction reverse current (I_s), and α varied as a function of ambient, starting with a measurement in dry air before the first vacuum bake. Between successive exposures to oxygen and water vapor the transistors were vacuum baked and measured in high vacuum at room temperature.

Figs. 9 to 12 (p. 498) show how the median value of each parameter varied as a function of water vapor pressure or oxygen pressure, starting from a high vacuum after vacuum baking.

From a vacuum-baked reference condition the effects of oxygen and water vapor may be summarized as indicated below.

1) Oxygen
   a) On *n-p-n* alloy junction transistors
      1. Decreases the junction breakdown voltage (V_B)
      2. Decreases the reverse current (I_s)
      3. Decreases α.
   b) On *p-n-p* alloy junction transistors
      1. Increases the junction breakdown voltage
      2. Increases the reverse current
      3. Increases α.

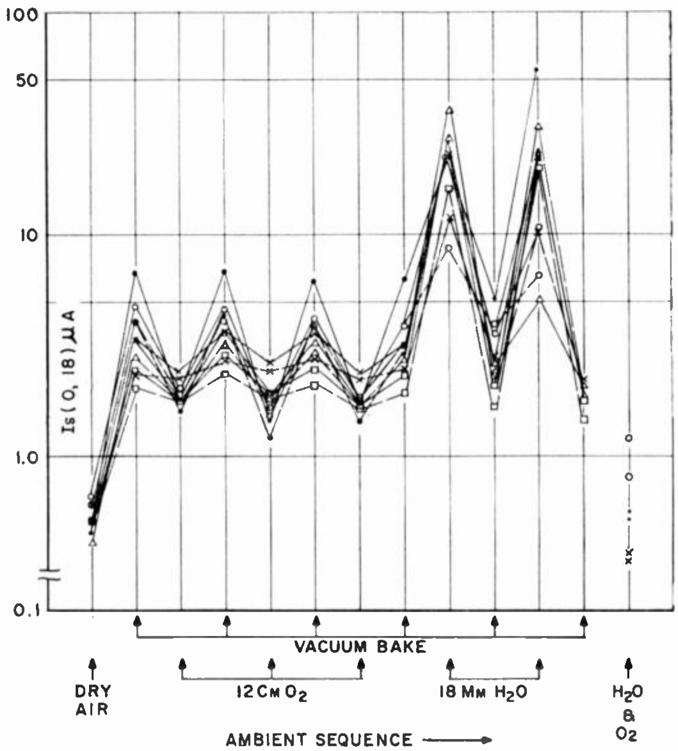

Fig. 3—Oxygen and water vapor effects on I_s of *n-p-n* alloy junction transistors.

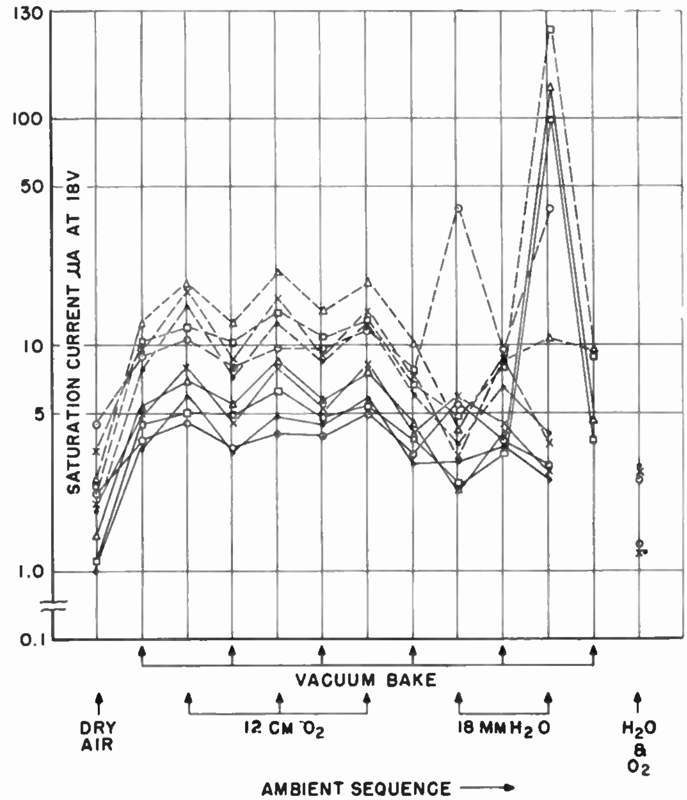

Fig. 4—Oxygen and water vapor effects on I_s of *p-n-p* alloy junction transistors.

2) Water vapor
   a) On *n-p-n* alloy junction transistors
      1. Increases the junction breakdown voltage
      2. Increases the reverse current
      3. Increases α.
   b) On *p-n-p* alloy junction transistors
      1. Decreases the junction breakdown voltage
      2. Decreases the reverse current until rather high vapor pressure is reached, then rapidly increases the reverse current.
      3. Increases α.

Some of the important features of the data shown in the figures are written below.

1) Vacuum baking produces a reference condition which can be re-established repeatedly after either oxygen or water vapor has produced a maximum change.
2) In general, the effects of water vapor and oxygen are counteracting, except that both substances cause an increase in alpha of *p-n-p* transistors.
3) On *n-p-n* transistors the highest reverse currents after vacuum baking are depressed most by oxygen.
4) The *n-p-n* junctions least sensitive to oxygen were actually the most sensitive to water vapor in regard to breakdown voltage.
5) In both types of transistors water vapor produces a greater change in α and I_s than does oxygen.

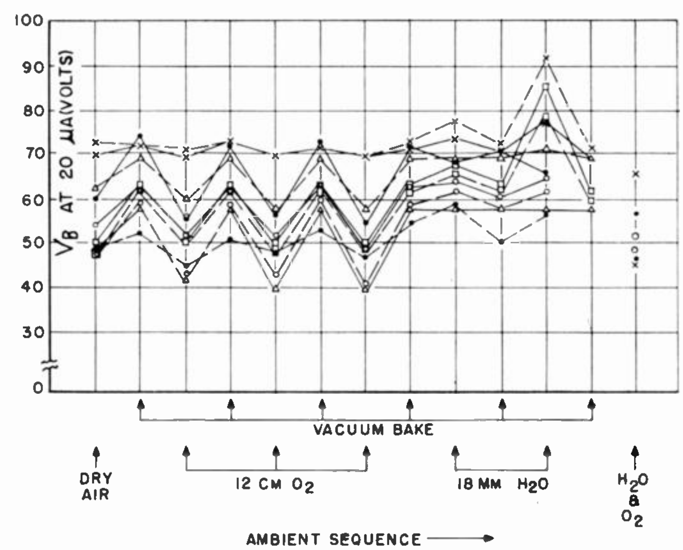

Fig. 5—Oxygen and water vapor effects on V_B of *n-p-n* alloy junction transistors.

6) Relative insensitivity of breakdown voltage to oxygen and water vapor does not necessarily mean that both α and I_s are also insensitive. Other results indicate that the I_s and α may be quite insensitive to water vapor and oxygen while the breakdown voltage can be strongly affected.
7) The effect of oxygen on all three parameters of both types of transistors definitely begins to be observable at a pressure of about 10^{-2} mm Hg,

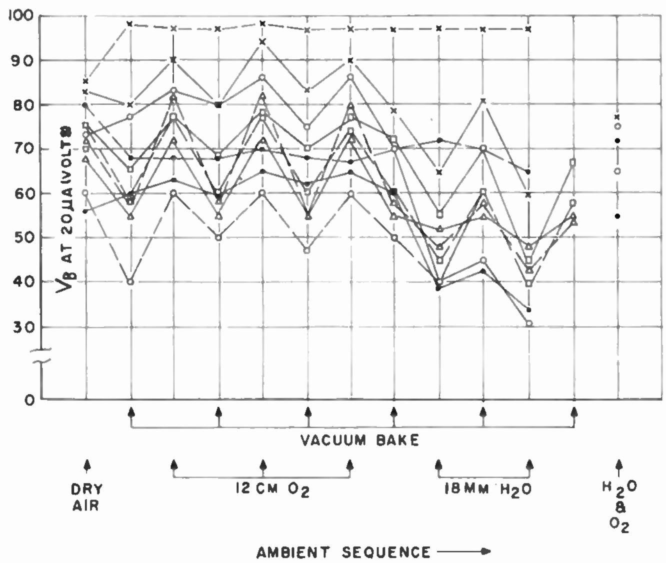

Fig. 6—Oxygen and water vapor effects on V_B of *p-n-p* alloy junction transistors.

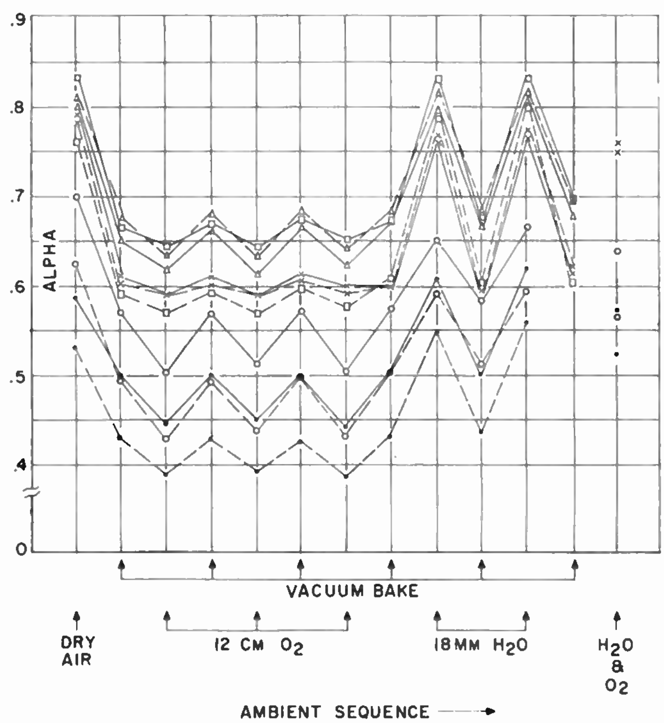

Fig. 7—Oxygen and water vapor effects on alpha of *n-p-n* alloy junction transistors.

which corresponds to the lowest measured quantity of pure oxygen admitted from a high vacuum condition.

8) After cycling with water and oxygen, the initial values of V_B, I_s, and α in dry air before the first vacuum bake can be very nearly regained by a combination of water and oxygen.

Additional observation on the effects of water vapor and oxygen is given as follows:

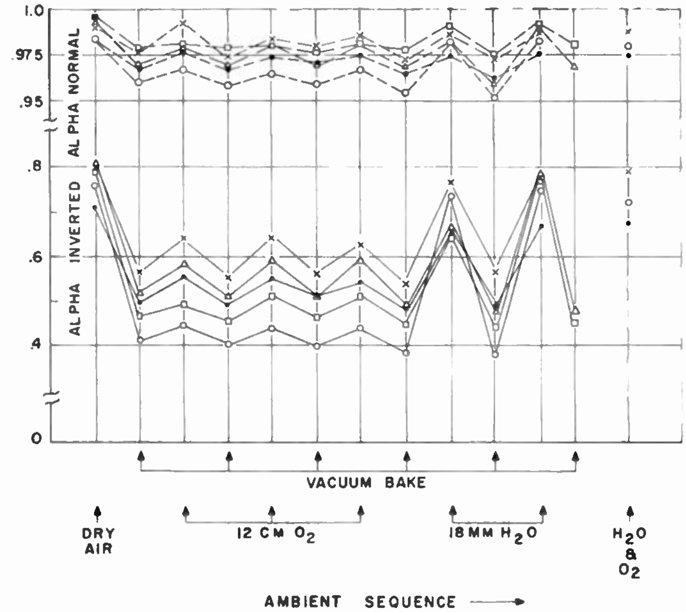

Fig. 8—Oxygen and water vapor effects on alpha of *p-n-p* alloy junction transistors.

1) In the presence of water vapor
   a) On *n-p-n* transistors
      1. Forward bias causes a rapid increase of I_s accompanied by an increase of α. Both I_s and α recede together upon removal of the forward bias.
      2. Reverse bias causes a much slower increase in I_s. The change caused by the reverse bias slowly disappears upon removal of the reverse bias.
   b) On *p-n-p* transistors
      1. Reverse bias causes a rapid increase in I_s. Initial values are regained at the same rate after removal of the bias.
      2. Forward bias causes no noticeable changes; however, when the characteristic is viewed by means of an ac sweep, as for example at 60 cycles for scope presentation, the forward sweep greatly reduces the apparent rate at which I_s increases during the reverse sweep.

The vapor pressure at which these effects begin to occur varies considerably among transistors. Many show very marked effects at 4.5 mm Hg vapor pressure of water when no oxygen is present.

2) In the presence of oxygen: Similar changes in I_s with bias are not observed. The breakdown voltage of *n-p-n* junctions fades to lower values with increase of applied voltage beyond the breakdown voltage; however, the opposite effect on *p-n-p* transistors is not so consistently observed.
3) Neither water vapor nor oxygen in the range of pressures used caused an observable change in junction capacitance, except that a possible slight increase in capacitance may have occurred on

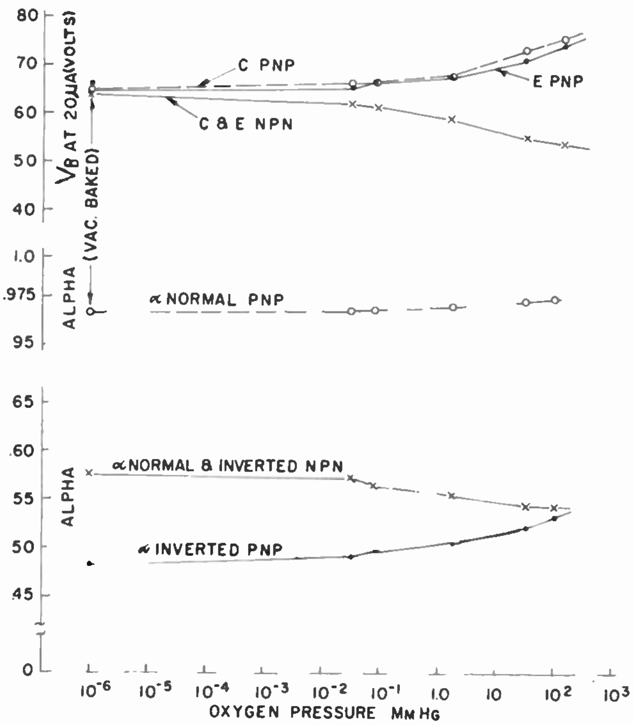

Fig. 9—Effect of oxygen pressure on V_B and alpha of *n-p-n* and *p-n-p* alloy junction transistors.

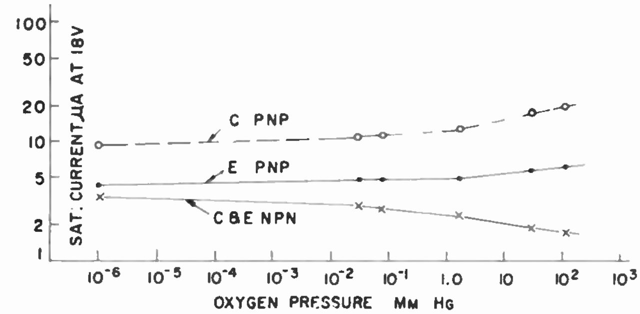

Fig. 10—Effect of oxygen pressure on I_s of *n-p-n* and *p-n-p* alloy junction transistors.

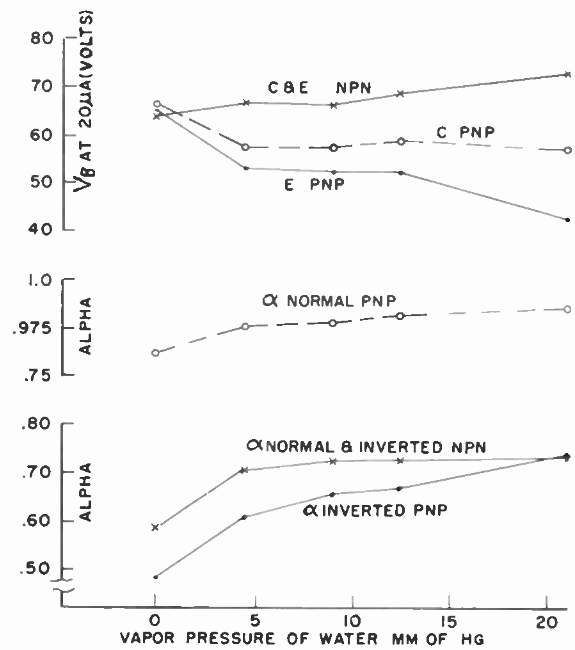

Fig. 11—Effect of water vapor pressure on V_B and alpha of *n-p-n* and *p-n-p* alloy junction transistors.

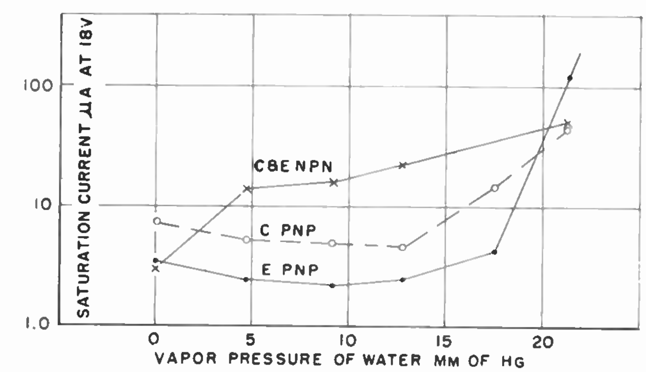

Fig. 12—Effect of water vapor pressure on I_s of *n-p-n* and *p-n-p* alloy junction transistors

some of the *p-n-p* transistors in oxygen. (Capacitance was measured at 20 kc with 4.5 v reverse bias on the junction.)

Discussion of Experimental Results

The two conditions not very well reproducible both involve water vapor effects: on the I_s of *p-n-p* transistors, and on the V_B of *n-p-n* transistors. The lack of reproducibility in both cases may be to a large extent due to the fact that the effects are very critical in the range of high water vapor pressure. For the *p-n-p*, this is the vapor pressure range where water ceases to decrease the I_s and starts to increase it rapidly. This switch does not occur at exactly the same vapor pressure for all *p-n-p* junctions. A variation in room temperature of a degree or two could affect the results significantly. In the case of V_B increase on *n-p-n*, the effect also occurs in most cases at a rather high vapor pressure, and the amount of increase varies considerably, many junctions showing rather little increase. Sometimes the failure to observe V_B rise was due to the fact that the forward half of the ac sweep voltage with which V_B is measured drove the saturation current to such high values that legitimate observation of V_B was impossible at the high vapor pressures where the V_B enhancement might occur. It was found that if the forward half of the sweep is blocked out by an external diode, good readings of V_B are possible in the high vapor pressure range.

Discussion of Mechanisms

A possible explanation for the direction of shift of V_B may be made in terms of the avalanche breakdown phenomenon reported by McKay.[2] In this process the breakdown in the interior is a function of the integrated electric field across the space charge region and of the geometry of the junction. Since the width of the space

[2] K. G. McKay and K. B. McAfee, "Electron multiplication in silicon and germanium," *Phys. Rev.*, vol. 91, p. 1079; September 1, 1953.

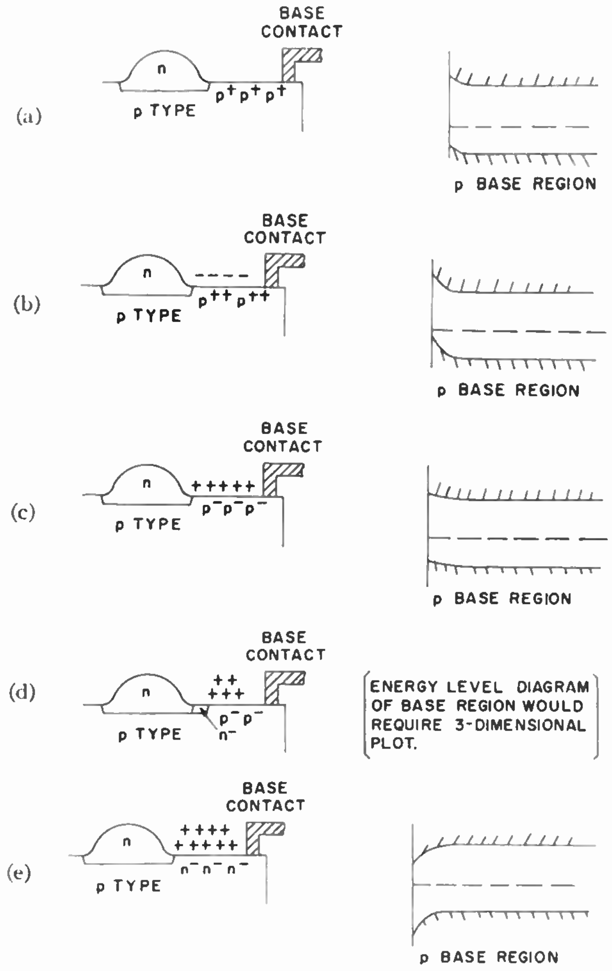

Fig. 13—(a) Normal condition, neutral ambient (surface more *p* type than interior because net V_B<body breakdown), *n-p-n* transistors. (b) Oxygen ambient, no bias, *n-p-n* transistor. (c) Water vapor ambient, no bias, *n-p-n* transistor. (d) Water vapor ambient, reverse bias, *n-p-n* transistor. (e) Water vapor ambient, high vapor pressure, no bias, inversion layer on base region, *n-p-n* transistor.

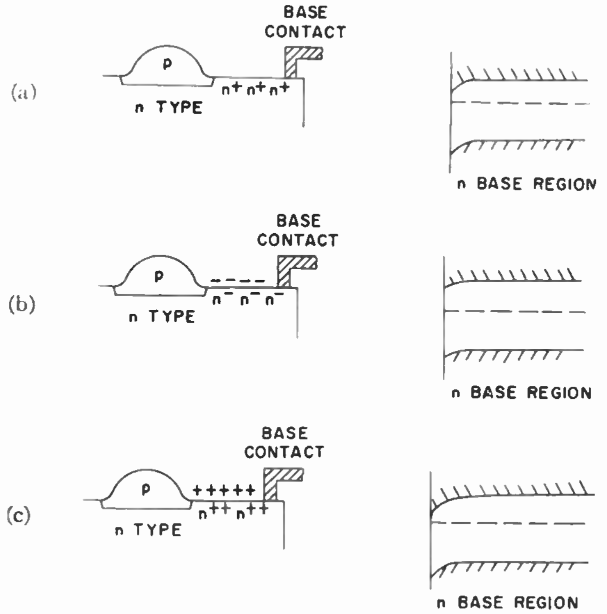

Fig. 14—(a) Normal condition, neutral ambient (surface more *n* type than interior because net V_B<body breakdown), *p-n-p* transistor. (b) Oxygen ambient, no bias, *p-n-p* transistor. (c) Water vapor ambient, no bias, *p-n-p* transistor.

charge region, for a given geometry, is a function of the resistivity, the breakdown voltage can be related directly to the resistivity. For the kind of junction in these transistors,[3] S. L. Miller[4] has shown experimentally that the body breakdown voltage varies directly with ρ^n, where ρ is the base resistivity and n is about $\frac{3}{4}$. The net V_B, then, is apparently determined by the lowest resistivity path from junction to base contact. Except in cases where base resistivity is purposely made very low, the V_B of these devices nearly always turns out to be much lower than the V_B expected from the resistivity of the bulk base material. One may say that the measured V_B of the junction is usually determined by breakdown occurring across the junction in the surface layers of semiconductor at lower voltage than for the body junction. If one makes a speculative extrapolation of the body behavior to the surface, one can conclude that the resistivity near the surface is lower than that of the interior, as depicted in Figs. 13(a) and 14(a). With this picture the effects of ambients can now be considered.

Let us first examine the effects of oxygen on these transistors. Upon contact with the surface the molecular oxygen apparently dissociates into atomic oxygen, which then satisfies its usual hunger for extra electrons and thereby produces negative ions. In the preservation of net charge neutrality the material immediately under the ionized oxygen acquires a positive space charge, which causes its resistivity to shift toward *p* type. On the *n-p-n* this process drives the surface resistivity to still lower *p* type, as shown in Fig. 13(b), thus decreasing the measured V_B. In a similar manner on the *p-n-p* the oxygen ions on the surface of the base region also shift the surface resistivity toward *p* type but not far enough to cause inversion. The net result is simply an increase in the *n* type surface resistivity as shown in Fig. 14(b), thus causing an increase in the measured V_B.

The behavior of water is more difficult to understand. Apparently water acts in such a way as to produce the net effect of positive ions on the surface.[5] In a manner analogous to the oxygen behavior, these positive ions on the surface of the base region induce a negative space charge near the surface, thus causing the surface resistivity to shift in the direction of *n* type. On the *n-p-n* transistor V_B is increased because the shift of surface resistivity toward *n* type has stopped short of actual inversion and merely caused an increase of *p* type resistivity, as shown in Fig. 13(c). On the *p-n-p* water vapor decreases V_B because the negative space charge induced

[3] Step junctions, with base resistivity at least two orders of magnitude greater than that of the emitter and collector.

[4] S. L. Miller, "Avalanche breakdown in germanium," *Phys. Rev.*, vol. 99, p. 1234; August 15, 1955.

[5] J. T. Law, "A mechanism for water induced excess reverse dark current on grown germanium *n-p* junctions," Proc. IRE, vol. 42, pp. 1367–1370; September, 1954.

by the positive ions on the surface causes the surface resistivity of the base region to shift to still lower *n* type, as shown in Fig. 14(c).

This picture of V_B behavior would also explain how water, which at first enhances V_B on *n-p-n* transistors, eventually causes V_B to fall suddenly to zero as the vapor pressure becomes very high. Such a sequence could happen as the surface layer, in shifting toward *n* type, first goes through high *p* type resistivity until actual inversion takes place, whereupon the junction becomes shorted by an *n*-type bridge to the metal base contact.

Now let us turn our attention to the behavior of I_s and α. If the effect of the ambient on these two parameters is due only to a change in surface recombination, then they should change in opposite directions. The only case in which such a combination of change occurs is that of water vapor on the *p-n-p*. To check this case quantitatively, a set of equations developed by Brattain and Garrett[6] were used. These equations give both I_s and α in terms of surface recombination for a structure similar to that of the transistors used in this study. With the observed data of Figs. 11 and 12 the results show to within about 10 per cent that the I_s and α change can actually be accounted for in terms of surface recombination change alone. In all other cases, however, α and I_s change in the same direction, indicating that one or more other mechanisms are also acting.

One method by which I_s and α could increase or decrease together would be by the formation or elimination of an inversion layer extending from the junction partway over the base region, thereby changing the effective junction area. Such an inversion layer could conceivably be established with the aid of the bias voltage, as shown in Fig. 13(d) if, for example, the water vapor pressure were high enough. However, if the junction area is changed, the junction capacitance should be correspondingly changed, but measurements showed no observable change in capacitance for the range of water vapor pressures used in this study. A slight increase in capacitance may have occurred with oxygen on some *p-n-p* units, but such an increase, even if real and not due to measurement error, would correspond to such a small change in junction area that it could probably not account for the observed changes in I_s and α.

Thus, if the absence of capacitance changes are accepted as a true indication of the absence of inversion layers and if the behavior of I_s and α cannot be accounted for in terms of changes in surface recombination, one must look for some other mechanisms.

This limited qualitative discussion is not intended to be complete or very accurate at this stage. Study of these phenomena in order to lead to an understanding of the mechanisms is a subject well worthy of separate treatment and will not be pursued further in this article.

[6] W. H. Brattain and C. G. B. Garrett, "Experiments on the interface between germanium and an electrolyte," *Bell Sys. Tech. Jour.*, vol. 34, p. 172; January, 1955 (equations A18).

It should be emphasized however, that no claim is made here that inversion layers cannot be established by water vapor or oxygen. In fact, it seems quite logical that for certain kinds of surfaces a high enough concentration of either of these ambients can indeed establish inversion layers, as depicted for water vapor in Fig. 13(e), and this possibility is borne out by observations of other workers for the case of water vapor. In this study, however, the effects observed on I_s, α, and V_B start at very low pressures of oxygen and water vapor, and the changes caused in these parameters are quite appreciable before actual inversion layers are established. Voltage bias on the junctions apparently adds still further complication to the action of these ambients.

Practical Significance

The practical significance of these water-oxygen effects lies in the behavior of these transistors under conditions of severe aging. Under nominally dry room ambient conditions, the counteracting effects of water and oxygen usually combine to yield very good transistor characteristics. If an attempt is made to capture and hold these characteristics by the standard type of room temperature hermetic sealing process, the stability of the transistor characteristics may be satisfactory, provided the transistor is afterward never heated much above room temperature. However, if the transistor is held for appreciable periods at an elevated temperature, such as 85°C with or without voltage bias applied, the changes in critical device parameters may be considerable when measured again at room temperature. These changes can usually be attributed to water vapor or oxygen driven off from the inside surfaces, thus upsetting the balance normally existing under nominally dry room ambient conditions.

The best technique known at present for the removal and exclusion of water vapor and oxygen involves vacuum baking and vacuum-tight sealing, leaving the transistor either in vacuum or in a very pure atmosphere of one of the various gases which do not affect it, such as nitrogen,[7] hydrogen, or helium. (The latter two are good heat conductors.) Evacuation without heating has been found to be inadequate because the internal surfaces of a sealed device may later evolve both water and oxygen if during evacuation the whole assembly is not heated to a temperature much higher than that to which it is subsequently subjected after seal-off. Under such circumstances the hermetic seal may be more harmful than helpful because it prevents the excessive contamination from escaping.

An argument can be made that vacuum baking makes the transistor surface extremely sensitive to either water or oxygen. Actually, the true sensitivity of the surface is probably unchanged, but removal of both water and oxygen causes an apparent increase in the

[7] Ordinary tank nitrogen, for example, usually has enough oxygen in it to make a noticeable effect on these devices.

sensitivity to either one of these substances because together, as normally observed under room ambient conditions, they counteract each other. An illusory reduction in sensitivity can be restored by admitting to the surface a little of both, but such a measure also restores the unreliability.

Although the best environment for *n-p-n* transistors is an absence of oxygen and water vapor, under certain conditions an atmosphere of oxygen may be desirable for *p-n-p*'s. For example, if the increase in I_s is tolerable, the oxygen enhancement of V_B and α may be attractive. However, the long term effects of oxygen in the absence of water vapor have not been evaluated.

It is known that certain surface oxides actually do reduce the sensitivity of the transistor to water vapor and oxygen. The use of such oxides, if established in a consistent and controlled manner, would be highly desirable, provided that no adverse results, such as high I_s or low V_B, would be produced on initial characteristics. Unless such oxides offer complete protection, however, the vacuum baking technique or its equivalent for final encapsulation will be necessary where good reliability is required; in fact it may always be necessary for good reliability as an extra safety factor, or particularly where the oxide does not cause complete insensitivity under all conditions and for long periods of time.

The experience of several thousand hours of severe aging tests has shown that the gradual deterioration which is characteristic of transistors hermetically sealed in the usual room temperature dry box environment can be essentially eliminated by the vacuum baking and sealing process in final encapsulation. Figs. 15(a), 15(b), and 15(c) show aging results for a typical batch of fifteen *n-p-n* transistors processed in this way and then held at 85°C with 28v reverse bias on each junction, but interrupted long enough for room temperature measurements.[8] By contrast, Figs. 16(a), 16(b), and 16(c), on the next page, show aging results for a group of ten *n-p-n* transistors hermetically sealed in a dry room temperature environment, then held at 65°C with 28v reverse bias and brought back to room temperature long enough for measurements. But these latter results, may not necessarily be the same as from another group of such transistors. In other cases reverse currents may not come down again with time and may in fact go even higher. Likewise breakdown voltage and alpha may change by different magnitudes. All of these features add to the unsatisfactory nature of the aging results from the transistors hermetically sealed in the ordinary "dry" room temperature environment and then subjected to aging at elevated temperatures. The remaining small fluctuations in the parameters of the vacuum-baked units are quite probably due largely to measurement error and other extraneous factors.

[8] Note again that these transistors, by design, have a low alpha for a special application. They can also be made with high alpha, but for these studies the low alpha units are particularly suitable because the low alpha is more sensitive to changes in the ambient.

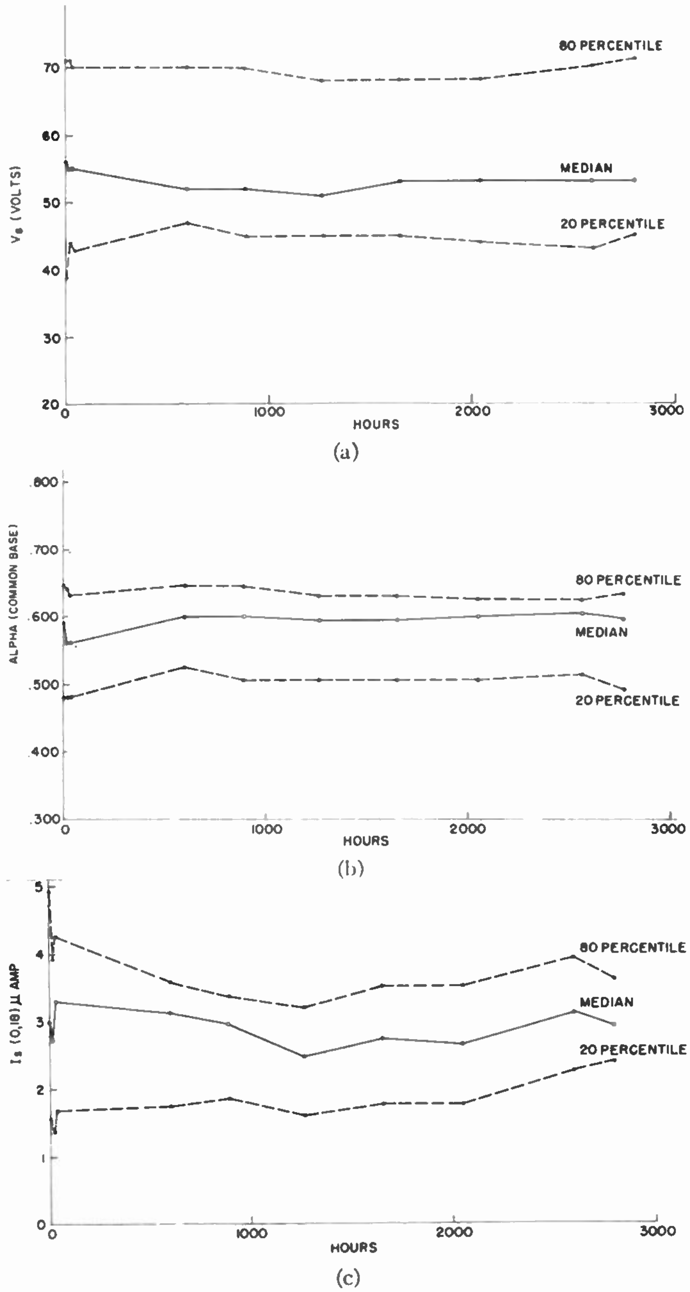

Fig. 15—(a) 15 *n-p-n* transistors, vacuum-baked and sealed, aged at 85°C with 28 v reverse bias, measured at room temperature. (b) 15 *n-p-n* transistors, vacuum-baked and sealed, aged at 85°C with 28 v reverse bias, measured at room temperature. (c) 15 *n-p-n* transistors, vacuum-baked and sealed, aged at 85°C with 28 v reverse bias, measured at room temperature.

Conclusion

The major significant changes caused by water vapor and oxygen on vacuum baked *n-p-n* and *p-n-p* germanium alloy junction transistors can be qualitatively summarized in Table I.

TABLE I

	O_2		H_2O	
	n-p-n	*p-n-p*	*n-p-n*	*p-n-p*
V_B	↓	↑	↑	↓
I_s	↓	↑	↑	↓↑
α	↓	↑	↑	↑

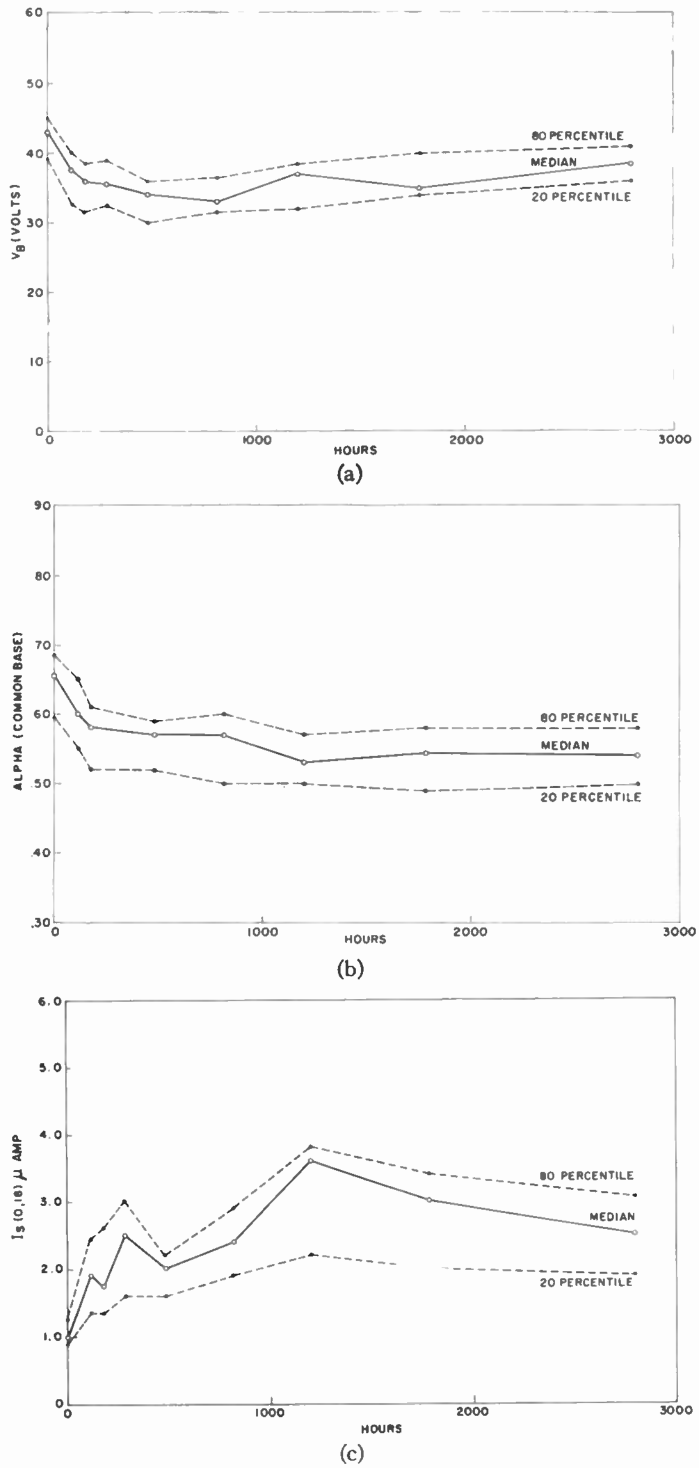

Fig. 16—(a) 10 *n-p-n* transistors, hermetically sealed at room temperature, aged at 65°C with 28 v reverse bias measured at room temperature. (b) 10 *n-p-n* transistors, hermetically sealed at room temperature, aged at 65°C with 28 v reverse bias measured at room temperature. (c) 10 *n-p-n* transistors, hermetically sealed at room temperature, aged at 65°C with 28 v reverse bias measured at room temperature.

The results are reversible and reproducible to a very good degree from the first cycle.

Although these observations were made on freshly-etched transistors, other experiments show that the effects are essentially the same for units subjected initially to oxygen atmosphere at room temperature with voltage sweep applied for as long as two weeks after final etching.

For most transistors as normally made at present and then vacuum baked, the effects of oxygen begin to be noticeable at pressures as low as 10^{-2} mm Hg, while water vapor at a pressure corresponding to water held at 0°C (4.6 mm Hg) causes a marked effect on the parameters under observation (V_B, I_s, α).

Excellent aging results have been obtained from those transistors from which oxygen and water vapor have been removed and excluded. Cases of departure from essentially flat aging on such units may be traced to imperfect removal and/or exclusion of oxygen and water. (Small leaks in the container can be a major source of trouble.)

The following conclusions may be drawn from these studies.

1) In ordinary applications where the transistor will not be operated much above room temperature or where extreme stability of characteristics is not required, the dry room temperature hermetic sealing process for final encapsulation may be adequate.
2) In severe applications requiring a high order of stability and reliability while operating for prolonged periods at elevated temperatures, the vacuum-baking process, or its equivalent, is necessary for final encapsulation.
   a) In the case of the *n-p-n*, water vapor and oxygen must be eliminated and excluded, leaving the transistor either in vacuum in a very pure atmosphere of one of the various gases which do not affect its characteristics. (Hydrogen or helium would be best from the standpoint of heat conductivity.)
   b) In the case of the *p-n-p*, similar processing and protection is necessary, particularly against water vapor, but if the increase in reverse current can be tolerated, an atmosphere of pure air or oxygen may be desirable because of the enhancement of α and breakdown voltage. However, the long-term effects of oxygen in the complete absence of water vapor have not been evaluated.

The alternative to the above procedures would be the development of a surface which would make these transistors completely and permanently insensitive to oxygen and water vapor under all conditions of operation.

In general, we conclude that except for the rare case of a sudden failure, the originally expected stability and reliability of these transistors can indeed be realized, but only with a much higher order of surface clean-up and protection than was first believed necessary.

Microwave Detector*

J. T. MENDEL†

Summary—**A new microwave detector is described which possesses extremely large bandwidth capabilities in addition to being free from power limitations. The sensitivity is comparable to that of a crystal. This new device operates on the electron beam of a traveling wave tube as a velocity sorter by utilizing the stop-band phenomenon of a periodic magnetic focusing system. The electrons are sorted according to their rf velocity modulation and thus the collector current is a function of the microwave power in the beam. Several important related devices are discussed including a high level mixer.**

Introduction

THE ADVENT of periodic focusing[1,2] of pencil electron beams has made possible a promising new method of microwave signal detection. By utilizing the "stop-band" phenomenon characteristic of periodic magnetic focusing an rf detector and mixer tube has been made which possesses bandwidth, power output, and sensitivity (rectified current vs rf power level) far in excess of those of conventional silicon rectifiers.[3] Not only does this device overcome many of the limitations of crystals but in addition it is mechanically quite simple and involves no unreasonable tolerances even at the high end of the microwave spectrum. Actually it need not be any more frequency sensitive than a broadband traveling wave tube.

This new mechanism is based upon the velocity discrimination properties of a periodic magnetic focusing system for pencil electron beams.[1,2] It has been shown both experimentally and theoretically that a spatially alternating axial magnetic field can overcome the forces of space charge in a beam. However, since this structure is periodic, and the perturbations of an electron in a magnetic field are also periodic, the two periods can be made to interact in such a manner as to produce violent defocusing. This is known as a "stop-band" since electron transmission is stopped.

This interaction is a function only of the electron velocity, the magnet spacing, and the magnetic field strength. Thus if these parameters are properly chosen the mechanism can be made to discriminate against certain velocity ranges by defocusing those electrons. The resolution (minimum detectable velocity difference) depends upon the length of the magnetic circuit and the period of the alternating lenses.

The sorting action of the periodic magnetic field operates on the spent electron beam after it leaves the rf circuit. A section of drift tube is added, following the helix (or other slow wave circuit) and beyond this the collector is placed as indicated in Fig. 1. In the drift

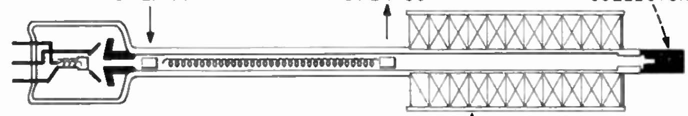

Fig. 1

tube the spent beam is focused with an alternately poled periodic magnetic field. For the case of no rf modulation on the beam, all of the electrons are focused by the periodic section and so the collector current is equal to the total beam current. When even a small amount of power is coupled to the input of the helix some of the electrons are slowed down below the dc beam velocity and are defocused to the drift tube by the periodic section. The net result is a decrease in collector current. For this action to take place the characteristic stop-band of the periodic section must lie just below the dc beam velocity. This is indicated for a typical structure in Fig. 2. It is apparent from

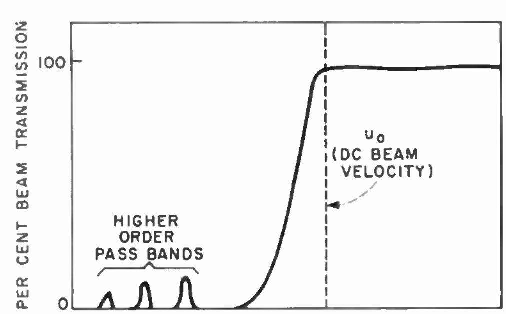

Fig. 2

this diagram that the sensitivity of the detector is primarily a function of the sharpness of the knee of the beam transmission curve. The sharper this knee, the greater the change in collector current, for a given power on the helix. As mentioned before this resolving power is determined by the length of the periodic magnet section, the spacing of the lenses, and the accuracy of each period. Imperfections in the electron stream such as transverse thermal velocities will naturally limit the maximum attainable resolving power. In addition to the sensitivity of the device it is important to know the proportionality between the change in col-

* Original manuscript received by the IRE, October 25, 1955; revised manuscript received December 29, 1955.
† Bell Telephone Labs. Inc., Murray Hill, N. J.
[1] A. M. Clogston and H. Heffner, "Focusing of an electron beam by periodic fields," *Jour. Appl. Phys.*, vol. 25, pp. 436–447; April, 1954.
[2] J. T. Mendel, C. F. Quate, and W. H. Yocom, "Electron beam focusing with periodic permanent magnets," Proc. IRE, vol. 42, pp. 800–810; May, 1954.
[3] This is an active device and possesses rf amplification, but even discounting this gain the sensitivity still compares favorably with that of a crystal.

lector current and rf power on the helix; *i.e.*, the detector law. To find this law the beam transmission curve of Fig. 2 can be expanded in a Taylor series about the operating velocity u_0

$$I(u_0 + \delta v) = I(u_0) + \frac{\partial I}{\partial v}\delta v + \frac{1}{2!}\frac{\partial^2 I}{\partial v^2}\delta v^2 + \cdots \quad (1)$$

where

u_0 = dc beam velocity

δv = small incremental change in velocity from the dc value.

The procedure is to find the velocity modulation, δv, in terms of the traveling wave tube output power and then the collector current can be expressed as a function of output rf power. From traveling wave tube theory,

$$v_{ac} = \frac{\eta E_{s0}}{\mu_0 \beta_e C \delta_1} \text{ (Pierce Notation)} \quad (2)$$

where v_{ac} is the peak ac velocity and E_{s0} is the peak electric field at the output of the traveling wave tube. In terms of power output from helix this becomes

$$\frac{v_{ac}}{u_0} = \left(\frac{2P_0 C}{I_0 V_0}\right)^{1/2} \text{ MKS Units} \quad (3)$$

Since v_{ac} corresponds to δv of (1), the change in rectified collector current is proportional to $(v_{ac})^2$ or to helix power output. Thus, the periodic detector behaves like a square law crystal, and the sensitivity is governed by the magnitude of $\partial^2 I_{\text{collector}}/\partial v^2$ [third term of (1)].

At this juncture it is well to consider the advantages of this type of velocity sorter over the conventional gridded type.

1) There is no danger of burning out delicate grids with the electron beam since the defocused electrons are collected on a drift tube which can be externally cooled to dissipate any amount of power.

2) There is no optical problem of dealing with secondary electrons since the drift tube can be operated well above the potential of the helix.

3) Space charge represents no serious problem since the beam is magnetically focused in the region of actual detection.

4) The tube structure is mechanically quite simple—just a metallic cylinder.

5) Broadband detection is inherent since only the shunting capacity of collector itself is involved and this may be no more than 1 $\mu\mu$f. Thus bandwidths in excess of 1,000 mc may be attainable.

6) The power level of the detected signal is no limitation as far as power dissipation is concerned.

7) Constant current output; *i.e.*, good for high impedance narrow band circuits too.

When compared to a conventional crystal used at the output of a traveling wave tube, the primary advantages are power handling capabilities and bandwidth, both of which are small for a crystal, although the crystal is much simpler in construction.

Experimental Results

The first experimental model (see Fig. 3) to test the principle of periodic detection utilized available parts which were not optimized for best performance but sufficed to give valuable information. It was expedient to operate the traveling wave tube at a frequency of 7,500 mc which was rather high for the size helix employed. With a large γa the gain was sensitive to small changes in the focusing fields, and thus measurements were made more difficult. A lower frequency, or lower γa would have been more suitable for the experiments performed.

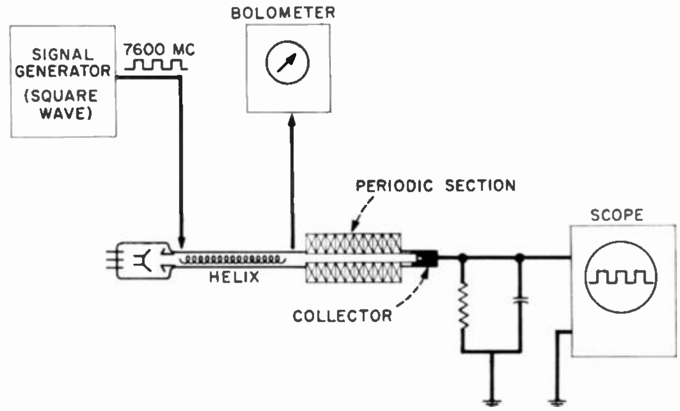

Fig. 3

The test procedure was as follows (Table I): The oscilloscope was calibrated so that the rectified current could be read directly and recorded as a function of rf power output from the traveling-wave tube.

TABLE I

Power Output from Traveling-Wave Tube	Rectified Current at the Collector (Magnitude of Square Wave)
1 mw	0.1 ma
10 mw	1.0 ma
20 mw	2.0 ma
30 mw	3.0 ma
40 mw	4.0 ma
50 mw	5.0 ma

These conditions seemed to yield the greatest sensitivity; *i.e.*, the largest detected current for a given power output. That this should be true is clear from Fig. 2 which indicates that the dc beam velocity should be such that almost all of the beam reaches the collector when there is no rf signal on the helix. Tests were made for a variety of operating conditions but in all cases the operation was essentially the same as that of Table II above, and in no instance was the sensitivity greater.

TABLE II

Conditions of Operation

Beam current	=10 ma
I collector (no rf)	= 7 ma
I drift tube	= 0.5 ma
I helix	= 2 ma

A rectified current of 100 μ amps per milliwatt of rf output power is about a factor 10 less than the best microwave crystals, operating at the same power level but there is no reason to believe that this cannot be improved upon by a more optimum design. Additional experiments with a longer drift tube and shorter magnetic period are necessary to make any comparison of sensitivities. Of course, because of the gain of the traveling wave tube, the periodic detector is much more sensitive than the crystal when compared on a basis of input power.

It is interesting to note that within the experimental accuracy, linearity of detected current vs power output held from 1 mw to 50 mw, which was the full range of measurement. This would indicate that the device has an inherently large dynamic range for which the detection law is a constant.

Approximate Analysis

The analysis of electron trajectories in a periodic magnetic field which is adjusted for operation near a stop band is quite complicated. Small ripple linear solutions are inappropriate since the equations become predominantly nonlinear. However, if one simplifies the problem by ignoring space charge a good qualitative picture of the beam behavior can be obtained. The justification for this procedure rests with the useful information that is thereby obtained.

Since the general dynamics of electrons in periodic fields is well covered in the literature,[1,2] only those features which are different and which apply to the detector will be presented here (see Fig. 4).

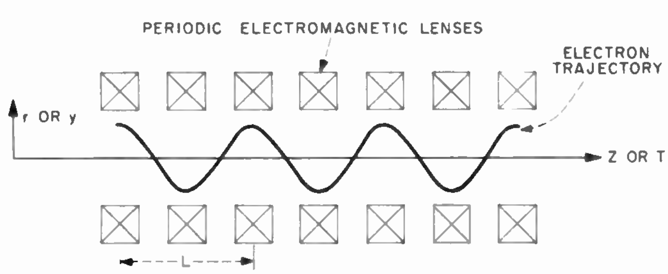

Fig. 4

In the absence of space charge the following equation holds for an electron in a periodic magnetic field.

$$\ddot{y} = \alpha(1 + \cos 2T)y = 0 \qquad (4)$$

where

$$y = \frac{r}{r_0} \qquad \omega = \frac{2\pi u_0}{L} \qquad \ddot{y} = \frac{d^2y}{dT^2}$$

$$T = \frac{2\pi Z}{L} \qquad \omega_L = \frac{1}{2}\eta B \qquad \alpha = \frac{1}{2}\left(\frac{\omega_L}{\omega}\right)^2$$

B = Peak magnetic field

L = Twice the distance between lenses

r_0 = Starting radius of the electron

Eq. (4) is a particular form of Mathieu's equation which possesses only one adjustable parameter, instead of the two which appear in the more general form. This restriction reduces the necessary range of investigation considerably and makes possible workable solutions.

Following the approach of McLachlan the solution of (4) in or near a region of instability can be written as follows.

$$y(T) = A_1 e^{\mu T}\phi(T) + A_2 e^{-\mu T}\phi(-T) \qquad (5)$$

where $\phi(T)$ is a periodic function of T with period π or 2π, and μ is real in an unstable region and imaginary in regions of stability. Evaluation of $\phi(T)$ can be accomplished in a straightforward manner but the work involved is considerable. Since it is a periodic function an explicit expression is not of much concern as far as the detection properties are concerned, and therefore only the exponential term will be considered in detail.

The term μ which governs the rate of radial growth of the trajectories is the critical factor of interest. It is only a function of α, and, therefore, in the case at hand it depends only on the velocity of the individual electrons (since magnetic field is held constant). Although an analytical expression could probably be derived for μ as a function of α it is much easier and more informative to solve directly for $y(T)$ with the aid of an analog computer. From these plots it is a simple matter to evaluate μ. Fig. 5 (page 506) (a through e) shows the electron trajectories and the envelope function for successively higher values of the parameter α (corresponding to successively smaller values of velocity if we assume B_0 held constant). The assumed initial conditions are,

$$\left.\begin{aligned} y'(0) &= 0 \\ y(0) &= 1 \end{aligned}\right\} \text{ at } T = 0$$

which means that the electron is traveling parallel to the axis at the peak of the first lens. The value of μ for each one of these cases can be calculated as follows.

$$y(T) = A_1 e^{\mu T}\phi(T) + A_2 e^{-\mu T}\phi(-T)$$

at $T=0$

$$y'(0) = 0 \qquad y(0) = 1$$

$$y'(0) = [A_1 - A_2][\phi'(0) + \mu\phi(0)].$$

Therefore,

$$A_1 = A_2 = \tfrac{1}{2}.$$

Thus,

$$y = \tfrac{1}{2}e^{\mu T}\phi(T) + \tfrac{1}{2}e^{-\mu T}\phi(-T) \qquad (6)$$

and the envelope will be $\frac{1}{2}e^{\mu T} + \frac{1}{2}e^{-\mu T}$. For $\mu T > 1$ the envelope is approximately given by $\frac{1}{2}e^{\mu T}$. Knowing T, the value of μ is then easily calculated from the figures. It should be noted that for one complete period of oscillation $T = 2\pi$.

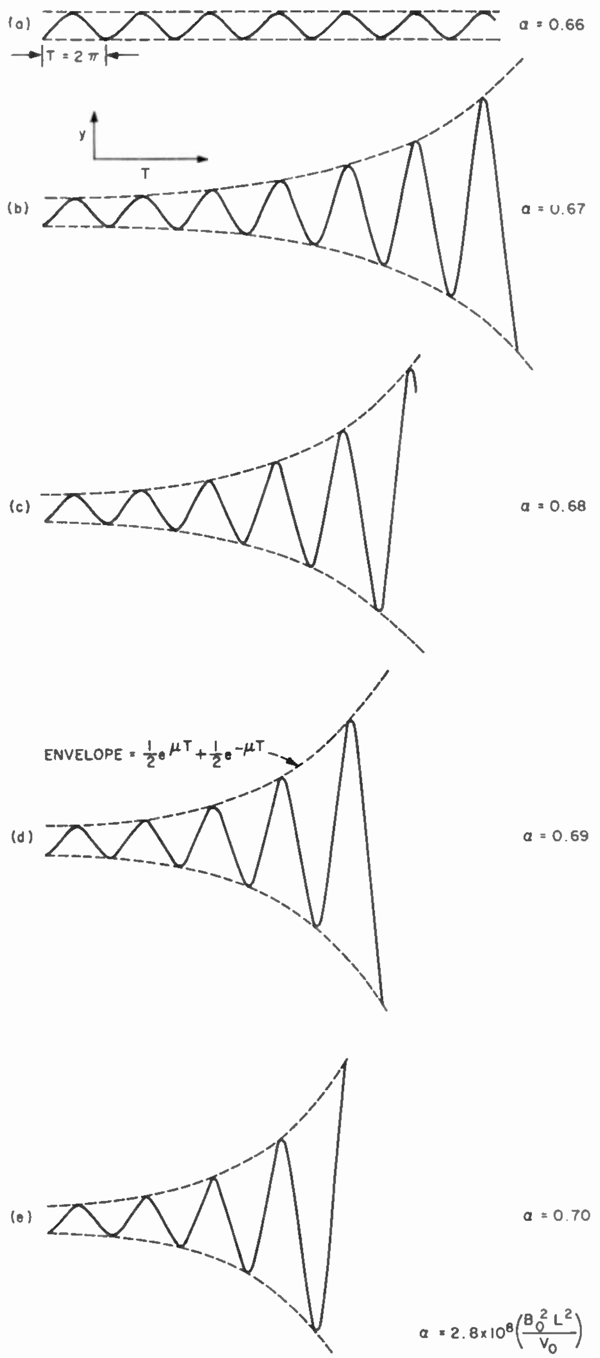

Fig. 5—Growth of the beam envelope (dotted lines) and individual electron trajectories (solid lines) as velocity is reduced with a fixed magnetic field. Initial slope of the electrons $[y'(0)]$ is assumed to be zero.

Fig. 6 shows the behavior of μ as a function of α as obtained from Fig. 5 (a through e). The greatest amount of discrimination between two different velocities occurs when the slope of this curve is a maximum or a value of α close to 0.66. Thus maximum detection efficiency should occur when the magnetic field is adjusted accordingly, and indeed this seems to be true from experimental measurements. Above a value of $\alpha = 0.68$ the slope of μ vs α decreases and eventually becomes zero and thus poor discrimination results from

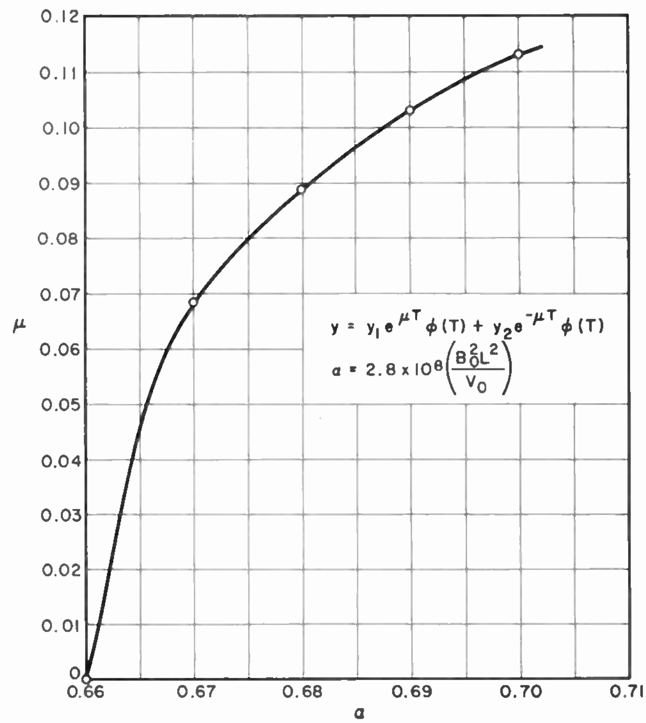

Fig. 6—The exponential growth factor μ plotted as a function of α.

operation in this region. A good qualitative picture of this behavior can be obtained from the Mathieu diagram as indicated in Fig. 7. The dashed line intersecting

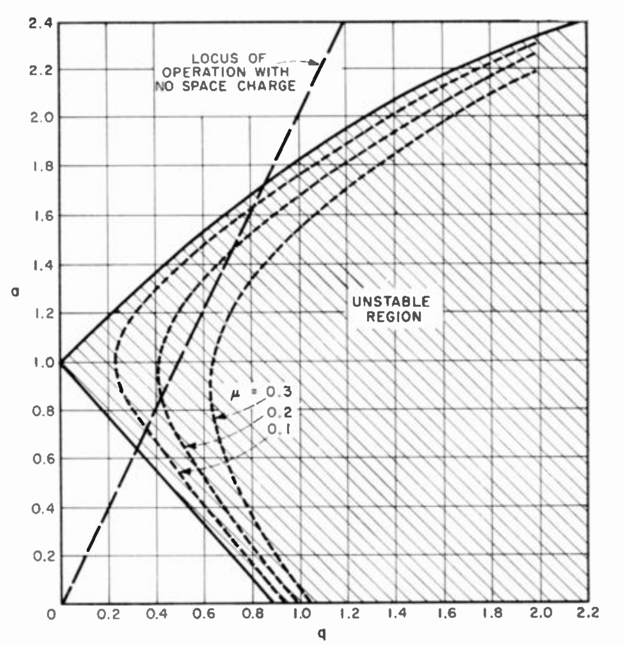

Fig. 7—Mathieu butterfly diagram illustrating the behavior of the detection mechanism in the unstable region. Here $a = \alpha$, $q = \alpha/2$ and μ is the exponential growth factor.

the origin defines the locus of operation for no space charge. Lines of constant μ are shown in the unstable area and it is the intersection of these with the dashed line that determines how μ varies as a function of α.

Sample Calculation

To evaluate the resolving power of this method of velocity sorting it seems best to consider a numerical example. Consider a drift space comprising 20 magnetic lenses or 10 periods which would be a total of 5 or 6 inches long. Since each period corresponds to an increase in the value of T of 2π the total length measured in the units of T would be 20π. Thus,

$$y_{\text{envelope}} = \tfrac{1}{2}e^{\mu 20\pi}. \qquad (7)$$
$$(\text{at } T = 20\pi)$$

As α is increased from 0.66 to 0.67 (which corresponds to a change in velocity of 0.76 per cent) μ changes from 0 to .07 and therefore,

$$y_{\text{envelope}} = \frac{1}{2}e^{1.4\pi} = 41 \qquad y = \frac{r}{r_0}.$$

Thus the radial position of an electron whose velocity is 0.76 per cent less than that of a stable electron is increased 41 times. This would indicate that a sharp sorting action could be achieved with a relatively short drift section.

To determine the power level in a TWT which corresponds to a .76 per cent electron velocity change (3) can be utilized.

$$\frac{v_{ac}}{u_0} = \left(\frac{2P_0C}{I_0V_0}\right)^{1/2}. \qquad (3)$$

For a typical low noise tube,

$C = .024$ $\quad I_0 = .5MA$ $\quad V_0 = 600$ volts

$\frac{v_{ac}}{u_0} = .0126P_0^{1/2}$ $\quad P_0$ in milliwatts.

Thus a power level of 0.36 milliwatt corresponds to a .76 per cent velocity change.

Effect of Initial Conditions

Since all of the electrons are not going to enter the periodic section with zero slope as previously assumed, it is worthwhile to consider what the effect is of an initial slope on the trajectories. There are two principal causes which lead to incorrect entrance conditions and which cannot be eliminated entirely. The first is due to transverse terminal noise velocities in the beam, but since this is small compared to other factors it will not be discussed. The second cause is attributable to nonlaminar flow conditions which exist in every beam to some degree. In principle it would be possible to adjust a laminar beam in a magnetic field such that all electrons were traveling parallel to the axis at some specified point. However, if the flow is nonlaminar; *i.e.*, electron paths crossing one another, it is not possible to make the trajectories behave in concert at any point. Some electrons will be traveling toward the axis while others are receding away from the axis. The magnitude of the slope will be a maximum when the trajectories are just crossing the axis and can be estimated from the magitude of the magnetic field in the uniform section.

$$\text{at } T = 0, \qquad \left.\frac{dy}{dT}\right|_{\text{max.}} = \frac{\eta BL}{4\pi u_0} \qquad (8)$$

where B = magnetic field strength in the uniform region. For the low noise tube used in the previous example (8) yields a value of

$$\left.\frac{dy}{dT}\right|_{\text{max}} = 0.1.$$

Fig. 8, (a, b, and c) illustrate the effect of this input derivative on the trajectories with α as a parameter.

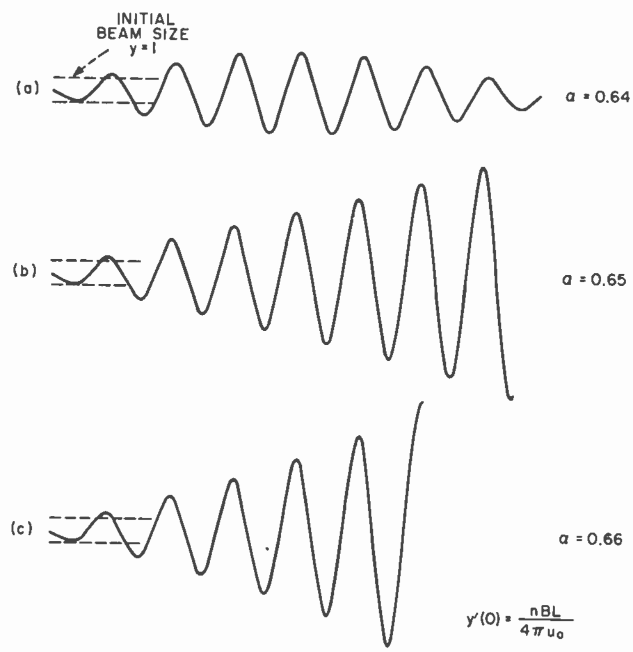

Fig. 8—Perturbations in the beam trajectories due to an initial entrance slope of $y'(0) = 0.1$. These gyrations are quasi-periodic and not growing waves.

Initially this effect masks the increase in the envelope radius due to small velocity changes which is the quantity of interest. However, the increase due to a velocity decrease grows exponentially whereas the other is a quasi-periodic disturbance which does not grow, and therefore the velocity effect eventually predominates. It should also be noted that all of the electrons will not be entering with this slope and thus the situation is more favorable than these curves would indicate.

Effect of Space Charge

Unlike many other types of velocity sorting detectors the periodic mechanism is capable of overcoming space charge forces while at the same time defocusing slow electrons. Although the problem of including space charge in the analysis is almost intractable (because of nonlaminar flow) the qualitative effect can be seen with reference to Fig. 7, the butterfly diagram. It is a fairly good assumption to let individual electrons thread through the beam independent of one another and influenced only by the average space charge forces. Careful studies of more realistic models indicate that this is not far from the actual case. Thus (4) becomes

$$\ddot{y} + \alpha(1 + \cos 2T)y - \beta y = 0$$

where

$$\beta = \frac{1}{2}\left(\frac{\omega_p}{\omega}\right)^2 \qquad \omega_p{}^2 = \frac{\eta\rho}{\epsilon} \qquad \omega = \frac{2\pi u_0}{L} \tag{9a}$$

rewriting,

$$\ddot{y} = \alpha\left[\left(1 - \frac{\beta}{\alpha}\right) + \cos 2T\right]y = 0. \tag{9b}$$

If $\beta \ll \alpha$; *i.e.*, the magnetic field is large compared to space charge forces, the dashed line (from the origin) of Fig. 7 is essentially unchanged and the operation is the same as if there were no space charge. When β is comparable to α in magnitude; *i.e.*, the magnetic field just strong enough to hold the beam at its initial size, the dashed line shifts downward and is parallel to the no space charge line of Fig. 7. As can be seen from the lines of constant μ such a move should not be detrimental to the velocity discrimination mechanism.

Noise Sensitivity

Wade[4] has shown that the noise figure of any velocity sorting scheme cannot surpass that of the traveling wave tube, assuming that the gain of the tube is sufficient to raise the noise above that of any subsequent element. Thus the periodic detector is limited by the particular traveling wave tube employed as far as noise figure and minimum detectable signal are concerned.

Other Applications

The ability of a periodic magnetic field to sort velocities has led to several suggestions of application to traveling wave tubes. Even at relatively low rf power levels there is a large spread of electron velocities in a traveling wave tube beam.[5] By selectively sorting the electrons within the beam it may be possible to alter the nonlinear characteristic so as to enhance the efficiency. Thus, the slow electrons which get out of phase with the circuit wave can be removed before they absorb energy from the circuit. This same technique could be utilized to produce signal limiting or expansion, or even slicing if more than one magnetic period were employed.

One possible application of more immediate interest is that of high level mixing. If two rf signals are coupled into the helix, the sum and difference frequencies will appear at the collector of the detector tube, which in turn could be the input to an IF amplifier. The advantage over a crystal would be one of power handling capability. Crystal mixing must be done at a very low level which restricts the system design to one which has only a small amount of rf gain before conversion. Bandwidth would also be a factor in favor of the magnetic mixer.

Conclusion

For certain applications the periodic magnetic detector possesses real advantages over existing devices both in simplicity and in performance. The rf bandwidth is limited only by the traveling wave tube to which it is attached and this could be of the order of several thousand mc. The video bandwidth is controlled by the collector capacitance and the output impedance. For broadband operation with a 50 ohm video impedance a 1,000 mc bandwidth is not unreasonable. Furthermore this bandwidth can be traded for a larger video output voltage merely by increasing the impedance without affecting the detection mechanism at all. In contrast a gridded detector or a crystal cannot be operated in conjunction with an arbitrarily high impedance without degrading the performance. The only elements of any complexity are the periodic lenses which are external to the tube and therefore do not add to the delicate problem of tube construction.

Acknowledgment

Many thanks to Noel E. Cram who performed most of the experimental work and to J. P. Laico who supervised the construction of one of the test models. Suggestions from C. C. Cutler were very helpful and his encouragement greatly appreciated.

[4] G. Wade, Tech. Rep. No. 75, Elect. Res. Lab., Stanford Univ., Calif.

[5] Unpublished work by C. C. Cutler.

Increasing the Reliability of Electronic Equipment by the Use of Redundant Circuits*

C. J. CREVELING†

Summary—**Electronic equipments which employ large numbers of tubes and components are often unreliable because a large number of parts may fail. A means of countering the undesirable effects of increasing equipment complexity is by introducing even more tubes and components as redundant circuitry. A requirement for obtaining the conditions, in which redundancy is helpful is for the causes of failure of parts to be independent. This calls for circuits to be designed so that as parts fail, their failure will not cause others to fail. The equations relating reliability to the number of circuit elements in the redundant and nonredundant cases are derived and applied to examples which show the degree of improvement which can be achieved. It is shown that equipment having hundreds of tubes could be made sufficiently reliable to satisfy most requirements. This greatly increased reliability can be maintained indefinitely by providing maintenance periods during which replacements are made for the parts which have failed while the operation has continued satisfactorily, the faulty condition having been obscured by the redundancy. This latter property of redundant circuit must be considered when establishing maintenance doctrines for equipment using these principles, since operational tests do not usually indicate the presence of faulty components or tubes. Rather, each tube and component must be temporarily disconnected from its redundant counterpart for the purposes of the test.**

Introduction

In recent years electronic equipment has shown a rapid trend toward increased size and complexity which has posed problems of decreased equipment reliability and increased difficulty of maintenance. It is obvious that excessive equipment failures and subsequent "down" time for trouble-shooting and repair seriously limit the usefulness of such equipment. It is equally obvious that in the military situation there will be electronic equipments of a complex nature used under conditions such as flight operations where it would not be possible to undertake repairs immediately. It therefore becomes highly desirable that a design method be evolved to keep the probability of equipment failure for the duration of such an operation as small as possible. It would then be possible for equipment probability of failure to remain low for considerable periods, by regularly scheduled maintenance.

Component Reliability

There are several attacks to the problem of improving equipment reliability. The most obvious is to improve the reliability of each component, hoping to arrive at the state where each component is so reliable that ensembles of such components of the desired size will still have adequate reliability. Unfortunately, the reliability of electronic equipment, which is primarily a function of the reliability of electron tubes or transistors (Appendix A), has become unsatisfactory as the number of tubes increases; the reliability of some large equipments is so poor that it needs to be improved by a factor of several times.[1]

To achieve greater equipment reliability by improving the reliability of the vacuum tubes calls for increasing their life expectancy by a large factor; in many cases development has already been carried to a point where such drastic improvement seems unlikely. For example, the so-called "reliable" tubes have an initial life expectancy in the region of 5,000 hours, and the longest quoted life expectancy for commercially available tubes is 10,000 hours. Although efforts to increase the life expectancy are continuing, it is not to be expected that any radical improvements can be made along this line. Even the expected life of the transistor is not more than ten times that of the vacuum tube. In addition, estimates of life expectancy for vacuum tubes usually specify that their operating conditions be carefully controlled. This is done as a matter of course in large permanent installations, but it is difficult to control voltages, temperature, pressure, and humidity under field conditions. Since variation of any of these may adversely affect the actual life of the tubes and components, it may be expected that in general the life quoted by the manufacturer may be well in excess of that realized under field conditions.

There is still, however, a widespread belief that the reliability of electronic equipment can be relieved from dependence on the life of their electron tubes by building the tubes to last as long as the rest of the equipment. It is supposed that this would result in an electronic equivalent of the "One Hoss Shay," running without care for a number of years and then spontaneously disintegrating, when everything wears out at once. Such wishful thinking ignores the random nature of vacuum tube failure. Because of the limited and unpredictable life of vacuum tubes the life expectancy of a large ensemble of tubes is only a small fraction of that of its individual tubes. This means that even with tubes having much longer life expectancies than those available at present, the probability of equipment failure due to tube failures, in large ensembles, would be disappointingly high. For example, in an equipment containing 1,000 tubes, if each tube has a mean life expectancy of 1,000 hours, there will be an average of one tube failing every hour. Then even if each of these tubes were replaced by a tube having a mean life of 10,000 hours,

* Original manuscript received by the IRE, July 11, 1955; revised manuscript received December 8, 1955.
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[1] R. R. Carhart, "A Study of the Current Status of the Electronic Reliability Program," Rand Corp. Research Memorandum RM-1131; August, 1953.

the equipment would still have tubes failing on an average of every 10 hours. Thus an intercontinental bomber on a 20-hour mission would hazard a considerable likelihood of being hampered by tube failure when relying on such an equipment.

Application of Redundancy

A more potent attack on the problem of increasing equipment reliability lies in the introduction of redundant circuits. Redundancy is the concept of providing duplicate circuits so that, should one circuit fail during operation, its counterpart would continue operating and the performance of the affected circuit would be substantially unimpaired. For greatest effectiveness and simplicity a counterpart should be provided for each circuit, arranging them so that each circuit in a pair operates simultaneously as a "twin," no sensing element being required to switch from one to the other when failure occurs.[2] This results in a great increase of reliability since, if the probability of failure of a unit is p_f, the probability of failure of both units making up the twin unit is p_f^2.[3] For example, if p_f were 0.001, p_f^2 would be 0.000001.

The probability of failure p_f or probability of survival p_s—terms which are mutually exclusive and whose sum is unity—can be determined from data on the life of the components in question. Individual component probabilities may then be combined to give the probability of failure or survival of the equipment composed of these parts. The probability of failure P_f of an equipment made up of n identical stages each having a probability of failure p_f, where any individual failures will cause failure of the equipment, is

$$P_f = 1 - (1 - p_f)^n. \tag{1}$$

If each individual stage is made redundant, the probability of failure becomes

$$P_f = 1 - (1 - p_f^2)^n, \tag{2}$$

whereas if the whole equipment were duplicated, the probability of failure would be

$$P_f = [1 - (1 - p_f)^n]^2. \tag{3}$$

It develops that redundancy becomes more effective as the organizational level at which it is introduced in the equipment is lowered. That is, an equipment which has several subdivisions will be made more reliable by duplicating each of these subdivisions than it would if the whole equipment were duplicated. There are simply more places where failures can occur without impairing the operation of the equipment. For example, a shift register having 20 identical stages, each of which has a p_f of 0.001 in a given period, would have a P_f of 0.02. If each stage of this register were made redundant, the resulting P_f would be 0.00002, or a decrease in probability of failure of 1,000 times. Had the equipment been duplicated in toto the resulting P_f would have become only 0.0004, a worthwhile improvement but much less than that attainable by the more thoroughgoing use of redundancy of the previous case. It should be noted, however, that when operating in a region where probability of failure is near unity, p_f^2 is not much less than p_f and the redundant circuits should be maintained in a stand-by condition and used to replace the operating unit if it fails.[4]

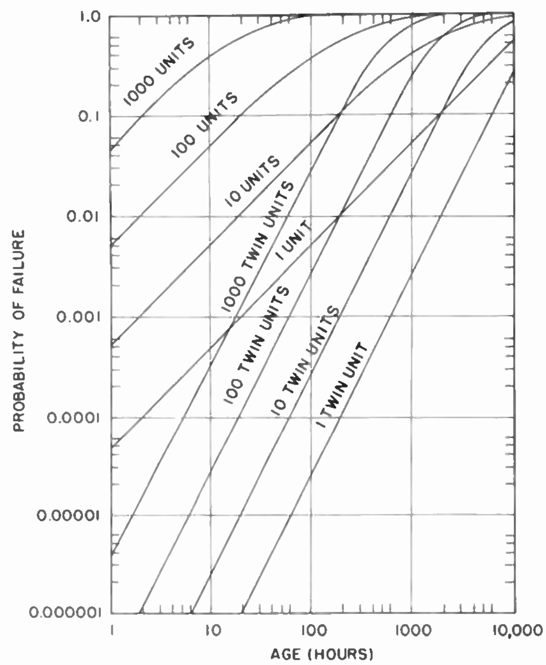


Fig. 1—Probability of failure of various equipment made up from series of identical units having a trapezoidal probability distribution with a mean life of 10,000 hours and a retirement age of 13,333 hours at which p_f is 0.5.

From (1) and (2) the nomographs of Fig. 1 above and Fig. 2, opposite, have been prepared to show the beneficial effects of redundancy in equipments having a multiplicity of identical unit stages. These unit stages are assumed to have a mean life of 10,000 hours with a trapezoidal distribution of probability of survival with age, and are arbitrarily retired when p_f has reached 0.5 as explained in Appendix A.

In Fig. 1 there is one family of curves indicating probability of failure of various numbers of units, as a function of time, and another family of curves for like numbers of twin units. For periods which are short com-

[2] Application of redundant circuits on anything less than a 1 to 1 basis as proposed above, poses the problem of providing some kind of sensing or supervisory device which will detect a defective stage and substitute a spare for it (there being less spare stages than original ones). The reliability of this device must then be considered in computing the over-all reliability of the equipment, and in the possible circumstance in which the reliability of the supervisory device might be low, there may be a net decrease in over-all reliability instead of the considerable gains theoretically possible.

[3] This assumes probabilities of failures of individual units are independent.

[4] Conditions of this type exist for amplifiers used in submarine cables where the total life is to be maximized. For this purpose each tube is used until it fails and then the stand-by is used to replace it, the total life being the sum of the individual lives.

pared with the mean life of the unit, redundancy may cause a decrease of P_f of as much as several orders of magnitude. This indicates that the redundant ensembles are much more reliable than nonredundant ones and shows that high short-time reliabilities can be attained with existing tubes which have modest mean lives, even in equipments which employ large numbers of these tubes.

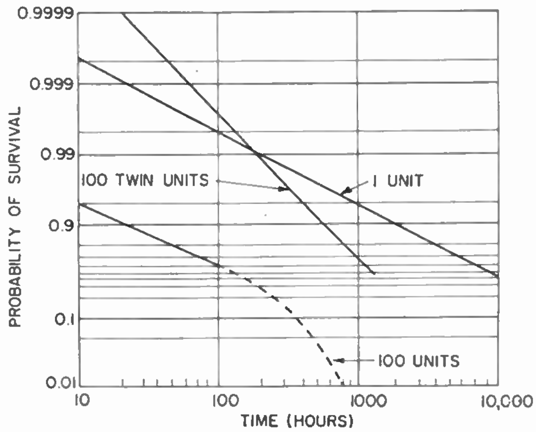


Fig. 2—Probability of equipment survival for a series of 100 conventional units and a series of 100 twin (redundant) units compared with that for a single unit.

Fig. 2 shows survival curves for 1 unit, 100 units, and 100 twin units of the same type as those of Fig. 1. It is seen that for a limited time, equipments having 100 such units may be made more reliable than their constituent units.

Special Considerations

To take advantage of redundancy requires special design considerations, since redundant circuits for power amplifiers, voltage amplifiers, quantizing amplifiers, etc., must be applied in different ways. The electronic circuit designer often has a choice of several types of circuit to perform a given function. Since some circuits are more amenable to the introduction of redundancy than others, this attribute should be taken into account when choosing the most suitable circuit for the job.

Equipment containing redundant circuits will generally be larger and heavier, require more power, and be more costly than equipment without redundancy. However, modern trends indicate that with the wider use of transistors and subminiature tubes and with miniaturizing techniques, it will be possible to build equipment equivalent to older types which will be smaller, lighter, and cheaper and which will have less power drain. In situations demanding increased reliability, some of these benefits can be plowed back into redundant circuits.

Redundancy also has the effect of concealing component failures. Therefore testing techniques must be devised which will reveal faulty circuits and allow them to be replaced.

Redundancy in Digital Circuits

Digital circuits operate on quantized signals. Furthermore the information flow throughout the circuits (as in a computer) must be requantized from time to time. This is usually accomplished by employing circuits which can only operate in certain discrete states as in the binary system. There are many devices used to accomplish this quantizing: multivibrators, blocking oscillators, regenerative amplifiers, etc. Of these, the circuit technique which appears to be the most readily adaptable to the introduction of redundancy on the circuit-element level is that which performs all logical operations by means of passive circuits, interposing active elements to compensate for the attenuation of these passive logic circuits. This technique of active and passive circuits with vacuum tubes in the active circuits has been used, for example, by the National Bureau of Standards in the SEAC.[5] The logic circuits being entirely passive, it may be presumed that their reliability will be good compared to that of the active elements. The active elements could then be mounted on subassemblies which could be inserted in the circuit. This would facilitate their removal for testing.

The beneficial effects of redundancy are then achieved by designing these active elements as "twin" units. The twin consists of a circuit in which two active elements (tubes, transistors, etc.) are made to do the work of one in such a way that if either fails, the operation of the other is unaffected. This qualification is a more particular way of restating the qualification that probabilities of failure of individual parts are independent; failure of each unit must be independent of the failure of the others, as well as independent of random causes affecting both units such as vibration, shock, voltage fluctuations, etc. This is important, since if care is not taken to see that this condition is obtained, the probability of failure may be increased instead of decreased. For example, connecting components in parallel which were more likely to fail by shorting than by open-circuiting would constitute an improper use of redundancy, since the probability of failure of the pair would then be $2p_f$ instead of p_f^2.

The design of a practical twin unit would require care be taken to insure that operating and bias potentials were properly fused and decoupled; input circuits were provided with series (stopping) resistors which would prevent grid shorts from shorting the signal; and similar precautionary measures were taken with other types of circuits.

Redundancy in Analog Circuits

The methods for introducing redundancy in analog circuits are necessarily different from the foregoing. For voltage amplifiers, one method which has been used with some success is to provide greater stage gain than

[5] S. Greenwald, R. C. Haueter, S. N. Alexander, "SEAC," Proc. IRE, vol. 41, pp. 1300–1313; October, 1953.

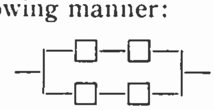
required and also apply negative feedback. The equations for the gain of an amplifier having negative feedback show that its gain can be made substantially independent of the gain of its constituent stages which are within the feedback loop. It is then possible to arrange a series of amplifier stages in such a way that if a stage fails, the resulting change of gain within the loop is not too great to prevent the over-all feedback from compensating it. This has been accomplished in the case of an intermediate-frequency amplifier by employing negative-feedback resistors from plate to grid of each stage. This feedback path is bilateral and provides a path for the signal around each stage should a stage become inoperative.

Redundancy can be introduced in power amplifiers by use of push-pull or parallel output stages. A familiar example is the high-quality push-pull audio amplifier having negative feedback, in which the removal of one of the output tubes during operation may have little discernible effect on quality. These amplifiers are redundant in that they are usually operated at less than rated output, so that a single tube of the output stage is capable of supplying the normal output, and the negative feedback normally used to reduce distortion compensates for change of gain when a tube fails.

It might be mentioned that the examples given do not provide protection against all types of failures. Defective tubes, wiring, or components may short out the signal or may change the operating potentials which are intended to supply the redundant part of the circuit. The use of decoupling circuits and overcurrent devices can minimize these effects and allow continued operation in the face of such failures.

Redundancy on the Component Level

The diode is one of the few components whose life may be considered to be limited in the same way as that of vacuum tubes or transistors. Theoretically, crystal diodes should have very long lives, being solid-state devices which show little or no deterioration due to use. They are, however, subject to deteriorating effects caused by their environment, and in actual practice it has proven necessary to replace them more frequently than, for example, conservatively rated resistors. It might be said then that their intrinsically long life has yet to be attained in practice. It turns out to be a comparatively easy matter to increase the reliability of diodes by introducing additional diodes as redundant members. Each diode can be replaced by a "quad," an assembly of four diodes connected in series-parallel in the following manner:



This quad will reduce the probability of failure by an "improvement factor" $1/G$ as shown in Appendix B where

$$G \cong \frac{p_s + p_0}{2p_s^2 + 4p_0^2}, \tag{4}$$

p_0 being the probability of a diode open-circuiting and p_s, the probability of short-circuiting. When $p_s = p_0 = 0.01$, G is $33\frac{1}{3}$ and when $p_s = p_0 = 0.001$, G is $333\frac{1}{3}$. Under normal operating conditions, p_s and p_0 would probably be considerably less than either of the above hypothetical values, and the improvement factor would be correspondingly larger.

This same expedient of using components in series-parallel can be used with resistors and condensers if the circuit is such that there is sufficient tolerance to changes in resistance or capacitance caused by failure of components within the quad.

Again it must be noted that the redundancy of the quad results in the masking of defects and requires testing techniques which will detect defective components. This might be done by grouping circuits into subassemblies in which all leads of all components are available at a terminal for test.

Conclusion

Although comments on the likelihood of maintaining electronic equipments having large aggregates of tubes in a reliable condition have tended to be somewhat pessimistic, it has been shown that by the introduction of additional tubes in redundant circuits, equipment having conventional vacuum tubes numbering in the hundreds if not thousands could be maintained so that the probability of failure would be acceptably low. The application of this redundancy to diverse types of equipment will require the adoption of the techniques enumerated here and will probably call for the development and refinement of these techniques.

Since redundant circuits may be more costly, heavier, larger, and more power-consuming than the ones they replace, the use of redundancy is, in general, most applicable where existing equipment is not sufficiently reliable and where penalty of failure is high.

Appendix A

Reliability of Equipment as a Function of Vacuum Tube Reliability

The reliability of equipment is dependent on the reliability of its components. The reliability of these components can be computed in terms of probabilities of failure from data concerning their life. To take into account the life of each individual component in computing equipment life would be a complex task especially since the expected life of components varies quite widely. Of these the vacuum tube is generally the least reliable and is the biggest factor in electronic equipment reliability. Since there are usually several passive components associated with each vacuum tube in a "stage" and these have lives which are long compared to that of the vacuum tube, it is not misleading to think of the life of a stage as being essentially that of the vacuum tube alone.

Studies of equipment failures have shown that a large proportion of vacuum tube failures is attributable to misapplication, poor design, or use under unfavorable environmental conditions. These factors are of no concern here since they are not strictly the fault of the components themselves. The effects of maintenance practices are of concern, however, since the reliability of vacuum tubes is a function of the distribution of their ages in the equipment, and this distribution is influenced by the way in which they are removed from service.

Comprehensive data on vacuum tube (or transistor) reliability are not generally available. Most vacuum tubes have been designed for use in equipments which use only a few tubes and in which low cost is an important factor. Generally, vacuum tube failure has imposed no serious penalty, so reliability has not been of primary importance. Experience of telephone companies has been published,[6] but since their equipment operates under carefully controlled conditions, such experience is not generally applicable to service or field use.

The reasons for vacuum tube failure are being studied but are not well understood. It has been stated by Gannett[6] that after some initial period, the survival curve follows an exponential law in decaying. Data published by Acheson and McElwee[7] however, indicate that a group of 555 premium subminiature tubes operating in an ambient temperature of 30°C did not obtain an exponential decay rate of 1 per cent per 100 hours until about 3,000 hours old. Prior to that time the rate of failure had been as low as 1/5 this value. This would indicate that if a rate of 1 per cent failures per 100 hours for this tube was tolerable, there would be no advantage in removing these from service in less than 5,000 hours of life (the limit for which data are given). A lesser rate of failures could be maintained by selecting tubes greater than 500 hours and less than 1,500 hours old. Tests run on like tubes at an ambient temperature of 175°C showed a constantly increasing rate of failure beyond 1,000 hours.[8] Under these conditions, tubes would have to be retired at some arbitrary age to keep rate of failure low.

On the basis of the meager data available it appears to be dangerous to generalize on the shape of the survival curve of the vacuum tube or to recommend replacement procedures without specific knowledge of the individual tube types and their respective circumstances of use. There is considerable agreement however that the initial period of tube life may have a higher than average rate of failure, but the removal of defective tubes in this period is properly the responsibility of the tube manufacturer and can be accomplished by a burn-in test where this is shown to be necessary.

A survival curve of a large number of tubes might resemble Fig. 3 in which the first vertical dotted line indicates the users receipt of the tube from the manufacturer, and the second vertical dotted line indicates a rejection point should the curve show any accelerating rate of failure.

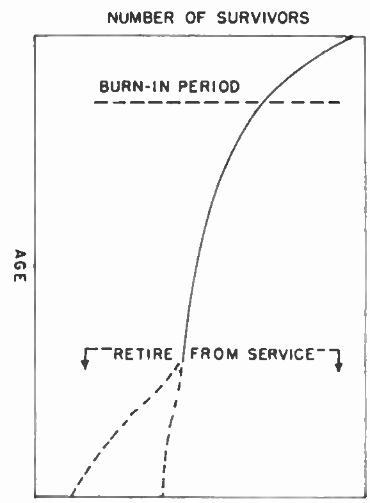


Fig. 3—A typical survival chart for vacuum tubes.

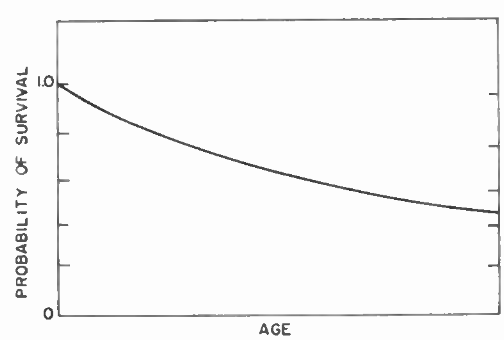


Fig. 4—Survival chart revised from Fig. 3 by initiating curve at end of burn-in period and terminating it at retirement age.

The survival curve of Fig. 3 can now be redrawn as in Fig. 4 with the point where the tubes were received from the manufacturer normalized to unity, the curve now representing the probability p_s of a tube being alive at any time after being put in service. The probability of a tube failing at any time is $p_f=(1-p_s)$. If the curves of p_f for a group of tubes used in an equipment are known, the probability of the equipment failing is

$$P_f = 1 - (1 - p_{f1})(1 - p_{f2})(1 - p_{f3}) \cdots (1 - p_{fn}) = 1 - \pi_i(1 - p_{fi}). \tag{5}$$

[6] D. K. Gannett, "Determination of the average life of vacuum tubes," Bell Lab. Record, vol. 18, pp. 378–382; August, 1940.

[7] M. A. Acheson and E. M. McElwee, "Concerning the reliability of electron tubes," Proc. IRE, vol. 40, pp. 1204–120; October, 1952, see Fig. 2.

[8] *Ibid.*, Fig. 3.

When all tubes are used under identical conditions and all have the same age, the p_{fi}'s become equal and (5) reduces to

$$P_f = 1 - (1 - p_f)^n. \qquad (6)$$

A more useful typical situation is one in which the equipment has a large number of tubes, all being used under substantially the same conditions. To optimize the reliability, tubes should be removed only when they are over-age (if, in fact, this proves to be necessary) and/or when they fail. The equipment will then gradually approach a state where it has a nearly uniform distribution of ages, all in the young category, and the rate of failure will be constant. The tube population is kept constant by replacing tubes which have failed (or reached retirement age). The failure rate of such a population will be less than that of a population consisting of all new tubes[9] and has been pointed out by Salzberg,[10] this makes the practice of wholesale tube replacements inadvisable.

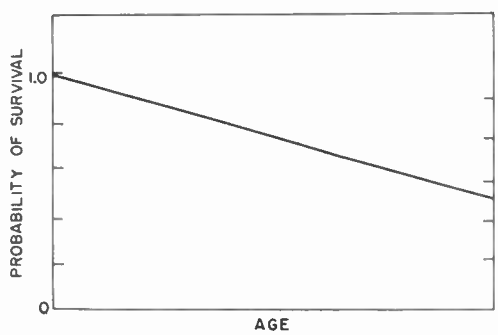


Fig. 5—Approximated probability of survival of a heterogeneous population (all ages in the young group). A curve of this type might be used to approximate a curve devised from actual experience.

For purposes of obtaining a simple relationship between p_s and the mean life, the survival curve of such a population may be approximated by the trapezoid of Fig. 5, from which p_s of an individual unit taken at random from such a population may also be approximated by a straight line. Then p_f will be considered to be linear function of time and may be used to construct charts such as those in Figs. 1 and 2 which depict how P_f and P_s vary with the number of tubes and as a function of time. Figs. 1 and 2 are based on an assumed mean life of 10,000 hours and a probability distribution of the form of Fig. 5 in which the tubes are retired at an age of 13,333 hours, it being assumed that p_s at this age would be 0.5. These hypothetical conditions are believed to be conservative; in any specific application, these assumed conditions should be based on data obtained from actual operation.

[9] "Investigation of Electron Tube Reliability in Military Applications," General Report No. 1, Aeronautical Radio, Inc.; September, 1953.

[10] B. Salzberg, "Tube Reliability; One Important Aspect" NRL Report No. 4125; March, 1953.

Appendix B

Reliability of a Component Quad

The probability of survival p_n of a single diode is

$$p_n = 1 - p_s - p_0 \qquad (7)$$

where p_s is the probability of short-circuiting and p_0 is the probability of open-circuiting. As shown in Table I below, there are 81 possible arrangements of shorted, open, or normal diodes in a quad, these arrangements being found by expanding

$$(p_n + p_s + p_0)^4 = 1$$

$$p_n^4 + 4p_n^3p_s + 4p_n^3p_0 + 6p_n^2p_s^2 + 12p_n^2p_sp_0 + 6p_n^2p_0^2 + 4p_np_s^3 + 12p_np_0p_s^2 + 12p_np_sp_0^2 + 4p_np_0^3 + p_s^4 + 4p_s^3p_0 + 6p_0^2p_s^2 + 4p_0^3p_s + p_0^4 = 1. \qquad (8)$$

TABLE I

All Possible Arrangements of Normal, Open, or Shorted Elements of a Quad*

NN NN	NN NS	NN SN	NN NO	NN ON	NN OO	NN OS	NN SO	NN SS
NS NN	NS NS	NS SN	NS NO	NS ON	NS OO	NS OS	NS SO	NS SS
			Operative					
SN NN	SN NS	SN SN	SN NO	SN ON	SN OO	SN OS	SN SO	SN SS
NO NN	NO NS	NO SN	NO NO	NO ON	NO OO	NO OS	ON SO	NO SS
ON NN	ON NS	ON SN	ON NO	ON ON	ON OO	ON OS	ON SO	ON SS
OO NN	OO NS	OO SN	OO NO	OO ON	OO OO	OO OS	OO SO	OO SS
					Inoperative			
OS NN	OS NS	OS SN	OS NO	OS ON	OS OO	OS OS	OS SO	OS SS
SO NN	SO NS	SO SN	SO NO	SO ON	SO OO	SO OS	SO SO	SO SS
SS NN	SS NS	SS SN	SS NO	SS ON	SS OO	SS OS	SS SO	SS SS

* In each arrangement the two elements in the upper row are in parallel with the two elements in the lower row.

Each of the terms in (8) represents the probability that a particular combination of normal, open, or shorted diodes will occur; the coefficient indicates the number of arrangements of the combinations which exist, the probability of each arrangement being represented by a product of unit probabilities to various powers. From Table I those causing circuit failure and those not causing circuit failure are determined by inspection, and by collecting the probabilities thereof,

$$P_s = p_n^4 + 4p_n^3p_s + 4p_n^3p_0 + 12p_n^2p_sp_0 + 4p_n^2p_s^2 + 2p_n^2p_0^2 + 8p_np_0p_s^2 + 4p_np_sp_0^2 \qquad (9)$$

where P_s is the probability of survival of the quad, and

$$P_f = 2p_n^2p_s^2 + 4p_n^2p_0^2 + 4p_np_s^3 + 4p_np_0p_s^2 + 8p_np_sp_0^2 + 4p_np_0^3 + p_s^4 + 4p_s^3p_0 + 6p_0^2p_s^2 + 4p_0^3p_s + p_0^4 \quad (10)$$

where P_f is the probability of failure of the quad. Substituting for p_n from (7), the latter equation reduces to

$$P_f = 2p_s^2 + 4p_0^2 - 4p_0^3 + p_0^4 - p_s^4,$$
$$= p_s(2 - p_s^2) + p_0^2(2 - p_0)^2. \quad (11)$$

An "improvement factor" G may be obtained from (7) and (11) as

$$G = \frac{1 - p_n}{P_f} = \frac{p_s + p_0}{P_f}$$
$$= \frac{p_s + p_0}{2p_s^2 + 4p_0^2 - 4p_0^3 + p_0^4 - p_s^4}, \quad (12)$$

or when p_s and p_0 are small, the higher order terms can be neglected and

$$G \cong \frac{p_s + p_0}{2p_s^2 + 4p_0^2}. \quad (4)$$

ACKNOWLEDGMENT

The writer wishes to thank Donald W. Lynch for invaluable assistance in formalizing the mathematical relationships which are the basis of this report.

Transformer "Miniaturization" Using Fluorochemical Liquids and Conduction Techniques*

L. F. KILHAM, JR.†, SENIOR MEMBER, IRE, AND R. R. URSCH†

Summary—Reduction of size and weight in electronic type transformers to keep pace with miniaturization in other phases of electronic component designing has led directly to higher temperature operation. Temperatures in the region of 185°C require inorganic materials throughout, including where used, the dielectric coolant. Design details and materials performance in "miniaturized" electronic type transformers are discussed in this paper. Specific Class *A* transformers are redesigned using these new techniques, thus giving size comparison.

INTRODUCTION

RELIABLE TRANSFORMERS for electronic applications, reduced in size and weight, have been designed using fluorochemical liquids and conduction techniques. These transformers, designed to operate in the region of 200°C, require inorganic materials throughout. Performance of high temperature materials in transformer operation with respect to corona, heat dissipation, toxicity, ability to self-heal, cost, and availability are most important. Dielectric properties of these materials at least equivalent to those of transformer oil are desired. Liquid and gaseous fluorochemicals fulfill these necessary requirements. Under a BuShips Contract entitled "Development of Transformers Utilizing Recently Developed Gases and Liquids Specifically Fluorochemicals," the Transformer Departments of the Raytheon Manufacturing Company have developed various transformer design techniques using these materials.

Although past experience indicated considerable size reduction by use of conduction techniques at higher temperatures, addition of fluorochemical liquids, vapors, or gases promises to improve and extend this advantage. It was demonstrated that liquid and gaseous fluorochemicals are compatible with materials commonly used in manufacture of high temperature transformers. Similarly liquid and gaseous fluorochemicals were tested and proved to be thermally stable at temperatures considerably in excess of 200°C high-temperature transformer operating conditions.[1] Properties of fluorochemical liquids and gases most applicable to magnetic component designs are to be discussed. Utilization of these properties in design techniques and their application to specific transformer problems will serve to show that transformers may be considerably reduced in size and weight by use of fluorochemical liquids, gases, and conduction techniques.

APPLICABLE PROPERTIES OF LIQUID AND GASEOUS FLUOROCHEMICALS

These inert, stable fluorochemical liquids have higher specific gravities and lower boiling points than materials commonly used in transformers.[2,3] They are nontoxic,

* Original manuscript received by the IRE, July 21, 1955; revised manuscript received November 17, 1955.
† Raytheon Mfg. Co., Waltham, Mass.

[1] M. Olyphant, Jr., and T. J. Brice, "Dielectric and coolant studies of inert fluorochemical liquids," Minnesota Mining and Mfg. Co., St. Paul, Minn., Proceedings 1954 Electronic Components Symposium, Washington, D. C.

[2] "Fluorocarbons," a brochure published by the Minnesota Mining and Mfg. Co.

[3] "Electrical Properties—Inert Liquids," a brochure published by the Minnesota Mining and Mfg. Co.

TABLE I

ELECTRICAL PROPERTIES OF FLUOROCHEMICAL MATERIALS

Identification Letter	Chemical Formula	Dielectric Constant 50 cycles to 100 kc	Power Factor %		Loss Factor		Resistivity ohms/cm³ 1.6kv/cm			Dielectric Strength A.S.T.M. D-877
			100 cycles	100 kc	50 cycles	100 kc	as received samples	after additional purification	after thermal aging	
Perfluoro Compounds										
A	$(C_4F_9)_3N$	1.89	0.0025	0.005	0.0025	0.0095	0.75×10^{14}	6.3×10^{14}	6.3×10^{14}	40 kv
B	$C_8F_{16}O$	1.84	0.016	0.029	0.118	0.053	2.5 "	6.3 "	0.63 "	40 kv
Conventional Oils										
C	transil	2.2	0.014	—	—	—	1.25×10^{14}	6.3×10^{14}	—	25 kv
D	silicone	2.7	0.01	0.01	—	—	6.3×10^{14}	—	—	30 kv

have power factors comparable to conventional transformer oils, and have excellent high temperature characteristics.

Dielectric properties of selected fluorochemicals are compared with transil and silicone oils in Table I above. Liquid fluorochemicals exhibit low dielectric constants, relatively high dielectric strength, and self-healing properties. These compounds are able to withstand repeated arcing without serious injury to their breakdown strength. Their dielectric constants and power factors were measured over a range of frequencies and are tabulated for 50 cycles and 100 kc as shown. Insulation resistance is tabulated for the measurements received from the manufacturers, after additional purification and after the thermal stability tests. In general, the dielectric properties of liquids and gaseous fluorochemicals compare favorably with the conventional transformer materials.[4,5]

Consideration of pressure developed in an enclosed system using liquid fluorochemicals must take into account the following points:

The unusually high coefficients of thermal expansion (several times that of mineral oil or silicone oil), the varying degree of volatility, and the absorption by the fluorochemical liquid of gas present in the expansion space may differ very materially from that experienced in a transformer filled with conventional transil oil.

Fig. 1 shows the pressures that may be developed using fluorochemical liquids with a gas space "void" in an enclosed container. It has been found that the pressure is affected by the solubility coefficient of the gas (see Fig. 2) which varies over the temperature range and is dependent on the gas-liquid system employed. The selection of the proper gas or gas mixture in the "void" space is an important consideration in controlling internal pressures commensurate with dielectric strength in transformers.

The heat transfer abilities of liquid fluorochemicals are excellent. These volatile liquid compounds with their low viscosity and high volume expansion offer excellent heat transfer possibilities by convective cool-

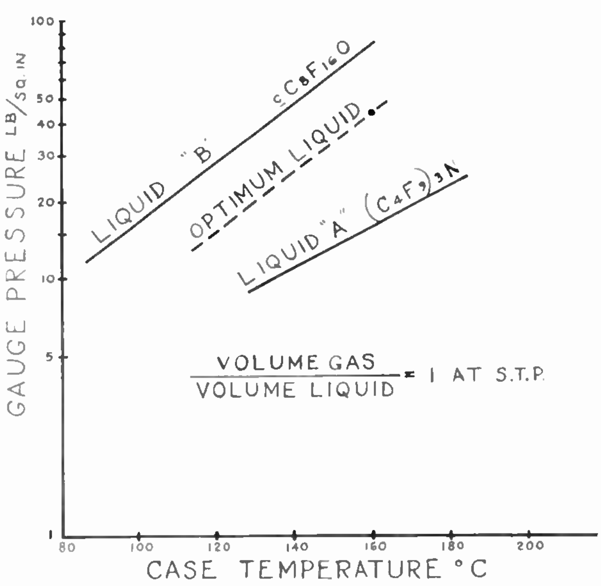


Fig. 1—Pressure curves—fluorochemicals in an enclosed container.

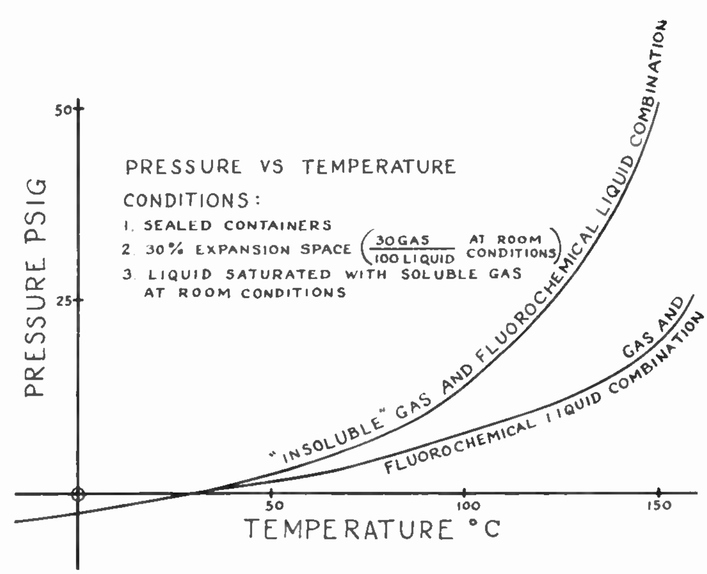


Fig. 2—Typical curve of pressure-temperature relationships of fluorochemicals showing effects of gas solubility.

[4] L. F. Kilham, Jr., and R. R. Ursch, "Fluorochemical Liquids and Gases as Transformer Design Parameters," Raytheon Mfg. Co., Waltham, Mass. IRE National Convention, New York, N. Y.; March, 1955.

[5] G. Camilli, G. S. Gordon, and R. E. Plump, "Gaseous insulation for high voltage transformers," A.I.E.E. Technical Paper, pp. 52–78, Winter General Meeting; January, 1952.

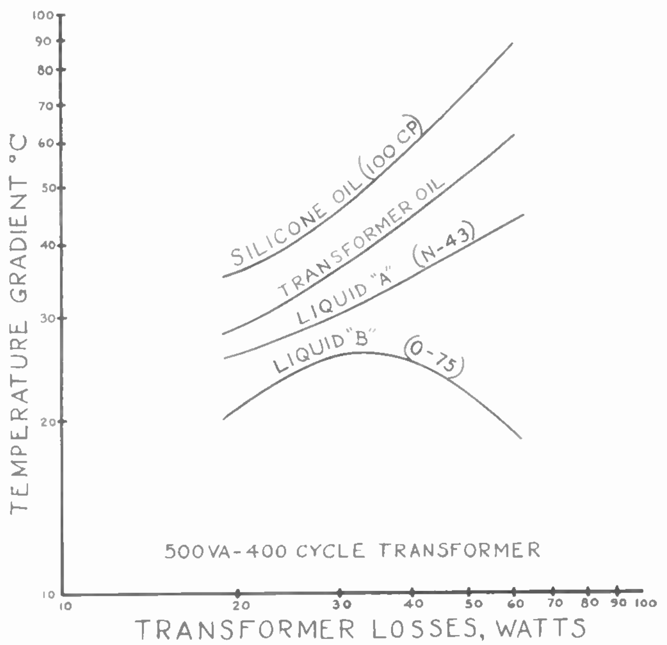


Fig. 3—Comparison of coil temperature rise above the container temperature with various dielectric liquids as coolants.

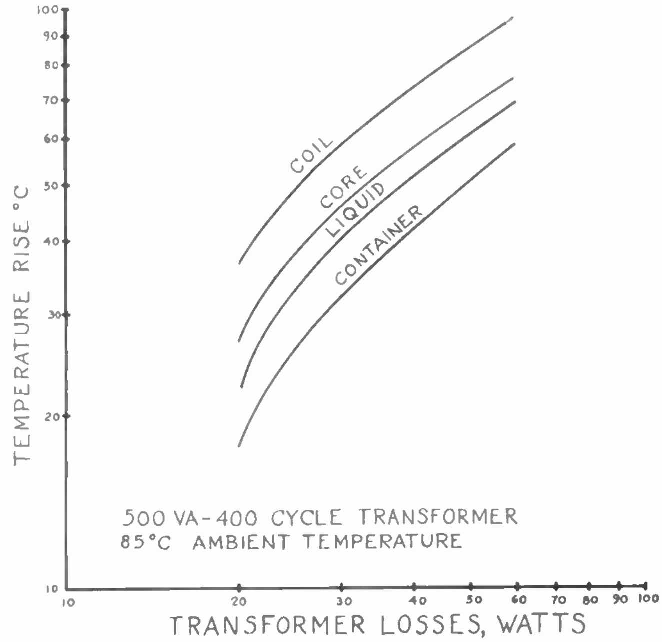


Fig. 4—Temperature gradients in a transformer with liquid *A* as the coolant.

ing.[6,7] Heat transfer characteristics of the fluorochemical coolants are compared in Fig. 3 to usual transformer-fill liquids whose characteristics are known. Using a transformer enclosed in a test cell with the liquid dielectric coolants under investigation, the cooling efficiency of the various liquids is tabulated.

A plot of coil temperature rise above case temperature vs watts input shows that for silicone oil (100 centistokes), the gradient between the highest measured coil temperature and the case temperature is about 85°C (at 60 watts loss), and for fluorochemical *B* (lower curve), the comparative gradient is only 20°C.

Additional test information shows that the temperature gradients existing between various parts of the transformer are small when using fluorochemical liquids as compared to conventional transformer-fill materials (Fig. 4). It should be noted that the internal gradients are low, with most of the temperature rise resulting from the high external gradient.

Conduction plates of copper[8] have been used successfully in many dry-type transformers of moderate voltages to transfer heat from the windings to an external radiator. The copper conducting plates are an integral part of the coil construction and consist of copper foil of sufficient thickness for proper transfer of the heat to be dissipated. Fig. 5 shows this in detail. The plates may serve as electrostatic shields in the coils and as thermal conductors around the core. From the relatively high thermal conductivity coefficient of copper (9.6 watts/in²/in/deg C), it is obvious that copper of relatively small cross section can reduce the gradient between the winding and the external radiator. When such a transformer is mounted on a sink[9] heat transfer is considerably increased.

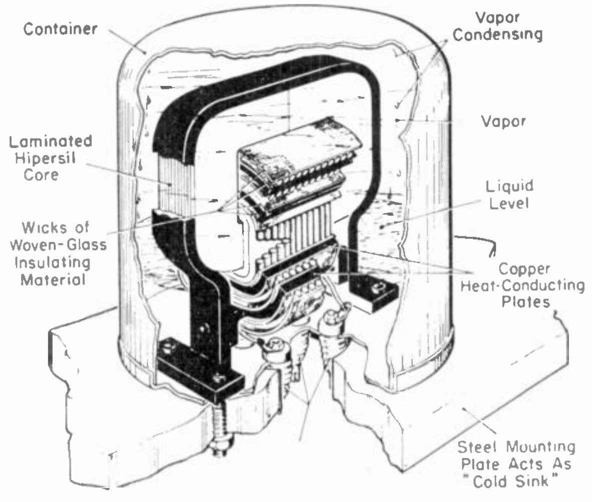


Fig. 5—Cutaway view of a typical Raytheon fluorochemical transformer.

Considering the dielectric properties and the heat transfer abilities of fluorochemicals together with the advantages of conduction techniques, new design methods have been devised to obtain minimum size and weight in specific transformers. A composite sketch showing a combination of several techniques is illustrated in Fig. 5.

[6] N. M. Bashara, "Some fluorochemicals for electrical applications," Proceedings of Symposium on Progress in Quality Electronic Components, Washington, D. C.; May, 1952.

[7] N. M. Bashara, "Some fluorinated liquid dielectrics," *A.I.E.E.;* vol. 72, Part 1, *Communications and Electronics*, pp. 79–85; January, 1953.

[8] J. P. Welsh, "A guide manual of cooling methods for electronic equipment operations," Cornell Aeronautical Lab., Inc., Report #HF-710-D-16, pp. 63, 64.

[9] G. Mathis, "Air cooled chassis design and application," Raytheon Manufacturing Co. Second Conference on Cooling of Airborne Electronic Equipment, Ohio State Univ.; June, 1953.

These techniques utilize the following desirable properties of fluorochemical liquids, vapors, and gases: adequate dielectric strength, nontoxicity, non-flammability, low K for high frequency response, corona inhibition, thermal stability at high temperature, ability to self-heal, and excellent heat transfer characteristics. It should be realized that these techniques when applied to a specific design should be selected in view of the specific conditions required in that design. For example, the design of an airborne component stressing light weight would select different techniques from a design in which volume, not weight is the governing item. Several design approaches, together with brief explanations, are listed below.

Liquid and/or Gas Fluorochemical Fill in Conjunction with Heat Conductor Plates

Dry-type transformers utilizing copper heat conduction plates similar to those shown in Fig. 5 are currently in production. Although considerable heat can be removed by this method resulting in a reduction of size, this technique has been somewhat limited because of dielectric and corona problems at higher voltages. Further size and weight reduction is possible when volatile fluorochemical liquids, vapors, or gases, with their excellent dielectric properties, are used in an enclosed system. A specific design application of this method will be detailed.

Combination or "Tailor-made" Fluorochemical Liquids and Gases

The fluorochemical liquids presently available and suitable for transformer designs vary in volatility. In most instances, the combination of operating ambient and watts-per-unit-area dissipation surface is such that the most suitable liquid for cooling would not be used because its volatility might give pressures in excess of desired container design limitations. A liquid of less volatility, however, gives less cooling. It is possible by a proper mixture of the two liquids to calculate an optimum dielectric coolant consistent with desired pressure limitations.[10]

The following formula is used for this calculation:

$$\frac{W_B}{W_A} = \frac{(P_0 - P_A)}{(P_b - P_0)} \frac{M_B}{M_A}$$

W_A = weight of less volatile liquid
W_B = weight of more volatile liquid
M_A = molecular weight of less volatile liquid
M_B = molecular weight of more volatile liquid
P_A = vapor pressure of less volatile liquid at the operating temperature
P_B = vapor pressure of more volatile liquid at the operating temperature
P_0 = maximum vapor pressure of the optimum liquid at the operating temperature.

[10] L. F. Kilham, Jr., and R. Ursch, *op. cit.*

Reference to Fig. 1 shows that an optimum liquid in which it is desired to keep pressure limits within 45 psig at 160°C for the conditions specified. Such a liquid mixture would have cooling properties intermediate between liquids A and B in Fig. 3.

It is possible to mix two or more gases to provide a gas solubility coefficient resulting in desired pressures consistent with adequate dielectric strength over the specified operating temperatures. Liquid combinations in conjunction with gas combinations to give optimum heat transfer and adequate dielectric protection within desired pressure limits may be selected.

Partial Fill and Use of Wicking Action of Liquid Fluorochemicals

The exceptional wetting action of liquid fluorochemicals is used to conduct these liquids through an inorganic wick to the interstices of the coil. Upon reaching the "hot spots" vaporization of the liquid occurs. The vapors rise to the upper portion of the container, where upon coming in contact with the inner container surface which is at a temperature below the saturation point of the vapor, condensation occurs and a major portion of the heat energy acquired by the vapor is given up to the surface of the container. Fig. 5 details a typical transformer showing the liquid sump, the inorganic wicks, and the vaporization and condensation of the liquid.

The heat of vaporization of the fluorochemicals used are in the order of 16.5 to 20.9 cal/gm while the specific heat of conventional transformer oils is around 0.5 cal/gm. Therefore, the quantity of heat that can be transferred by the vapor phase heat exchange method is roughly 30 to 40 times greater per gram of coolant than that of the quantity removed by an equal weight of nonvolatile liquid.

The partial fill and wick action method provides minimum cost by the use of a minimum amount of liquid, adequate heat transfer as described, considerable dielectric protection from the vapors under operating conditions, and reduction in weight and size of the transformer.

Volatile Fluorochemical Liquid Fill

The transformer is completely immersed in a liquid fluorochemical. Heat transfer is accomplished by the high auto-convection of the liquid plus the ability of the volatile liquid vapors in the "void" space to transfer heat to the case. Both methods of heat transfer are considerably more effective than those in a comparative transil oil transformer.

Techniques Applied

The use of fluorochemical liquid and heat conduction techniques may be applied to many transformer problems. The proper techniques to be used will depend upon the particular design problem under consideration. One or more of the above-mentioned techniques may be used in an optimum transformer design for a specific set of operating conditions. Following is a design approach to a specific transformer problem.

Specifications of transformer to be redesigned for minimum size and weight are given below.

Ambient temperature of 125°C
Chassis mounting to provide a cold plate of 100°C
Frequency—60 cycles
Primary voltage—117 v rms
Sec. 5,600 v rms ct
Load: Full-wave bridge rectifier with an RC filter
Secondary current—105 ma.

The techniques selected are shown in the composite sketch of Fig. 5. Partial fill and wick action were used to meet the objective of minimum weight. Copper heat conducting plates were also used to augment heat removal to the cold plate. A transformer designed, tested and constructed following these techniques shows a size reduction, when compared to a conventional transformer built to the same specifications, of approximately 2 to 1 by volume (see Fig. 6). Comparative ability to transfer heat of the various techniques is shown in Fig. 7.

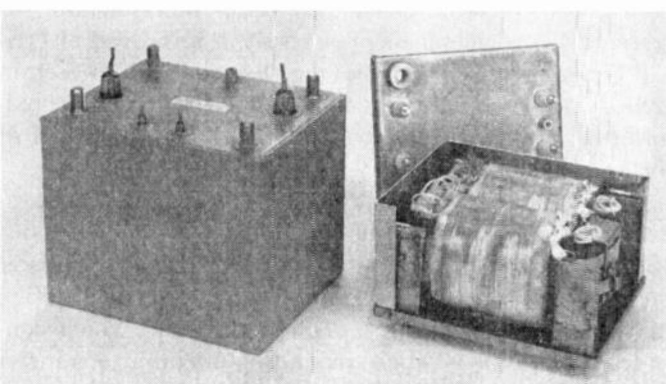


Fig. 6—Size reduction of a transformer employing fluorochemical liquid and heat conducting tabs.

Considering another transformer redesign problem, a unit of the following requirements was redesigned. This transformer is a low-capacity, high-test voltage magnetron filament transformer in which it is desired to reduce size, provide a hermetic enclosure, reduce corona, and provide equivalent electrical performance. The operating ambient is 125°C. The electrical specifications for the transformer are as follows:

Primary voltage: 115 volts rms 400 cycles
Secondary voltage: 3.7 volts rms
Secondary current: 44 a
Capacity between secondary and primary tied to ground: less than 20 micromicrofarads
Secondary test voltage: 43 kv peak.

The use of fluorochemical liquids or gases as a fill material in this particular design has certain dielectric advantages in addition to heat transfer possibilities. The low dielectric constant and high dielectric strength are particularly desirable properties. In the original design of this transformer, there is an air space between the high-voltage winding and ground which must be maintained to meet high-voltage requirements and maintain, at the same time, a minimum capacity. If a dielectric material such as transil oil with a higher dielectric strength replaces the air, the spacing may be reduced without affecting the dielectric requirements. However, the reduced spacing, together with the greater dielectric constant of transil oil, increases the capacity.

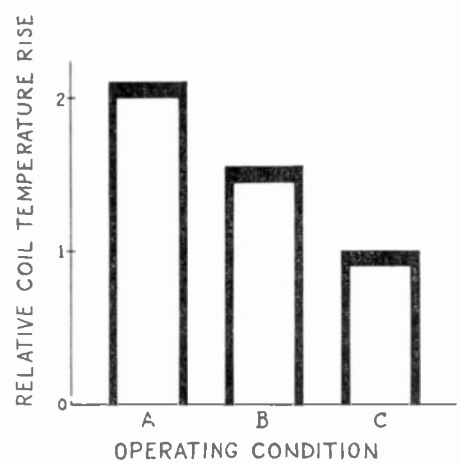


Fig. 7—Comparison of temperature rises using various heat transfer techniques. A. "Miniaturized" transformer-heat run per mil-T-27A. B. Condition "A"+conduction techniques+cold sink. C. Condition "B"+partial fluorochemical fill+wick action. 500 VA, 60~plate transformer—volume 80 cubic inches.

In this redesign, the use of a fluorochemical liquid dielectric coolant allows the secondary coil to become smaller in size. This, together with the lower K of the fluorochemical liquid, results in lower capacity. Thus, the combination of low dielectric constant, high dielectric strength, and excellent cooling properties opens the way to the miniaturization of high-temperature transformers.

The reduction in volume of the core and coils in this particular design is approximately 8 to 1. Including the hermetic enclosure to seal the transformer liquid system, a reduction of 4 to 1 by volume as well as 2 to 1 by weight is achieved.

The hermetic enclosure in this instance is a ceramic housing of high-density alumina which also serves as the high-voltage terminal. Fig. 8 shows the reduction of size.

Tests of this design indicate that the unit is capable of operating in dead-air ambients of 140°C with a hot-spot rise of 35°C. The high-voltage winding capacitance to ground is 15 mmf. Repeated dielectric breakdowns of over 50 per cent above rated test voltages do not lower the breakdown strength of the liquid. Electrical performance is excellent, including considerable reduction in corona. The maximum operating internal pressure at full load, 125°C ambient, is less than 12 psig.

A less expensive version of this design, shown in Fig. 9 can be used with lower operating ambients (up to 85°C).

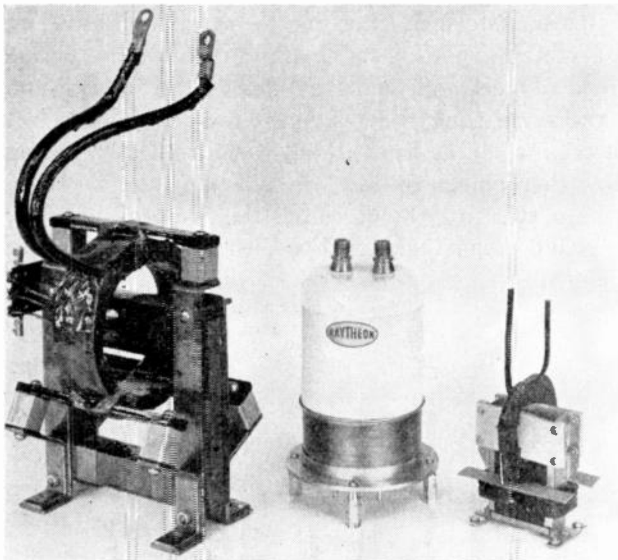


Fig. 8—Size reduction of a fluorochemical-liquid filled magnetron filament transformer.

Fig. 9—A modified version of the Raytheon fluorochemical filled magnetron filament transformer with minimum fluorochemical requirements.

The primary coil and core is embedded in a resin, while the secondary coil, together with a small amount of fluorochemical liquid, is enclosed in a fabricated shell. The temperature rise is within the limitations of the materials used. Capacity and dielectric strength are adequate.

Conclusion

Size and weight of many electronic-type transformers may be considerably reduced by utilizing fluorochemical liquids and/or gas and conduction design techniques, at operating temperatures up to 200°C. Fluorochemical liquids have electrical properties which are equivalent to the lower-operating-temperature transil oils. These properties, together with excellent heat transfer ability of the conduction plates and the fluorochemical liquids, lead directly to smaller and lighter transformer designs. Utilization of optimum heat transfer mechanisms such as heat sinks in the equipment design results in further transformer size and weight reduction.

The low dielectric constants, the self-healing ability, and the corona inhibiting properties of liquid and/or gaseous fluorochemicals may result in improved electrical performance. By employing the solubility of certain gases in selected fluorochemical liquids as demonstrated, internal pressures may be minimized and the need of bellows, pellets, etc., may be eliminated.

It is important to note that some recent military specifications are considering not only the ambient, but also the "ability of the surrounding medium to transfer heat."[11] To make proper use of these described design techniques, a "coefficient of heat transfer" which considers the ability of the surrounding medium to transfer heat, in addition to the usual statement of ambient, is required. Such a coefficient would consider the marked difference in ability to dissipate heat between still air at 100°C and a 100°C heat sink.

Utilization of fluorochemical liquids and conduction techniques by transformer designers, when combined with improved external heat transfer systems of equipment designers, points the way to transformer "miniaturization."

[11] L. F. Kilham, Jr., "Prepared discussion," Proceedings 1954 Electronic Components Symposium, Washington, D. C.

IRE Standards on Electron Devices: Definitions of Terms Related to Storage Tubes, 1956*

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* Approved by the IRE Standards Committee, December, 1955. Reprints of this Standard, 56 IRE 7.S1 may be purchased while available from the Institute of Radio Engineers, 1 East 79th Street, New York, N. Y. at $0.25 per copy. A 20 per cent discount will be allowed for 100 or more copies mailed to one address.

Definitions

Blemish (in charge-storage tubes)

An imperfection of the storage surface which produces a spurious output.

Charge Storage Tube

A *storage tube* in which information is retained on a surface in the form of electric charges.

Cloud Pulse (in a charge-storage tube)

The output resulting from space charge effects produced by the turning on or off of the electron beam.

Decay (in charge-storage tubes)

The reduction in magnitude of stored charge by any cause other than *erasing*.

Decay Time (in a charge-storage tube)

The time interval during which the magnitude of the stored charge decays to a stated fraction of its initial value.

Note: The fraction is commonly $1/e$, where e is the base of natural logarithms.

Erase (in charge-storage tubes)

To charge or discharge *storage elements* to eliminate previously stored information.

Erasing Speed (in charge-storage tubes)

The rate of *erasing* successive *storage elements*.

Hold (in charge-storage tubes)

To maintain *storage elements* at equilibrium potentials by electron bombardment.

Level (in charge-storage tubes)

A charge value which can be stored in a given *storage element* and distinguished in the output from other charge values.

Memory Tube

Deprecated (see *Storage Tube*)

Prime (in charge-storage tubes)

To charge or discharge *storage elements* to a potential suitable for writing.

Priming Speed (in charge-storage tubes)

The rate of *priming* successive *storage elements*.

Read (in charge-storage tubes)

To generate an output corresponding to the stored charge pattern.

Read-Around Number (in charge-storage tubes)

The number of times *priming*, *writing*, *reading*, or *erasing* operations can be performed on *storage elements* adjacent to any given element without loss of information from that element.

Note: The sequence of operations should be specified.

Read-Around Ratio (in charge-storage tubes)

Deprecated (see *Read-Around Number*)

Read Number (in charge-storage tubes)

The number of times a *storage element* is *read* without rewriting.

Reading Speed (in charge-storage tubes)

The rate of *reading* successive *storage elements*.

Redistribution (in charge-storage tubes)

The alteration of charges on an area of a storage surface by secondary electrons from any other area of the surface.

Regeneration (in charge-storage tubes)

The replacing of charge to overcome *decay* effects, including loss of charge by *reading*.

Retention Time, Maximum (in charge-storage tubes)

The maximum time between *writing* into and *reading* an acceptable output from a *storage element*.

Spill (in charge-storage tubes)

The loss of information from a *storage element* by *redistribution*.

Storage Element (in charge-storage tubes)

An area of a storage surface which retains information distinguishable from that of adjacent areas.

Note: The *storage element* may be a discrete area or an arbitrary portion of a continuous storage surface.

Storage Time (in charge-storage tubes)

Deprecated (see *Retention Time, Maximum* and *Decay Time*)

Storage Tube

An electron tube into which information can be introduced and then extracted at a later time.

Write (in charge-storage tubes)

To establish a charge pattern corresponding to the input.

Writing Speed (in charge-storage tubes)

The rate of *writing* on successive *storage elements*.

A Systems Approach to Electronic Reliability*

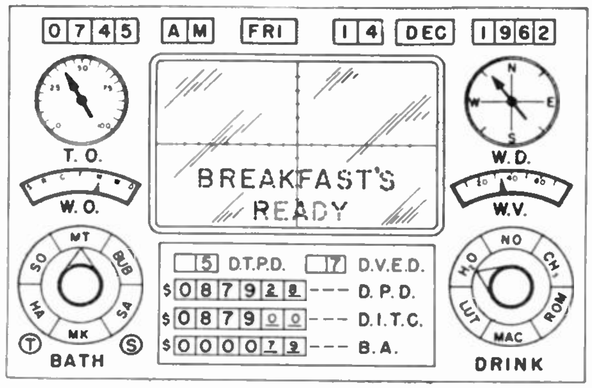
W. F. LUEBBERT†

Summary—**Certain broad general principles of electronic reliability become obvious when reliability is viewed as one of a number of important aspects of the broad general problem of achieving most efficient and effective performance of some desired task using electronic equipment. 1) The reliability of an electronic device or system depends not only upon the intrinsic properties of the device, but also on the tasks it is required to perform, the characteristics of the demand for performance, and the conditions under which successful performance of the tasks is demanded. 2) The over-all reliability objective is not merely the prevention of failures; instead it is the achievement of successful system performance. 3) Failure is important only because of its effect on the external situation, and it is not important in itself.**

These principles are perhaps so self-evident as to sound like platitudes, yet their implications are seldom fully recognized. Accepting them as valid leads to the conclusion that to attempt to insure success merely by improving the failure characteristics of individual parts of an equipment or system is to neglect many fascinating and attractive approaches to reliability improvement. The achievement of successful, failure-resistant systems is the type of problem suitable for attack by the classic scientific method and by specific techniques used in operations research. A rationale for the development of reliable systems is developed from such a viewpoint and presented in the form of a system development check-list.

"IT IS IMPERATIVE that electronic equipment be reliable, but it is not. For example, a study made in 1950 showed that only 33 per cent of Navy electronic equipment was operating satisfactorily, and later reports indicated that the situation was getting worse."—*Extracted from Report EL200/17, Research and Development Board, Department of Defense.*

Unreliable electronic equipment can cost time, money, and even lives. A sergeant, who once worked for the author, summed up the importance of the reliability problem when he said: "It doesn't matter how pretty a piece of equipment is, or how well it was designed, or what the book says it should be able to do; if you can't make it work when you need it, it's just extra junk to lug around. It's worse than that; you can get hurt if you depend upon it!" The armed services need simple, rugged equipment which will work when needed. Too often the equipment they get is like the hypothetical 893 tube electronic medicine cabinet shown in Fig. 1.



(Figure reproduced with kind permission of A. S. Brown from his article, "Discussion of the complexity and unreliability of military equipment and need for simplification and increased life," TRANS. IRE, vol. PGQC-1, pp. 11–22; August, 1952.)

Fig. 1—The DIGI-TELE-WHIRL (Digital-television-whirlwind) 893 tube electronic medicine cabinet. The top line shows time, day and date. The dial at top left labled T.O. shows the Temperature Outside. The meter below it labeled W.O. indicates Weather Outside (Sunny, Rainy, Cloudy, Fair, etc.) The bath knob below this enables one to select "HA" hard water, "SA" salt water, "SO" soft water, etc. for one's bath while the push buttons T and S enable one to choose either Tub or Shower bath. Similarly the knob on the bottom right enables one to select any of a number of different liquids to drink. Above it is a meter labeled W.V. which indicates Wind Velocity and above it a dial labeled W.D. which shows Wind Direction.

In the middle of the panel, in lieu of the conventional mirror, one has a picture tube which, by a newly invented television method picks up the image in front of it and reproduces it on the face of the tube, mirror fashion. By another device the housewife's voice in the kitchen is converted to print and superimposed momentarily on the picture.

For shaving a 2 cm diameter antenna fed with one trilliwatt at a wavelength of the diameter of a whisker is used. When brought within 1 cm of the face, the whiskers disappear painlessly, likewise the hair and eyebrows. Further research is proceeding to remedy this minor deficiency. Circular polarization will be tried.

At the bottom center is a data chart. 5 DTPD indicates 5 Days Till Pay Day, 7 DVED indicates 7 Days Vacation Earned to Date, $879.28 DPD indicates $879.28 Due Pay Day, $879.00 DITC indicates $879.00 Due Income Tax Collector (Note: this bad news appears only when a special push button is depressed), and $0.97 BA indicates $0.97 in Bank Account. This latter number appears in black if + and red if −.

Most equipment manufacturers and engineers realize that the reliability of many kinds of electronic equipment is intolerably poor. There is, however, a general tendency to believe that the development of more reliable and rugged vacuum tubes, transistors and other components will provide a solution to the reliability problem, and that the current difficulties with reliability are merely "growing pains," which, though currently important, are of little long-term significance. Although there is much truth in this viewpoint, it is an oversimplification which has been used too often to justify the designing of equipment for maximum performance "to advance the state of the art" while paying only lip service to reliability.

There is, of course, good reason to believe that the current critical nature of the reliability problem in complex electronic systems is transient, but "growing pains" are often precursors of important developments and changes of viewpoint. Already, the nature and operation of some types of electronic systems have become so complex that they are often described like biological organisms. For example, it is not uncommon to hear a fire control system described in terms of radar "eyes,"

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communication "nerves," computer "brains," etc. This is but one symptom of a growing consciousness of the importance of the systems viewpoint. The developing interest in this viewpoint holds forth the hope that someday, perhaps not too far in the future, there will be a well-developed body of knowledge available to enable the designers, builders, and users of electronic equipment to make rational choices between the sometimes conflicting demands for high performance, good reliability, ease of operation, maintainability vs cost and difficulty of production, etc. When this is true, the reliability problem, *per se*, will have been solved, and only a reliability aspect of the over-all problems of system design, construction, and utilization will remain.

Every electronic product manufactured, and every electronic system built has one fundamental reason for existence: Under the conditions of actual use, it must successfully perform a set of assigned tasks when (and for as long as) it is required to do so. If the equipment or system is consistently successful, it is considered reliable; if it has a tendency to fail (perform unsatisfactorily) it is considered unreliable.

Factors Influencing Reliability

What does affect reliability? The simplest, and possibly the most accurate, answer to this question is, anything and everything; but this answer is not very enlightening because it is too broad to be useful. One can, however, obtain one possible classification of factors, a particularly convenient and useful one, by referring back to the statement of the fundamental reason for equipment existence. This statement tends to bring out the following important factors:

1) The conditions of use (environment).
2) The task to be performed (including also criteria for successful performance).
3) The inherent properties of the equipment, or system, which influence its capability to perform the assigned task under the conditions of use.
4) The frequency and duration at which successful performance is expected.

This classification brings out clearly the foundation for the first basic principle to be considered here: *The reliability of an electronic device, or system, depends not only on the intrinsic properties of the system, but also on the tasks which it is required to perform, the characteristics of demand for performance, and the conditions under which successful performance of the tasks is demanded.*

The implications of this principle are many and interesting. For example, the principle implies that improved tubes and components alone can never insure reliability, since increases in the severity of task requirements, demand for performance, and conditions of use may increase the tendency to failure faster than component improvements can decrease it by improving the intrinsic capability of the equipment to perform the task successfully. A few minutes' reflection should lead the reader to other useful corollaries to this principle.

Reliability and Systems Worth

Another important principle is the following: *Reliability cannot be considered by itself, rather it is only one of a number of factors which determine over-all system worth.* Just how many factors determine over-all system worth depends, of course, on the scheme of classification one uses. A common breakdown lists seven basic factors:

1) Performance capability.
2) Reliability.
3) Accuracy and precision.
4) Vulnerability (for military systems primarily).
5) Operability (ease of operation by user).
6) Maintainability (ease of maintenance).
7) Procurability (including producibility, availability in time, and ease of logistical supply).

This breakdown is essentially arbitrary and the factors enumerated overlap considerably. Consequently, this list may be modified, or discarded completely, and replaced by another which may prove to be more useful in a particular case. The important thing is not what the other factors are, but the fact that factors other than reliability and performance do exist and must be considered.

It is obvious from a consideration of this principle that the most effective balance of these various factors is important, rather than giving the highest possible weight to any one of them. It should also be noted that engineering practices and techniques, which produce improvements in reliability (*e.g.*, simplicity, interchangeability, etc.) will in many cases also improve operability, maintainability, and procurability. However, in many other cases, the demands of one factor (*e.g.*, high performance) may be in conflict with the demands of another (*e.g.*, high reliability). In such cases, it is obvious that it may be best to accept a low reliability, if this will help to achieve a desirable over-all objective. Conversely, it might be necessary to sacrifice some other desirable factor in order to achieve satisfactory reliability. This idea is worthy of further examination.

Assume, for example, that a manufacturer has a very large contract for the development of a radically new type of radio transmitter. He desires to develop a transmitter which has both high performance and high reliability; yet he realizes that emphasis on high performance often leads to low reliability and, conversely, that emphasis on high reliability often leads to low performance capabilities. How can he make a rational choice between these alternatives? An operations research specialist[1] might suggest that he attempt to estimate the performance and reliability attainable for different relative emphasis, and for different amounts of engineering effort expended, and plot the estimates on

[1] Operations research is a rapidly developing field which sometimes appears to contain elements of black magic, but which really deals with the application of mathematical techniques to the analysis of the over-all picture of complex situations.

a "trade-off" curve shown in Fig. 2. A typical engineering job might produce a transmitter which, when placed in a severe military environment, would have only a 50-50 chance of performing satisfactorily for a predetermined length of time. This condition would be represented by point A (probability of success 0.5; relative performance 1.0). If more emphasis were placed on reliability and less on performance (*e.g.*, by operating the tubes at lower plate dissipation, etc.), then one could "trade" performance for reliability and move to point B. If one decided to expend the time and effort to do an especially thorough and effective engineering job, one might achieve higher reliability and the same performance, point C, or higher performance and the same reliability, point D, or some other point along the new curve through these points. If such curves can be drawn, even approximately, they provide a useful basis for intelligent determination of the amount of engineering effort necessary to achieve certain results, and/or, of the relative amount of emphasis to be placed on reliability.

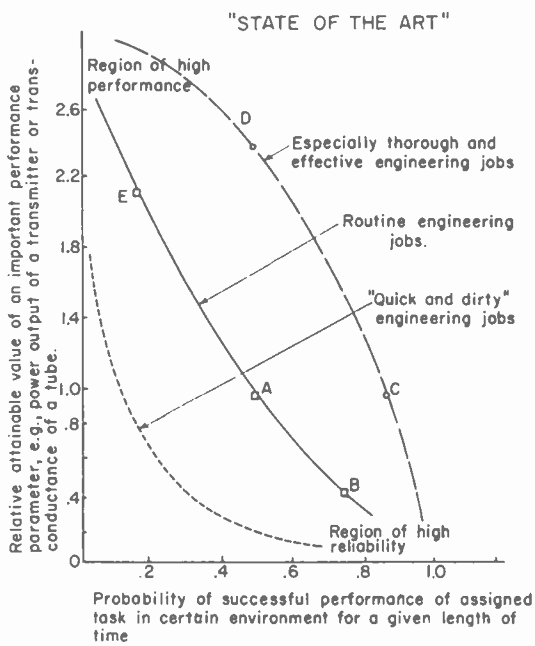


Fig. 2—Trade-off curve of performance vs probability of successful performance.

Unfortunately, there exists a strong tendency among designers to emphasize performance unduly and to end up around point E, where poor reliability may make the equipment virtually useless. The application of the trade-off analysis would prevent such overemphasis of performance and aid in making intelligent decisions. Certainly there is nothing new in comparing alternatives or in presenting the results graphically; engineers do it constantly. What is significant is that, by using an over-all viewpoint and available techniques, one can make rational decisions about the employment of engineering time and effort, rather than blindly depending upon intuition and experience to divine the "best" solution.

The application of the trade-off viewpoint is not restricted to this particular example. The trade-off curves could have been parametric in weight of equipment or of some other relevant quantity, rather than "state of the art." In fact, one need not plot the measure of reliability vs cost of the equipment. One can even use the trade-off viewpoint with a different measure of reliability than probability of success. Outside of its applications to reliability problems, the trade-off viewpoint is often valuable in many other aspects of systems engineering.

Achieving Successful Performance

Another useful principle is the following: *The over-all reliability objective is not merely the prevention of failures; instead, it is the achievement of successful system performance.* It is desirable to have a system which has no failures, but there is always some small probability of failure; consequently it is important to accept the possibility of failures in system planning. The following is a brief summary of some important ideas suggested by the above principle:

Intrinsic Survivability

A system should be designed, produced and utilized with due attention to over-all task requirements and environmental conditions so that there is a minimum number of essential parts subject to failure, and so that these parts have a high probability of performing their sub-tasks successfully.

Failure Inhibition

Full use should be made of applicable measures to inhibit failure. For example, techniques for protecting equipment from environmental extremes, utilization by personnel fully cognizant of the capabilities and limitations of the equipment, and adequate reserves of performance capability, all tend to inhibit failures in equipment.

Failure Prediction and Prevention

Provision should be made for the prediction and prevention of failures. Thus, routine operational checks, marginal checking techniques, and preventive maintenance techniques may be helpful in predicting wear-out or drift-out-of-tolerance types of failure.[2]

Failure Detection and Isolation

Provision should be made for detection of the existence of failures, and for the isolation of the most frequent and most important failures, as soon as they occur. Failure alarms, qualifying tests undertaken either by operators, or automatically by the equipment, may be useful in detecting failures, since it may not always be immediately obvious in normal operation whether equipment is operating properly.

Functional Replacement

When practical, provision should be made for the rapid replacement of defective equipment or parts by others, not necessarily identical, but capable of performing the same function. Redundancy, either in the

[2] It is important to note that routine replacement of parts (such as tubes) and tests which may cause destructive environmental stresses, may do more harm than good.

form of stand-by or alternate means may be valuable. Rapid repair which brings the faulty equipment back to serviceable condition is, of course, another method of getting the equipment back into service.

Failure Effect Minimization

Realizing that some failures will occur which cannot be corrected instantaneously, attempt to design and utilize the system so that the failures which are most likely to occur, and not be corrected, are those which have minimum effect on performance of the system task. In fact, it could even be desirable to induce additional failures of a type which can be easily corrected, or which have little effect on task performance, if by so doing you can reduce other failures which are difficult to correct and have a critical effect on task performance. Since the effects of a given type of failure may vary with time, perhaps being extreme during an operational mission and negligible during a maintenance period, it may be desirable to induce certain failures during a maintenance period, if so doing will decrease their probability of occurrence at a critical time. The task requirements of the electronic equipment or system are really subtasks of some larger system, perhaps the over-all conduct of a military operation, and can themselves be chosen so that failures of the electronic equipment or system have minimum effect on the over-all situation.

The Importance of Failure

Probably the most important principle which one should keep in mind in the study of reliability is this: *Failure is important only because of its effect on the external situation, and not important in itself.*

This principle is easily illustrated by a few examples. In a household which has two radios the failure of one will usually cause only minor inconvenience. On the other hand, the failure of an aircraft receiver, which is needed to make a successful blind landing, may cause a crash and cost many lives. This serves to illustrate the concept that higher failure rates may be tolerated for equipment which performs tasks which do not have a critical effect on the external situation than for equipment which performs critical tasks. If the aircraft had two receivers capable of receiving the blind landing instructions, and if only one fails, then the aircraft can still make a safe landing with no ill effects. This situation illustrates that higher failure rates may be tolerated in individual items of equipment, if alternate (redundant) means of performing the required tasks are available.[3]

Reliability is primarily a problem with new equipment, particularly new equipment operating in new and often unknown environments. Experience shows that the first solution to a set of requirements is usually unduly complex, marginal in performance, difficult to build and even more difficult to operate and maintain, and, in general, notoriously bad from the reliability viewpoint. Too much electronic equipment, even equipment which is used and mass-produced for years, never gets far beyond this early stage of development. Many items of equipment truly deserve the appellation "super-collossal gadgetry." The equipment is too complex; it uses too many tubes; operational tolerances are too tight; it is too expensive to build and maintain; and so on. To put it bluntly, the job of engineering the equipment has not been brought anywhere near completion.

The Scientific Method

What kind of approach can be used to encourage the development of successful, reliable equipment? An obvious and useful approach is to apply the viewpoint of the classical scientific method as indicated below.

Determination of Requirements

With due consideration to the importance of the tasks to be performed by the equipment, and the cost in time, money, manpower, etc., needed to design, produce and utilize the equipment, determine the required reliability. This, of course, will demand specification of conditions of use, of the tasks to be performed, and criteria for successful performance, and also the characteristics of the demand (duration, frequency, etc.).

Collection of Data

Data on component and system failures, and their causes, must be collected in statistically significant amounts; this also applies to data on task, environment, and demand for performance.

Analysis and Criticism

Data must be analyzed to determine whether requirements are being met, or can be met, to establish areas and causes of failure, or potential failure, and to bring out possible methods of improvement.

Innovation and Improvement

Action must be taken to use the knowledge gained to modify and improve the reliability characteristics of the equipment.

Test and Trial

Tests must be made to determine whether proposed "improvements" really do improve not only reliability, but also over-all systems value, and that they are practically realizable and economically feasible.

Surveillance and Feedback

A continuous and critical surveillance must be carried out to determine whether requirements are actually being met in the conditions of practical use of the equipment, to make sure that "improvements" improve in actual use as well as in test, to anticipate and examine new and unsuspected sources of trouble, and to review and modify requirements. The results of this surveillance must be fed back to the places where the information can be used, in a form that makes it usable.

[3] Provided, of course, that failure of the primary means does not also imply failure of the alternate means. This might be the case if the thing which causes receivers to fail is their power source and both the primary and alternate means obtained their power from the same supply.

Systems Development Check List

Taking the ideas and principles treated, thus far, and adding to them some practical experience and a generous portion of the scientific method, one can devise a rationale for the development of reliable electronic equipment and equipment systems. A convenient way to summarize this rationale, and to make the ideas in it readily available, is to recast it in the form of a check-list:

1) Is a special attempt being made to gather and collate available information about the tasks which the system is to perform, the environment in which it is to perform them, and possible methods of achieving successful performance of the task, *before* design is undertaken?
2) Is an attempt being made to anticipate what additional information (particularly about task and environment) may be helpful in setting reasonable specifications, and in selecting favorable design alternatives? Are appropriate experiments being designed and conducted to obtain such data in statistically significant amounts?
3) Is a system analysis carried out to determine what tasks are most important to the over-all objective to be attained, to determine what reliability is necessary, and to assist in determining the design alternatives most likely to achieve the objective (taking cost, weight, and other limitations into account)?
4) Is the system so designed that it can perform the assigned tasks with appropriate margins of safety with respect to both task and environment?
5) Is due attention paid to prediction of environmental extremes, and to protecting the system against their effects?
6) Is a conscious effort made to make the system simple and to resist any tendency toward gadgetry?
7) Is the system so designed that the humans who build, use, maintain, repair, and/or replace it find their tasks simple and foolproof? Has information about their training and skill been considered in the equipment design?
8) Does the system utilize standardized, interchangeable parts and subassemblies which are resistant to failure or excessive drift in the expected environment?
9) Are the used circuits and assemblies tolerant to variations in component characteristics, and are reasonable, accurate tolerances set and enforced?
10) Is the system designed so that it is easy and convenient (or possibly automatic) to carry out preventive maintenance, marginal checking, or other tests aimed at anticipating drift-out or wear-out failures before they actually occur?
11) Recognizing that some failures are inevitable, do the conditions of use make it obvious when failures occur, or is it necessary to design "qualification" tests to determine, perhaps automatically, whether the equipment is operating properly? Is provision made for the rapid isolation of failures which do occur?
12) Are alternate or stand-by means or some other form of redundancy used, when appropriate, so that failure of part of a system does not necessarily cause failure of the whole system?
13) Are appropriate test procedures, repair and replacement routines, special test equipment when necessary, etc., developed with the system so that there is the best possible chance that defective systems will be repaired or replaced before their failure can have an appreciable effect on the external situation?
14) Realizing that all failures cannot be corrected before they affect the external situation, are the above techniques applied so that these failures which are most likely to occur, and have the greatest effect on the external situation, are guarded against, whereas least protection is provided against failures which are unlikely to occur and have little effect on the external situation?
15) Equipment must actually be produced. Is this fact considered as an integral and very important part of the design problem, and are possible production engineering aspects considered throughout the entire development process? Is full use made of mock-ups, breadboard models, pre-production models, pilot runs, etc., to improve the producibility of the system, and to aid the production engineering process?
16) Realizing that quality of design is of little value unless accompanied by quality of conformance to design, is due attention given to making quality control simple and effective? Is advantage taken of statistical techniques?
17) Is a comprehensive test program, to determine the capabilities and limitations of the system as designed and produced, considered an integral and important part of the development program?
18) Is a special attempt made to work in close coordination with the equipment user to provide him with information, and to ensure the development of effective procedures and doctrines for use of the system?
19) Is critical and continuous surveillance conducted at all stages of design, production, test, and actual field use, so that the experience gained in these processes can be incorporated into improving the system, anticipating and examining new and unsuspected applications of the system or its parts, anticipating and examining new or unexpected environmental stresses and their effects on the system, and reviewing and modifying requirements and design criteria for subsequent systems?
20) Most important of all, is there effective feedback of information from where it is gathered to where it can be used? Is the information fed back in a form that can be readily understood and in a

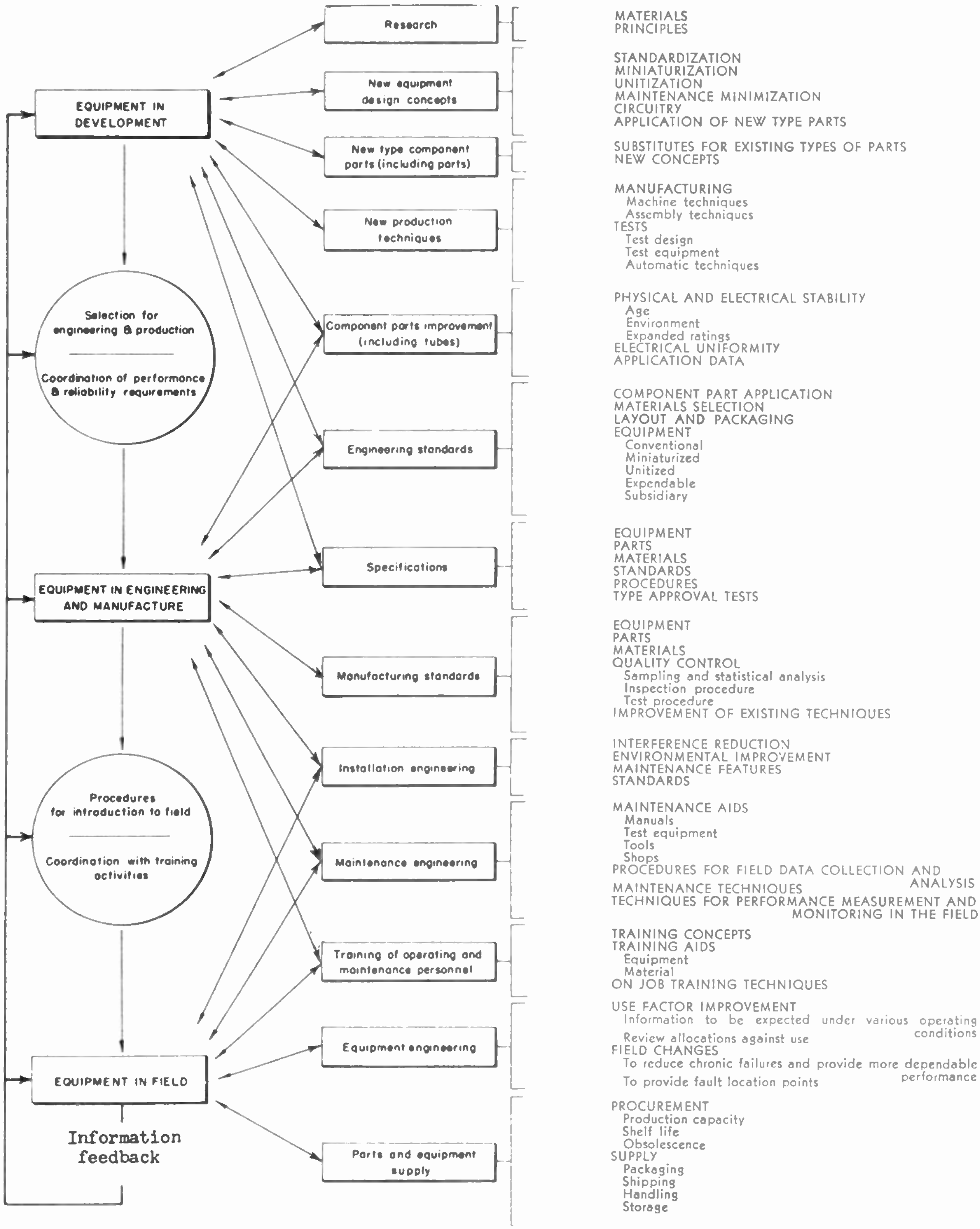


Fig. 3—Factors influencing reliability.

form where pertinent information can readily be separated from nonpertinent information?

There is no sharp distinction between this check-list and many others which deal with "good engineering practice." The differences are primarily in emphasis. However, this approach, which emphasizes over-all systems planning, the need to obtain and make use of environment and task information, and the importance of surveillance and feedback of information, points up some of the most frequently neglected aspects of the problem of developing reliable systems.

A slightly different viewpoint which shows more clearly the practical compartmentalization of effort is shown in Fig. 3 (above). This figure is a modified version of a chart which was originally prepared by the Vitro Corporation.

A Magnetic Thyratron Grid Control Circuit*

J. H. BURNETT†, MEMBER, IRE

Summary—**A new compact magnetic method for driving thyratron grids has recently simplified high-power fast response servos and regulated power supplies.**

Since large power gain is available in the thyratron, power amplification can be sacrificed in the magnetic circuit to gain speed of response comparable to the thyratron inverse cycle. The new grid circuit provides a rapid rate of rise of thyratron firing voltage, wide-control range and high-voltage amplification. It eliminates from the grid circuit most of the usual interference and "noise" pick-up. Dc or ac input signals to the magnetic circuit work equally well, and the signal voltage source can be isolated from the power circuit.

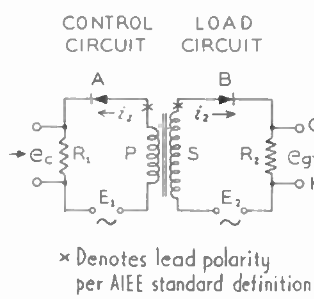
A NEW, compact magnetic method for driving thyratron grids, which simplifies the design of high power, fast response servos and regulated power supplies, is described in this paper.

An earlier grid circuit[1] produces a phase shift of firing voltage by change of the half cycle flux storage in the core of a small transformer, through the primary of which is passed half-wave rectified current that is varied by either dc or ac command voltages. The new grid circuit (see Fig. 1(a)) which provides a greatly improved rate of rise of grid firing voltage, differs from the earlier circuit by the addition of a source of ac supply voltage in the secondary circuit of the grid transformer and a blocking rectifier, changing the mode of operation to that of a reset magnetic amplifier.[2]

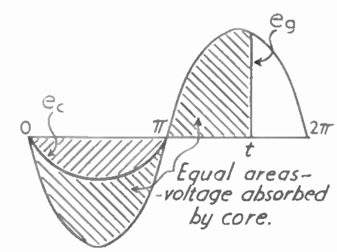
This device is ideal for providing a 180 degree range of firing voltage for thyratron grid drive. The combination with a thyratron gives enormous power gain in a rugged, compact system with one cycle response.

The inherent integration of the command signal over one-half cycle gives some surprising economic results, other than the obvious ones due to isolation of command voltage circuits from the power circuits, the ability to use either ac, or dc, or combination commands, and the ability to work from low impedance sources with high amplification. This one-half cycle integration is long enough to filter out all of the usual sources of "noise," stray pickup, and false feedback in the grid circuit that has accounted for so much of the engineering time necessary in the past to perfect a thyratron grid circuit, and make it commercially reliable. The usual interference from transients, especially those created in the power circuit by firing of other thyratrons in the unit, by commutators, by power circuit capacitance to ground, moderate line voltage or command voltage distortions, are all automatically removed without the penalty of slow response time. Hence, all of the stabilization and precision deterioration problems that result from introducing delays into the response time are minimized.

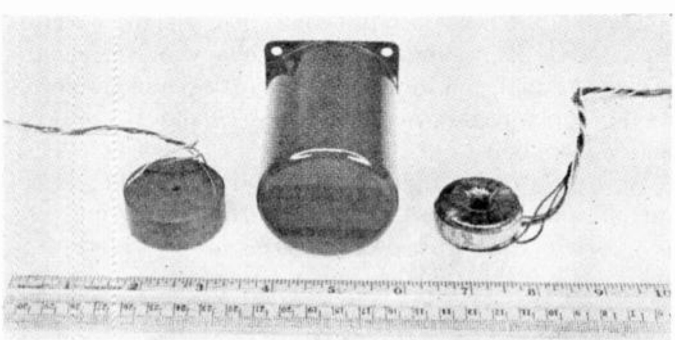
This apparently small change has made a vast improvement in the engineering aspects of thyratron circuit design. Once the power circuit has been carefully checked out to be sure no stray voltages are finding their way into the circuit between the magnetic grid circuit output and the thyratron grid, the unit may become a packaged "building block" type of equipment that may be engineered into a wide variety of input circuits without the necessity of carefully checking out a breadboard model with "cut and try" methods to locate and eliminate the sources of the various stray voltages picked up.



(a)



(b)



(c)

Fig. 1—(a) Basic circuit. (b) Idealized waveform. (c) Left: Grid transformer of Delta-max toroidal core, design details of which are described in this paper. Center: Complete packaged unit of another manufacturer, containing supply transformer, grid transformers blocking rectifiers and resistors, for two thyratrons. Right: Grid transformer of Delta-max toroidal core, wound for 180 degree control with 0.7 volts change in command voltage.

This is true even on systems where the command voltage has been amplified up from extremely low power levels.

Packaged units are small ($1\frac{3}{4}$ inches diam. × 2 inches). Fig. 1(c) shows a unit with complete grid circuits for two thyratrons.

Theory of Operation

The magnetic grid circuit consists of a small transformer, two small dry rectifiers and ac supply voltages for the control and for the output windings. Operation of the circuit is illustrated in Fig. 1(a). The magnitude of

* Original manuscript received by the IRE, May 13, 1955; revised manuscript received October 17, 1955. This paper was presented at the IRE National Convention, March 21, 1955.

† Electrons, Inc., Newark, N. J.

[1] J. H. Burnett, "Thyratron grid circuit design," *Electronics*, vol. 24, no. 3, p. 106; March, 1951; corrections, *Electronics*, vol. 24, no. 4, pp. 304–305; April, 1951.

[2] R. A. Ramey, "On the mechanics of magnetic amplifier operation," *Trans. AIEE*, vol. 79, part II, pp. 1214–1223; 1951.

the current which flows in the command winding P of the grid transformer is varied by changing resistance R_1, or by applying a varying dc or half-wave rectified ac signal voltage in series with source E_1. Neglecting winding resistance and the voltage drop across rectifier A, the time integral of the voltage per turn absorbed by the winding P during this half cycle is proportional to the volts-seconds change in the state of the core away from saturation. It may be written:

$$\int_0^{\pi} V_1\, dt, \quad \text{where} \quad V_1 = \frac{E_1 - e_c}{\text{Turns of winding } P}.$$

Polarities of supply voltages E_1 and E_2, and windings P and S, are such that during this half cycle when command circuit current is flowing, rectifier B blocks current flow in the grid load circuit. During the succeeding half cycle, the source E_2 furnishes voltage which is absorbed by winding S until the core is flux saturated, at which time, t, appreciable current can start to flow in the output resistance R_2. Neglecting the voltage drop of magnetizing current in S and R_2 and rectifier B, the available flux setting voltage per turn of the secondary is

$$V_2 = \frac{E_2}{\text{Turns of winding } S}.$$

then

$$\int_{\pi}^{t} V_2 dt = \int_0^{\pi} V_1 dt.$$

These equal areas are indicated in Fig. 1(b) which is drawn with the assumption that primary and secondary turns are equal. For different turns ratios, and the corresponding changes in magnitudes of E_1 and E_2, the reasoning is similar.

With relatively high impedance loading the rise of current and hence that of the thyratron grid firing voltage *e.g.* which is developed across R_2, is very rapid up to nearly the instantaneous value of supply voltage, less the drop across rectifier B and the resistance drop across S when flux saturated, and then follows the sine wave shape of this voltage. The firing angle is varied by change of command voltage, or by change of command circuit resistance.

Typical Design

In a typical design of the magnetic grid circuit, the grid transformer may consist of a Delta-max toroidal core as pictured in Fig. 1(c). The control winding turns would be chosen for the desired sensitivity within the limits set by the command circuit internal impedance. For illustration, a 12 volt control winding was chosen in the unit to be described. The output winding is wound for 120 volts which provides satisfactory firing voltage for most thyratrons over a wide range. Then the source voltage E_1 must be 12 volts, and E_2 120 volts rms. A grid load resistance R_2 in the order of 20 K may be used, but the exact value is not critical. Rectifiers with high back resistance are most desirable to minimize undesired flux resetting. In the low voltage control circuit, a low forward drop rectifier is preferred, such as the IN92 Germanium diode. A selenium type 9GA8 may be used as rectifier B.

A total change of 10 volts rms across the command winding is required to obtain the full 180 degree range of grid firing voltage. The maximum command winding current is 1.3 ma at zero output voltage of the thyratron. This current decreases with increase of command voltage e_c. For the transformer described, if e_c is zero, the resistance R_1 is 1,300 ohms or less for zero thyratron output voltage. Increasing resistance R_1, with e_c equal to zero, advances the firing angle.

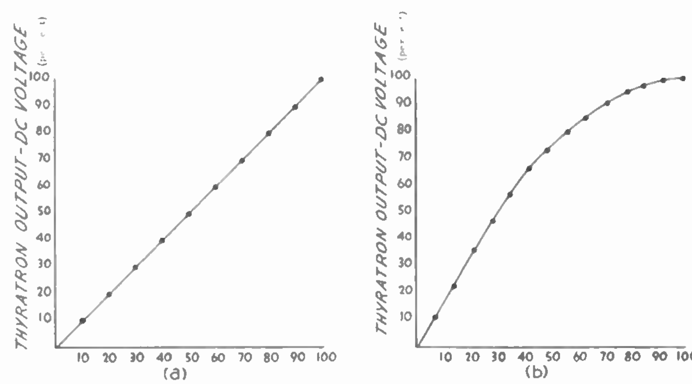


Fig. 2—(a) Ac control voltage (per cent). (b) Dc control voltage (per cent).

The response of this magnetic grid circuit is shown in Fig. 2 in per cent of thyratron full dc output voltage vs per cent of ac and dc signal voltages required to obtain full thyratron output voltage.

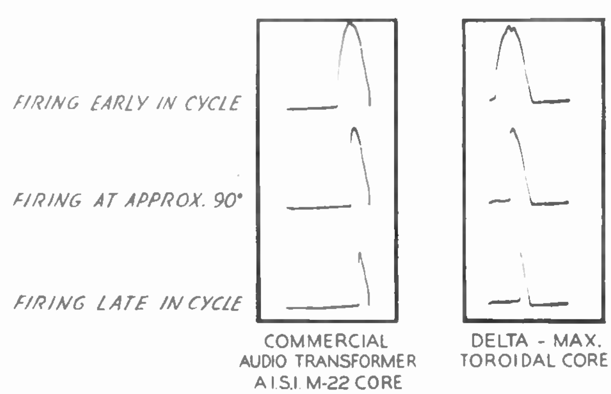


Fig. 3—Comparison of output waveform with different core material.

Typical oscillograms of the waveform of the output voltage of this magnetic grid circuit at various signal input voltages, are shown in Fig. 3.

Since most modern thyratron tubes have slightly negative grid characteristics, it usually is necessary to provide a negative hold-off bias for the thyratron grid.

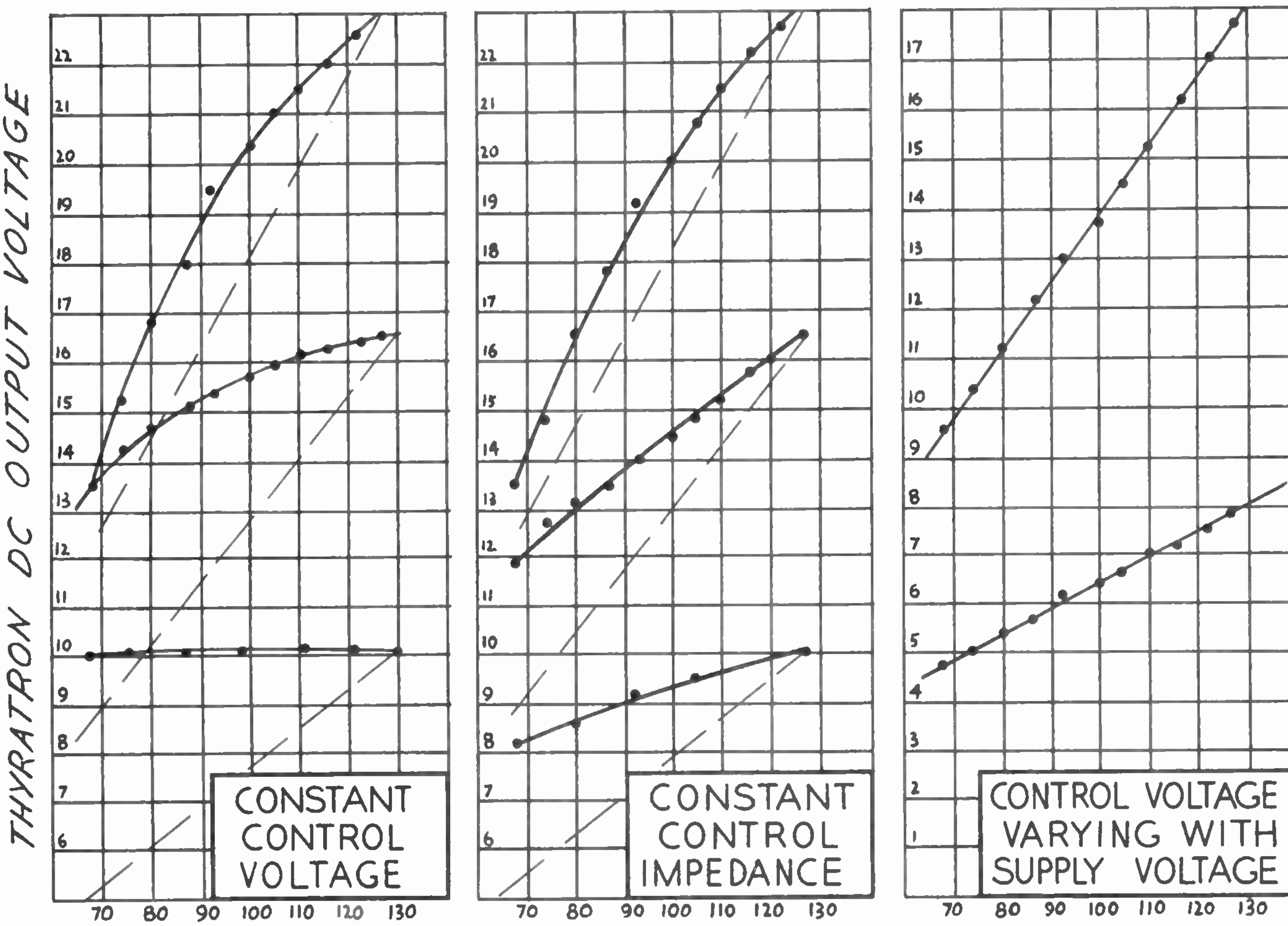


Fig. 4—Effect of line voltage variation on thyratron output voltage with magnetic grid circuit. Dotted curves indicate output for constant firing angle. Horizontal axis is line voltage in per cent of normal.

As the magnitude of dc negative bias is increased, the range of control of firing angle is obviously decreased unless the output voltage of the magnetic grid circuit is also proportionally increased. Thus for any design, the magnitude of dc negative bias used sets a limitation on the realizable range of thyratron firing angle with the magnetic grid circuit. For example, in the design described, a 10 volt bias negative to the critical grid voltage reduces the maximum retard angle by 3.5 degrees.

The magnetic grid circuit is independent of line frequency variations. It is relatively insensitive to line voltage variations, since the control and grid load circuit supply voltages are affected in the same proportion. The self-regulating characteristic of the magnetic grid circuit for various constant signal voltages is shown in Fig. 4(a). The effect of line voltage variation with various constant signal circuit resistance values, and with various signal voltages that are proportional to line voltage are shown in Figs. 4(b), and 4(c), respectively.

As shown in Fig. 5 a small capacitor (500 $\mu\mu f$ in this case) should be connected directly from grid to either side of the filament, as near the tube as possible. This capacitor limits to an insignificant magnitude voltages induced directly on the grid through interelectrode capacitance of the thyratron, resulting from transient changes in anode voltage. It also serves to minimize any unwanted voltages appearing on the grid through coupling to the supply line across stray capacitance of grid supply and filament transformers. Limiting of reversing ac or dc signal voltages can be simply accomplished by connecting a diode in series with signal voltage.

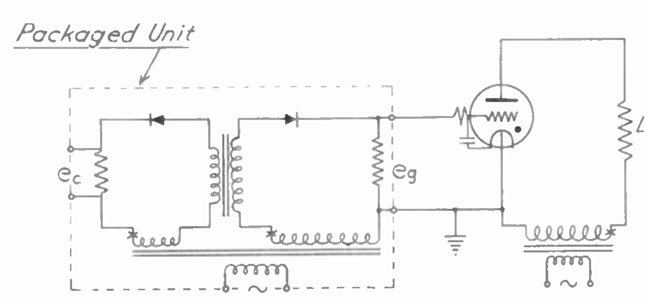


Fig. 5—Packaged magnetic grid circuit for thyratron amplifier.

For thyratron bridge circuits it is often necessary to provide simultaneous firing pulses for two tubes which conduct in series. In this case a single winding reactor may be utilized as shown in Fig. 6. An output transformer may be connected across R_2 with separate secondaries supplying tubes to be fired at same instant.

A transistor connected directly in the control circuit as shown in Fig. 7(a) may be used to obtain an increase of sensitivity of the order of 100. With the addition of a 2N34 to the unit described in detail above, a change in signal voltage of 0.13 volt between base and emitter of the transistor is sufficient to shift the output of the magnetic grid circuit 180 degrees, a sensitivity of 0.001 volt/degree. In a full wave rectifier circuit, a single transistor is used to control two grid circuits and hence any drift in the transistor affects both tubes equally, changing the loop gain, but not unbalancing the firing angles.

The realizable power gain per cycle of this transistor magnetic grid circuit and two C16J thyratrons in a full wave circuit is of the order of a billion!

Fig. 7(b) shows a typical vacuum tube cathode follower input to the magnetic grid circuit.

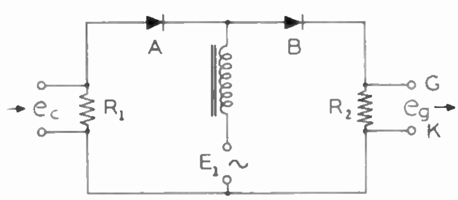


Fig. 6—Single winding circuit.

Use of Commerical Audio Transformers

Many commercial audio transformers can be used in this grid circuit. Fig. 3 provides a comparison of the output voltage wave forms obtained with Delta-max toroidal core units, and those obtained with standard commercial audio transformers having A.I.S.I. M-22 steel cores, each having an output grid load resistance of 100 *K*. Other standard audio transformers with M-15 steel cores work equally well.

The circuit of Fig. 6 illustrates a convenient arrangement for exploring quickly whether a particular transformer has a suitable core for use in this application. A potentiometer is substituted for R_1, e_c is omitted, and e_g connected to an oscilloscope. Although standard audio transformers which have been so used exhibit a control range less than 180 degrees, a small dc positive bias voltage in series with the command circuit can be added in some cases to obtain a full 180 degree range of magnetic grid circuit output voltage with commercially available types of transformers.

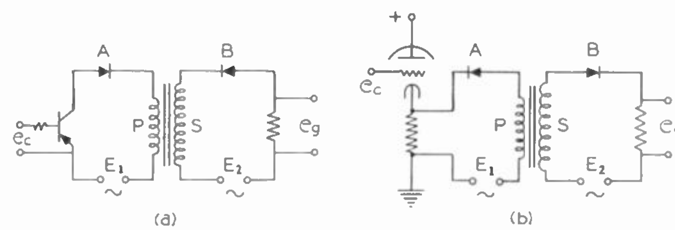


Fig. 7—(a) Transistor pre-amplifier. (b) Vacuum tube input.

Applications

Application of the new grid circuit has been made in 1) Single phase full-wave rectifiers for regulated dc voltage supply. 2) AC current regulated supply. 3) Half-wave reversing dc armature voltage control for positioning a dc servo motor with ac error voltage from synchros, together with dc tachometer feedback voltage. 4) DC shunt motor speed control units up to 25 hp.

With the sacrifice of only one cycle in response speed compared to the instantaneous response that is possible with thyratron units, the magnetic thyratron grid circuit eliminates most of the painstaking laboratory engineering which has formerly been needed for the design of commercially reliable thyratron amplifiers.

Correction

The following correction to **IRE Standards on Terminology for Feedback Control Systems, 1955,** which appeared on pages 107–111 of the January, 1956 issue of Proceedings of the IRE has been brought to the attention of the editors.

The *Note* following *Definition 2.1.6* for **Loop Return Signal** should read:

"The loop return signal is a specific type of loop feedback signal."

Analysis of a Regenerative Amplifier with Distributed Amplification*

B. S. GOLOSMAN†

Summary—**It is shown that regenerative circuits of the discontinuous-output type such as the monostable multivibrator do not have their high speed performance improved by the use of distributed amplification in the amplifier portion.**

Introduction

IN BOTH REGENERATIVE and nonregenerative amplifiers the high speed performance is proportional to the transconductance/capacity ratio of the tubes used. The success of distributed amplification in improving the high speed performance of nonregenerative amplifiers suggests its use in regenerative amplifiers. Below is a comparison of such a regenerative circuit with and without distributed amplification.

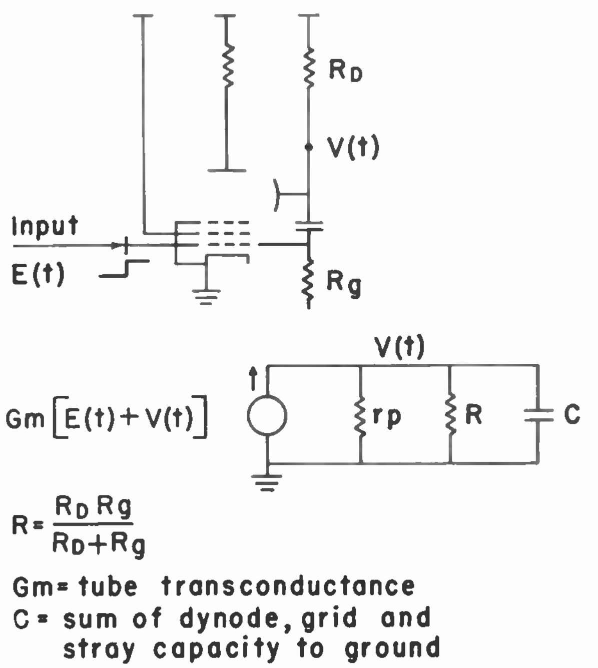


Fig. 1—One-tube multivibrator.

Fig. 1 shows an EFP 60 one-tube regenerative monostable multivibrator[1] and its equivalent circuit. For $r_p \gg R$ and a step input, the Laplace transform is[2]

$$V(s) = \frac{G_m E}{C} \frac{1}{S\left[S + \left(\frac{1}{RC} - \frac{G_m}{C}\right)\right]}. \quad (1)$$

For $(1/R) \ll G_m$ and $C = 2C_g$ where C_g is the grid input capacity, the inverse transform is

$$V(t) = E\left[e^{G_m t/2C_g} - 1\right]. \quad (2)$$

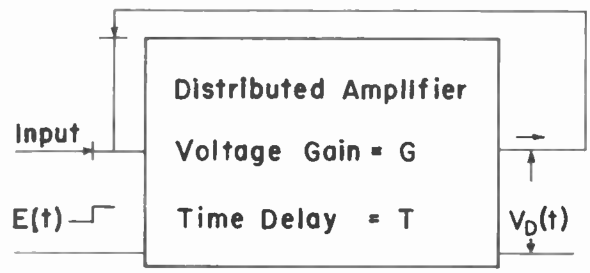


Fig. 2—Regenerative distributed amplifier.

Fig. 2 shows a regenerative distributed amplifier. It is assumed that the output voltage is of the same polarity as the input. For a step input the Laplace transform of this circuit is

$$V_D(S) = GE \frac{1}{S(e^{ST} - G)}. \quad (3)$$

The inverse transform is[3]

$$V_D(t) = GE \frac{1 - G^{n'}}{1 - G}$$

$$n'T < t < (n' + 1)T. \quad (4)$$

To obtain the smooth function which is equal to $V_D(t)$ when t/T is a whole number, and which is otherwise greater than $V_D(t)$, let $n' = t/T$. Then, for $G \gg 1$

$$V_D(t)(\text{smooth}) = E(G^{t/T} - 1). \quad (5)$$

Eqs. (4) and (5) are plotted in Fig. 3.

In a distributed amplifier[4]

$$g = \frac{nG_m Z_0}{2} \quad (6)$$

* Original manuscript received by the IRE, August 1, 1955; revised manuscript received November 4, 1955. Extract from a thesis submitted in partial satisfaction of requirements for the M.S. degree at University of California, Berkeley, Calif.

† NACA Ames Aeronautical Lab., Instrument Development, Moffett Field, Calif.

[1] N. F. Moody, *et al.*, "Millimicrosecond pulse techniques," *Electronic Eng.*, p. 218; May, 1952.

[2] S. Goldman, "Transformation Calculus and Electrical Transients," Prentice Hall, New York, 1949.

[3] J. Cossar and A. Erdelyi, "Dictionary of Laplace transforms," Admiralty Computing Service, Dept. of Scientific Research and Experiment, London, page VII-11.

[4] E. S. Ginzton, *et al.*, "Distributed amplification," Proc. IRE, vol. 36, pp. 956–969; August, 1948.

where

G_m = tube transconductance.

Z_0 = characteristic impedance of the plate line, assumed equal to the impedance of the grid line.

n = number of tubes per stage. A stage is defined as n tubes with all plates connected to one plate line and all grids connected to one grid line.

g = voltage gain per stage.

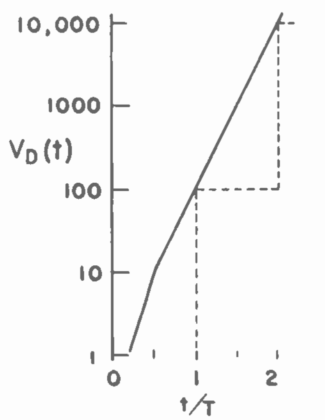


Fig. 3—Regenerative distributed amplifier output.

It is assumed that the plate line is terminated in Z_0 at both ends.

$$G = g^m \tag{7}$$

where

m = number of stages connected in cascade. A connection in cascade is defined as connection of the plate line of a stage to the grid line of following stage.

G = voltage gain of an amplifier with m stages connected in cascade.

$$T = mnZ_0C_g \tag{8}$$

where

T = time delay of the amplifier.

C_g = the capacity of each section of the grid line, and the plate line, assumed equal to the grid input capacity. A section is defined as one tube with its piece of the grid line and of the plate line.

From (6), (7) and (8) the gain/delay ratio is

$$\frac{G}{T} = \frac{\left(\frac{nG_mZ_0}{2}\right)^m}{mnZ_0C_g}. \tag{9}$$

For a single stage $m = 1$, so

$$g/\tau = G_m/2C_g \tag{10}$$

where τ = delay per stage.

This may be compared with the gain/rise-time ratio $= G_m/2.2\ C_t$ where C_t = sum of grid and plate capacity, and with Wheeler's bandwidth over which unity gain may be obtained $= G_m/\pi\sqrt{C_gC_p}$ where C_g = grid capacity and C_p = plate capacity.

Eq. (10) shows that for a given gain g, τ is independent of the number of tubes, so the cut-off frequency $f_c = 1/\pi Z_0C_g$ may be increased by reducing Z_0; this will reduce the gain, which may be increased to the original value by increasing the number of tubes, with no net change in τ or the G/T ratio. So the rise time $= 0.35/f_c$ can be made short enough to be neglected.

Eqs. (6), (7) and (8) show that for a given value of G_m, of Z_0, and of C_g the gain is proportional to n^m and the delay is proportional to mn, the total number of tubes. Ginzton[4] has shown that to obtain a given overall gain G with the least number of tubes mn, the gain per stage g should equal $e = 2.72$. This value of g will also yield the best gain/delay ratio, since delay is proportional to the number of tubes.

From (10), for m stages, with $g = e$,

$$T = \frac{2meC_g}{G_m}. \tag{11}$$

From (5), (7) and (11)

$$V_D(t)(\text{smooth}) = E(e^{G_mt/2eC_g} - 1). \tag{12}$$

Comparing (12) with (2) it is seen that since the exponent of e is smaller by a factor of $e = 2.72$ for the distributed amplifier the output of the nondistributed amplifier will rise faster.

Conclusion

If the output ends of the distributed amplifier plate lines are open circuited (6) becomes

$$g = nG_mZ_0 \tag{13}$$

and (12) becomes

$$V_D(t)(\text{smooth}) = E(e^{G_mt/eC_g} - 1). \tag{14}$$

Comparing (14) with (2) it is seen that since the exponent of e is smaller by a factor of $e/2 = 1.36$ for the distributed amplifier, the output of the nondistributed amplifier will rise faster.

Distributed amplification principle is very tempting but its inherent time delay restricts its application in regenerative circuits. The reference below[5] describes attempts to construct a regenerative amplifier using distributed amplification with no delay in feedback loop.

Acknowledgment

The author wishes to express his gratitude to Professors R. S. Mackay and J. R. Woodyard of the University of California, Berkeley, for their help and encouragement.

[5] B. S. Golosman, "A study of regenerative feedback around distributed amplifiers using secondary emission tubes," M.S. Thesis, University of California, Berkeley, Calif., January, 1955.

Keep-Alive Instabilities in a TR Switch*

T. J. BRIDGES†, P. O. HAWKINS†, AND D. WALSH†

Summary—**The reliability of a pulse radar system is often reduced by intermittent burnout of the receiver crystal. It has been found that one possible cause of this is the occurrence in the Transmit-Receive (TR) switch of occasional transitions from a glow to an arc of the keep-alive gas discharge. Following a transition, the keep-alive discharge is extinguished for a period of the order of fifty microseconds, during which time the protection afforded by the switch is greatly reduced. Methods of preventing this effect have been applied to TR cells. In the best method, two keep-alive electrodes are fitted to the same cell. This considerably reduces the chances of the cell being without a keep-alive discharge at any time. Cells of a particular design originally protected crystals for less than 100 hours, and this was attributed to the occurrence of glow-arc transitions. After modification as above, crystal protection for more than 2,000 hours was obtained.**

Introduction

One function of a TR cell in a pulse radar set is the protection of the associated crystal receiver from electrical burn-out. The incidence of the high rf power from the transmitter during the pulse produces an rf discharge in the cell. Most of the power is reflected, but some is transmitted to the crystal. This leakage power, as it is termed, consists of two parts, the "spike" and the "flat."

The spike occurs at the beginning of the pulse, while the rf discharge is still being formed. Although of high peak power its duration is very short (order of 10^{-8} seconds). Because of this its effect on the crystal is characterized by its energy content, rather than its power level.

The spike can be very damaging to the crystal, and its value is reduced to a minimum by providing a small dc keep-alive (or primer) discharge in the cell. With this discharge operating the spike energy is usually of the order of 0.1 ergs per pulse. If the discharge is switched off, the spike energy rises by a factor of a few times in broadband cells, to several hundred times in high-Q tunable cells.

The flat is a steady power transmitted during the rest of the pulse, usually less than 100 milliwatts in value and generally has no ill effect on the crystal.

In a correctly designed TR cell, with the keep-alive discharge operating, the measured leakage is lower than that known to cause crystal damage. In spite of this, crystal burnout sometimes occurs which is uncorrelated with any obvious fault (such as a keep-alive short circuit, or a rise in spike energy with life). This seriously reduces the reliability of the system.

A severe problem of this form was encountered during the development of a TR cell for a wavelength of 8 mm. Experiments were started to find the cause, and any possible cures of the trouble.

Description of Experiments

The TR cell was of the high-Q tunable type. Prototype cells had a single keep-alive electrode made of Nilo K alloy.[1] The keep-alive electrode was glass-sheathed up to its tip, where a small bare metal area about 0.005 inch in diameter was left, forming the cathode for the glow discharge. Limiting the cathode area in this way ensured that the 50 microampere discharge was in the abnormal cathode fall condition at the running current. This gave a positive slope to the current-voltage characteristic, thus preventing relaxation oscillations,[2] and also made sure that the dc discharge directed electrons towards the rf discharge gap. The gas filling was a mixture of 25 mm Hg of argon, the optimum value for rf breakdown by 8 mm waves, and 5 mm Hg of water vapor which ensured a short deionization time.

With the keep-alive discharge on, the spike leakage measured with a thermistor bridge was about half that required to burn out the crystal. However, upon running under approximately operational conditions, the cells protected crystals for a short number of hours and then allowed them to burn out. Once having caused a crystal burnout the cell would continue to do so whenever a new crystal was inserted. In a batch of five cells tested in this way crystal protection lasted from 20 to 106 hours (average 74 hours).

Careful measurements of spike, flat, insertion loss, and recovery time showed that these characteristics remained unchanged. The only effect noticed was a steady rise in the voltage required to maintain the keep-alive discharge. (This did not affect the discharge current as suitable manual adjustment was made.) A typical example of this behavior is shown in Fig. 1.

Explanation of the Effect

The only obvious change in the cell was associated with the keep-alive discharge and so attention was drawn to this as a possible source of the burnout mechanism. Previous work had shown that the rise in voltage was caused by the formation of an oxide film on the keep-alive electrode (cathode) surface during life. This oxidation was associated with the presence of water

* Original manuscript received by the IRE, April 28, 1955; revised manuscript received October 20, 1955.
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[1] Nickel-Cobalt-Iron glass sealing alloy (Henry Wiggin & Co., Ltd.).

[2] L. D. Smullin and C. G. Montgomery, "Microwave Duplexers," M.I.T. Rad. Lab. Ser., vol. 14, McGraw-Hill Book Co., Inc., New York, N. Y., p. 201; 1948.

vapor in the cell. Other writers[3] have shown that the presence of an oxide film on the cathode in a glow discharge greatly facilitates the change from a glow to an arc[4] discharge. Momentary glow to arc transitions have been shown to occur spontaneously in TR cell keep-alive discharges and have been suggested as the cause of abnormal keep-alive sputtering. Transitions were seen to occur in bursts of 10–20, at intervals of 5–60 minutes between bursts.[5]

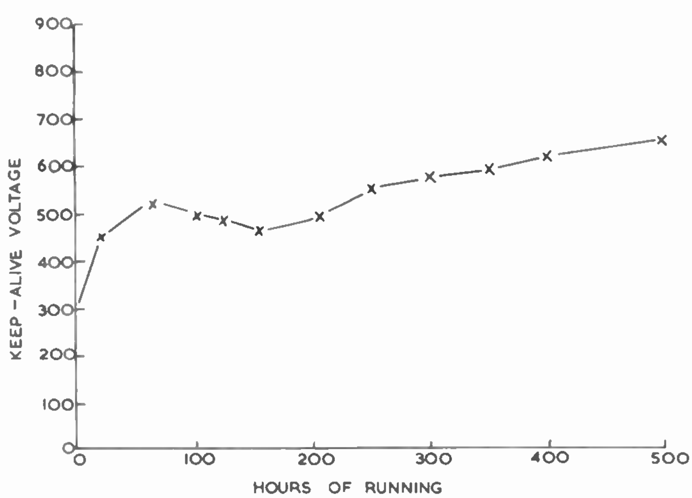

Fig. 1—Rise of keep-alive voltage during life.

In the present paper it is shown that the occurrence of a glow-arc transition at the keep-alive electrode can extinguish the discharge for a period of the order of 50 microseconds. If a magnetron pulse occurs during this time, an abnormally large spike results, which can cause burn-out of the crystal. In a typical example (see Appendix B) one abnormal spike occurs on the average every seventeen minutes. These occasional large spikes cannot be detected with ordinary leakage measuring gear, which measures mean power.

Effect of Glow-Arc Transition (See Fig. 2)

During the glow regime the potential difference across the keep-alive discharge is several hundred volts, and the discharge current is a few hundred microamperes. When the glow spontaneously changes to an arc, the potential difference falls to a few tens of volts and the current rises to the order of amperes. This current cannot flow through the high dropping resistor which feeds the glow discharge, but is supplied from the stray capacitance of the electrode. The value of this is only a few $\mu\mu$f, and so the arc discharge lasts a very short time—of the order of 10^{-10} seconds.

After this the arc stops. There will be no further discharge until the stray capacitance of the electrode has charged up through the dropping resistor to the striking potential of the glow discharge, when the glow will resume. The time during which the keep-alive discharge is extinguished is calculated in Appendix A. It is there shown that under normal operating conditions it is close to

$$\frac{V_s C}{I} \text{ seconds}$$

where V_s is the striking potential of the glow discharge (volts), C is the stray capacitance of the electrode (farads), and I is the normal discharge current (amperes). Taking typical values:

$$V_s = 500 \text{ volts}$$
$$C = 10 \text{ micro-microfarads}$$
$$I = 100 \text{ microamperes.}$$

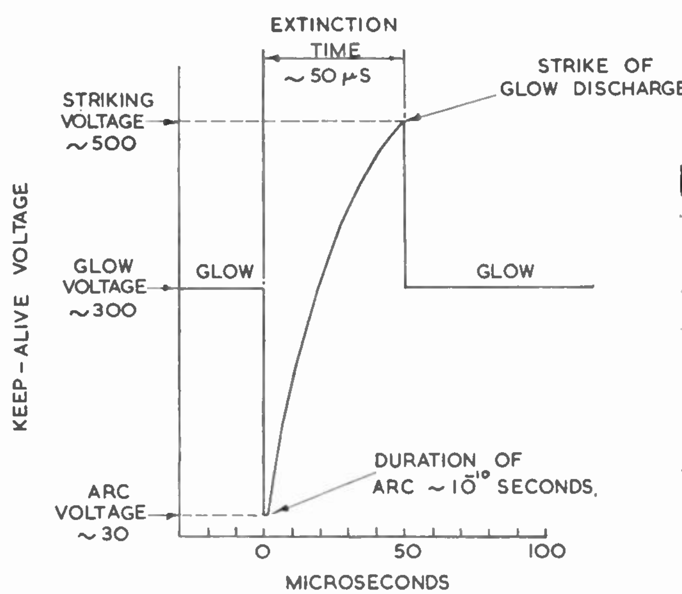

Fig. 2—Keep-alive voltage following a glow-arc transition.

The extinction time is close to 50 microseconds. Discontinuities in voltage consistent with this picture have been observed by connecting the keep-alive electrode to a cathode ray oscilloscope. It appeared likely then that the crystal damage caused by the prototype cells was due to occasional extinction of the keep-alive glow discharge, caused by glow-arc transitions. The fact that burn-out occurred only after an initial lapse of 20–100 hours would correspond to the time needed for a layer of

[3] T. Jurriaanse and M. J. Druzvesteyn, "Transition from glow discharge to arc discharge," *Physica*, vol. 3, pp. 825–840; August, 1936.
F. E. Haworth, "Experiments on the initiation of electric arcs," *Phys. Rev.*, vol. 80, pp. 223–226; October, 1950.
[4] J. D. Cobine, "Gaseous Conductors," McGraw-Hill Book Co., Inc., New York, N. Y., Ch. VIII, IX; 1941.
[5] J. C. French, "Electrode deterioration in transmit-receive tubes," *Jour. Res. Nat. Bur. Standards*, vol. 45, pp. 310–315; October, 1950.

oxide to form on an initially clean cathode surface. The high initial rate of rise of keep-alive discharge voltage, in roughly the same time, also corresponds with the formation of the oxide (Fig. 1).

Eliminating the Effect

The following two ways[6] of preventing transitions, or of avoiding their undesirable effects appeared promising:

1) The use of a material for the keep-alive electrode, which would not oxidize in the discharge and hence would not cause glow-arc transitions.
2) The fitting of two equivalent keep-alive electrodes to the tube, each of which on its own would normally be sufficient. The probability of there being no keep-alive discharge in the cell at the same time as a magnetron pulse is thereby greatly reduced. In a typical example (see Appendix B) the average time between abnormal spikes is increased from seventeen minutes with one keep-alive electrode, to 330,000 hours with two, an improvement of a million times.

Cells of both these types were constructed and greatly improved crystal lives were at once obtained. Rhodium keep-alive cells protected crystals for periods between 300–700 hours. They did, however, burn out crystals after this time. This was associated with another form of instability that developed. Unlike the glow-arc transition this took the form of an intermittent *increase* of keep-alive voltage. The explanation of this is unknown.

Another feature of the rhodium keep-alive cells was a relaxation oscillation of the discharge that invariably started after a short period of running. The frequency was about $\frac{1}{2}$ mc. Although the electrode area was restricted as explained above, sputtering of metallic rhodium soon increased the effective area and enabled oscillation to start. The oscillations had no effect on the protection afforded by the cell. This is because even at the minimum amplitude the current (approximately 10 microamps) prevented the spike leakage increasing to a dangerous level.

Cells were also made with two keep-alives. It was found that the leakage power of the cells was the same with either or both the keep-alives running. Lives without crystal burn-out of over 2,000 hours have been obtained with some of these cells. Of 27 cells life tested with an aggregate life of over 23,000 hours only one crystal burnout was encountered which could not be explained by some obvious fault.

If one keep-alive became inoperative crystal burnout invariably occurred although this caused no increase in the measured leakage power.

[6] T. J. Bridges, P. O. Hawkins, and D. Walsh, British Provisional Patent Application 28983/51.

Discussion

It has been shown that the occurrence of glow-arc transitions in the keep-alive discharge of a TR cell should have a large effect on the crystal protection afforded by the cell. Experiments were performed on cells which caused crystal burnout which was apparently due to this effect. By using keep-alive electrodes made of rhodium or preferably by using two keep-alive electrodes, the effect could be eliminated and reliable crystal protection obtained during life.

The cell was of high-Q tunable variety. This type is characterized by a large ratio of abnormal spike (with keep-alive off) to normal spike (with keep-alive on), in this case 100–200 times. This means that abnormal spikes from these cells could be very damaging to crystals and possibly only one was required to cause complete burnout. Transitions were thus made very obvious.

In other TR cells, particularly the broadband type, the ratio of abnormal to normal spikes may be a few times. Because of the statistical spread of burnout properties of crystals many TR cell-crystal combinations may be unaffected by the occurrence of a transition in this case. A crystal with a low burnout value (still within the specification) could still be damaged. Furthermore, although one spike two or three times greater than normal might not damage a crystal, the effect could be cumulative and burnout would then occur after a number of abnormal spikes. In broadband cells, therefore, the effect of transitions may not be so obvious, but could explain intermittent lack of crystal protection which does occur.

Appendix A

Calculation of Extinction Time

In the equivalent circuit of the keep-alive the factors which are properties of the cell design are:

I = discharge current (amperes)

C = stray capacity of electrode (farads)

V_g = glow potential of discharge (volts)

V_s = striking potential of discharge (volts)

The variable is V_b battery or supply voltage and R is the dropper resistance (ohms) which is set to give current I at battery voltage V_b.

$$R = \frac{V_b - V_g}{I}. \tag{1}$$

If the arc voltage is neglected, then after a transition the stray capacity will charge up according to

$$V = V_b\left\{1 - \exp\left(-\frac{t}{RC}\right)\right\}.$$

V = voltage across electrodes,

t = time after transition.

When V reaches V_s the discharge strikes

$$V_s = V_b\left\{1 - \exp\left(-\frac{t_d}{RC}\right)\right\};$$

t_d = time during which discharge is extinguished.

$$\therefore \frac{t_d}{CR} = \log_e\left\{1 - \frac{V_s}{V_b}\right\}.$$

Substitute (1)

$$-\frac{I}{V_sC}t_d = \left\{\frac{V_b}{V_s} - \frac{V_g}{V_s}\right\}\log_e\left\{1 - \frac{V_s}{V_b}\right\}.$$

As V_b becomes large, $\log_e(1 - V_s/V_b) \rightarrow - V_s/V_b$ and as V_g/V_s is of the order of 1.

$$t_d \rightarrow \frac{V_sC}{I} = t_0. \quad (2)$$

This is the extinction time for large battery voltage. For lesser voltages t_d is shown in terms of t_0 in Fig. 3. t_d is reduced for lower voltages. Since, however, for certainty of striking V_b/V_s should be at least 1.5, very little benefit can be obtained.

Taking typical values:

V_s = 500 volts

I = 100 microamperes

C = 10 micro-microfarads

$\therefore t_0$ = 50 microseconds.

Appendix B

Chances of Coincidence of Radar Pulse and no Keep-Alive Discharge

Extinction time of keep-alive discharge following glow-arc transition—t_d seconds

Repetition rate of radar—p pulses per second

Average rate of glow-arc transitions (assumed the same at every keep-alive)—r transitions per second

The radar pulse is assumed to be short compared with t_d.

The chance of one keep-alive discharge being extinguished at any given instant in time is rt_d. There are p instants every second when this can result in an abnormal spike. The average number of abnormal spikes per second is therefore rt_dp.

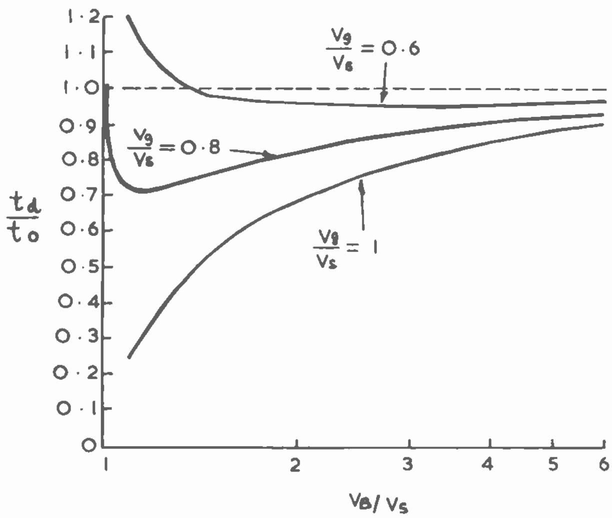

Fig. 3—Extinction time of keep-alive discharge as function of battery voltage.

With two keep-alive discharges in the cell, the chance of both being extinguished at any given instant in time is $(rt_d)^2$, and it follows that the average number of abnormal spikes per second is $(rt_d)^2p$.

To take an example:

t_d = 50 microseconds

p = 1,000 pulses per second

r = 1/50 transitions per second.

With one keep-alive discharge, the average number of abnormal spikes per second is $rt_dp = 10^{-3}$ per second. The average time between abnormal spikes is therefore about 17 minutes.

With two keep-alive discharges, the average number of abnormal spikes per second is $(rt_d)^2p = 10^{-9}$ per second. The average time between abnormal spikes is 330,000 hours.

The provision of two keep-alives has increased the time between abnormal spikes by a million times.

Acknowledgment

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The Optimum Tapered Transmission Line Matching Section*

R. E. COLLIN†, MEMBER IRE

Summary—**The tapered transmission line matching section is analyzed by treating it as a high pass filter. The ideal reflection coefficient characteristic for the taper that gives the smallest pass band tolerance for a given cutoff frequency and vice versa is derived from the expression for the reflection coefficient of the optimum designed n section quarter-wave transformer by taking the limit as n tends to infinity. By neglecting the square of the reflection coefficient in the differential equation for the reflection coefficient on the taper, a synthesis procedure is derived for obtaining an optimum taper. The procedure is similar to that used in designing an optimum line source distribution in antenna theory. Conditions are derived for the maximum allowable pass band reflection and change in impedance level for which the theory remains accurate. For a maximum reflection coefficient of 0.1 in the pass band the theory remains accurate for all frequencies above the cutoff value provided the change in impedance level does not exceed 7.5. This optimum taper is compared with the well-known exponential and Gaussian tapers and is found to be 13.9 per cent and 27 per cent shorter respectively for the same cutoff frequency and pass band tolerance.**

Introduction

VARIOUS aspects of the problem of the tapered or non-uniform transmission line have been treated by several authors.[1-3] A rigorous solution is possible only when the characteristic impedance varies in a specified manner. When the series impedance and shunt admittance both vary as a power of y (the distance along the taper), the voltage and current on the tapered line may be expressed in terms of Bessel functions. The current and voltage on the tapered line satisfy second order differential equations. In contrast to this Pierce has shown that[4] the input impedance looking into the tapered section at a point y satisfies a general first order Riccati equation. Likewise Walker and Wax have derived[5] a first order Riccati equation for the voltage reflection coefficient on the tapered line. These authors[5] have also given an iteration method of arriving at approximate solutions to the general problem. Schelkunoff also outlined[6] a somewhat similar procedure. However, the problem of what constitutes an optimum taper and the synthesis of a taper having a specified optimum behavior do not appear to have been analyzed. It is the purpose of this paper to present such an analysis.

Specifically the problem to be treated is as follows. Given two uniform lossless transmission lines of normalized characteristic impedance, unity and Z_2, what type of intermediate tapered section must be used to ensure the two lines will be as well matched as possible over as broad a band of frequencies as possible for the shortest physical length of taper? The analogous problem of the optimum multisection quarter-wave transformer has been treated[7] by the author in a previous paper. In this paper the author points out the similarity between the designs of a multisection quarter-wave transformer and of a broad side antenna having as narrow beam width as possible with the smallest possible side lobe level. Similarly, as will appear later, the design of an optimum taper is very closely analogous to the problem of the design of an optimum line source distribution in antenna theory. This latter problem has been treated[8] in considerable length by Taylor. It will be shown that the theory developed by Taylor may be applied with very little modification to the problem being studied. For this reason it was decided to use Taylor's notation as far as possible in order to avoid confusion.

The basic theory will be developed first, followed by an outline of a suitable design procedure. A comparison of the optimum designed taper and the exponential and Gaussian tapers will also be given. The theoretical development utilizes the infinite product expansions of the sin x and cos x functions. For this reason a short Appendix on the technique of expanding an integral function into an infinite product is included.

The Ideal Reflection Coefficient Characteristic

For an optimum designed n section quarter-wave transformer used to match a transmission line of normalized characteristic impedance unity to a line of nor-

* Original manuscript received by the IRE, October 20, 1955; revised manuscript received December 20, 1955.

† Canadian Armament Research and Development Establishment, Valcartier, Quebec, Canada.

[1] J. W. Arnold and P. F. Bechberger, "Sinusoidal currents in linearly tapered transmission lines," Proc. IRE, vol. 19, pp. 304–310; February, 1931.

[2] A. T. Starr, "The non-uniform transmission line," Proc. IRE, vol. 20, pp. 1052–1063; June, 1932.

[3] A. W. Gent and P. J. Wallis, "Impedance matching by tapered transmission lines," Jour. IEE, pt. IIIA, vol. 93, pp. 559–563; 1946.

[4] J. R. Pierce, "A note on the transmission line equations in terms of impedance," *B.S.T.J.*, vol. 22, pp. 263–265; July, 1943.

[5] L. R. Walker and N. Wax, "Non-uniform transmission lines and reflection coefficients," *J. App. Phys.*, vol. 17, pp. 1043–1045; December, 1946.

[6] S. A. Schelkunoff, "Remarks concerning wave propagation in stratified media," *Comm. on Pure and App. Math.*, vol. 4, pp. 117–128; June, 1951.

[7] R. E. Collin, "The theory and design of wide-band multi-section quarter-wave transformers," Proc. IRE, vol. 43, pp. 179–185; February, 1955.

[8] T. T. Taylor, "Design of line-source antenna for narrow beam width and low side lobes," Trans IRE, PGAP, VAP-3, pp. 16–28; January, 1955.

malized characteristic impedance Z_2 the modulus of the reflection coefficient ρ is given by:[9]

$$|\rho| = \frac{k\left|T_n\left\{\dfrac{\cos\beta\dfrac{L}{n}}{\cos\beta_0\dfrac{L}{n}}\right\}\right|}{\left[1+k^2T_n^{\,2}\left\{\dfrac{\cos\beta\dfrac{L}{n}}{\cos\beta_0\dfrac{L}{n}}\right\}\right]^{1/2}} \quad (1)$$

where

T_n = Tchebycheff polynomial of order n

β = phase constant (the same for all sections)

β_0 = lower cutoff value $=2\pi/\lambda_0$ where λ_0 is the lower cutoff wavelength

L = total length of n sections

$$k=\frac{Z_2-1}{2\sqrt{Z_2}}T_n^{-1}\left\{\frac{1}{\cos\beta_0\dfrac{L}{n}}\right\}=\text{pass band tolerance.}$$

The argument of the Tchebycheff polynomial

$$T_n \text{ is } \frac{\cos\beta\dfrac{L}{n}}{\cos\beta_0\dfrac{L}{n}}$$

and $T_n^{-1}(u)$ is written for $1/T_n(u)$.

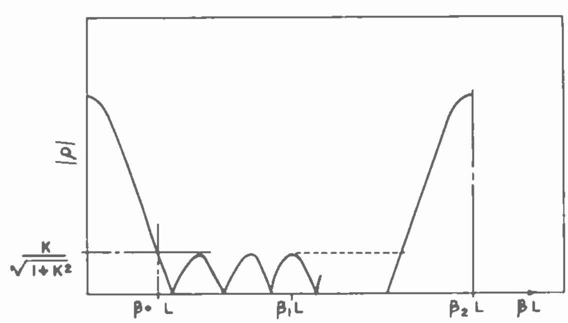

Fig. 1—Typical plot of modulus of reflection coefficient as a function of βL for an n section Tchebycheff transformer.

The Tchebycheff polynomial enters into the design of the optimum n section transformer because it has the property that it oscillates between ± 1 as the argument increases from -1 to $+1$. For values of the argument outside the range -1 to $+1$ the Tchebycheff polynomial increases monotonically. The variable $\cos\beta(L/n)$ is therefore normalized by dividing by $\cos\beta_0(L/n)$ so as to confine the equal amplitude oscillations within the desired pass band. A typical plot of this relationship as a function of βL is given in Fig. 1. The pass band is the region between the lower cutoff value of $\beta_0 L$ to the corresponding point at upper end of curve. This optimum designed transformer has the property that, for a given pass band tolerance, pass band is of maximum width and hence $\beta_0 L$ is a minimum. Conversely, for a given pass band width, the pass band tolerance is a minimum. Reflection coefficient is a periodic function of βL of period $\beta_2 L$. N simple zeros are in pass band.

If the total length of the transformer is kept constant at a value L, then as n increases each section becomes shorter and in the limit as n approaches infinity, the multisection transformer becomes a section of tapered transmission line of length L. Furthermore, the frequency at which the center of the pass band occurs; *i.e.*, $\beta_1 L$, moves off to infinity. Thus in contrast to the quarter-wave transformer which has an infinite number of pass bands separated by attenuation bands, the tapered transmission line is a high pass filter. This is the viewpoint adopted in this article. The lower cutoff frequency is obtained from the value of $\beta_0 L$. In general the reflection coefficient from a tapered transmission line matching section will exhibit one large or main lobe centered about $\beta L=0$, followed by an infinite number of minor lobes of much smaller magnitude above the cutoff frequency. The tapered transmission line may, of course, be designed for use at a frequency coincident with a zero of ρ. However since it is usually desirable to have a good match over a broad band of frequencies, it is necessary to design the taper so that the mismatch represented by the minor lobe maxima is tolerable and consequently a high pass filter is obtained. The Tchebycheff transformer design optimizes the relationship between minimum pass band tolerance and minimum cutoff frequency and therefore it is of considerable interest to determine the limiting form of (1) as n approaches infinity.

Consider the limiting form attained by

$$T_n\left\{\frac{\cos\beta\dfrac{L}{n}}{\cos\beta_0\dfrac{L}{n}}\right\}$$

as n tends to infinity. Using the relation that $T_n(\cos\theta)=\cos n\theta$, it is readily shown that the zeroes of the above function are located at

$$\cos\beta\frac{L}{n}=\cos\beta_0\frac{L}{n}\cos\left(S-\frac{1}{2}\right)\frac{\pi}{n},\; S=1,2,\cdots n.$$

For $n \gg S$, the arguments of the cosine terms are very small since $\beta_0 L<\beta L\ll n$ and hence

$$1-\frac{1}{2}\left(\beta\frac{L}{n}\right)^2\approx\left[1-\frac{1}{2}\left(\beta_0\frac{L}{n}\right)^2\right]\cdot\left[1-\frac{1}{2}\left(S-\frac{1}{2}\right)^2\frac{\pi^2}{n^2}\right]$$

or

$$\beta L\approx\pm\sqrt{(\beta_0 L)^2+(S-\tfrac{1}{2})^2\pi^2}.$$

[9] R. E. Collin, *loc. cit.*

As n tends to infinity this relation gives the exact location of the zeroes of the Tchebycheff polynomial. Considered as a function of βL, T_n is an even function of βL. Such a function is determined to within a constant[10] by a knowledge of the location of its zeroes. Thus it follows that

$$\lim_{n\to\infty} T_n\left[\frac{\cos\beta\frac{L}{n}}{\cos\beta_0\frac{L}{n}}\right] = C\prod_{s=1}^{\infty}\left(1-\frac{\beta^2L^2}{\beta_0^2L^2+(S-\frac{1}{2})^2\pi^2}\right). \quad (2)$$

The constant C is readily evaluated from the limiting value of T_n when βL equals zero. Thus

$$C = \lim_{n\to\infty} T_n\left[\frac{1}{\cos\beta_0\frac{L}{n}}\right].$$

Since n is very large,

$$\cos\beta_0\frac{L}{n} \approx 1 - \frac{1}{2}\beta_0^2\frac{L^2}{n^2}$$

and

$$T_n\left[\frac{1}{\cos\beta_0\frac{L}{n}}\right] \approx T_n\left(1+\frac{1}{2}\beta_0^2\frac{L^2}{n^2}\right).$$

Using the result $T_n(\cosh\theta) = \cosh n\theta$ where

$$\cosh\theta = 1 + \frac{1}{2}\beta_0^2\frac{L}{n^2},$$

the following limiting value of T_n is obtained when the approximation $\theta = \beta_0 L/n$ is made:

$$\lim_{n\to\infty} T_n\left[\frac{1}{\cos\beta_0\frac{L}{n}}\right] = \cosh\beta_0 L = C. \quad (3)$$

Thus

$$\lim_{n\to\infty} T_n\left[\frac{\cos\beta\frac{L}{n}}{\cos\beta_0\frac{L}{n}}\right] = \cosh\beta_0 L\prod_{s=1}^{\infty}\left(1-\frac{\beta^2L^2}{\beta_0^2L^2+(S-\frac{1}{2})^2\pi^2}\right)$$

$$= \cos L\sqrt{\beta^2-\beta_0^2} \quad (4)$$

The derivation of this result is given in the Appendix. Thus the limiting form of the reflection coefficient becomes

$$|\rho| = \frac{k\left|\cos L\sqrt{\beta^2-\beta_0^2}\right|}{[1+k^2\cos^2 L\sqrt{\beta^2-\beta_0^2}]^{1/2}} \quad (5)$$

where

$$k = \frac{Z_2-1}{2\sqrt{Z_2}}\,\frac{1}{\cosh\beta_0 L}. \quad (6)$$

The modulus of the reflection coefficient has one major lobe centered at $\beta L = 0$ and an infinite number of minor lobes of equal amplitude and separated by simple zeroes. Ratio of major lobe maximum to minor lobe maxima is:

$$\eta = \frac{\frac{2\sqrt{Z_2}}{Z_2+1}\cosh\beta_0 L}{\left(1+\frac{(Z_2-1)^2}{4Z_2\cosh^2\beta_0 L}\right)^{1/2}}. \quad (7)$$

This is the ideal characteristic required for the reflection coefficient and will give the minimum cutoff frequency for a given pass band tolerance and vice versa.

When only the first order reflections are taken into account in the design of an optimum n section transformer it is found that the modulus of the reflection coefficient is given by:[11]

$$|\rho| = k\left|T_n\left[\frac{\cos\beta\frac{L}{n}}{\cos\beta_0\frac{L}{n}}\right]\right|. \quad (8)$$

The limiting form of this expression is:

$$|\rho| = k\left|\cos L\sqrt{\beta^2-\beta_0^2}\right|. \quad (9)$$

This result also follows from (5) when $k^2 \ll 1$. This expression gives a major to minor lobe ratio of

$$\eta = \cosh\beta_0 L. \quad (10)$$

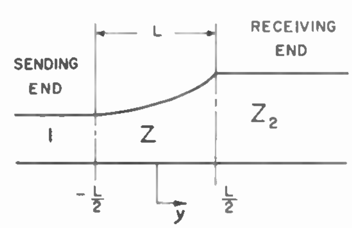

Fig. 2—Matching of two transmission lines by means of an intermediate tapered section.

Integral Equation for Reflection Coefficient

With reference to Fig. 2 let L be the length of the tapered section used to match a line of normalized impedance unity to a line of normalized impedance Z_2. The co-ordinate y measures the distance along the taper from the center point. It will be assumed that all the transmission lines are lossless and that the characteristic impedances are independent of frequency. Also the taper is assumed to be gradual enough so that the higher order modes excited are of negligible amplitude.

[10] E. C. Titchmarsh, "The Theory of Functions," Second edition, Oxford Univ. Press, New York, chap. VIII, 1939.

[11] R. E. Collin, *loc. cit.*, Appendix.

The conditions necessary for single mode propagation are not easy to establish. A qualitative estimate of the magnitude of the higher order modes may be obtained by replacing the taper by a large number of steps. The presence of higher order modes is accounted for by the shunt reactances which exist at each step. When these shunt reactances are negligible the effect of higher order modes may be neglected. From Reference 5 the reflection coefficient satisfies the differential equation:

$$\frac{d\rho}{dy} = 2j\beta\rho - \frac{1}{2}(1-\rho^2)\frac{d}{dy}\ln Z \qquad (11)$$

where the propagation constant β and the normalized characteristic impedance Z are in general functions of y. A first approximation to the integration of this equation is readily obtained if ρ^2 can be neglected as compared to unity. In practice this is usually valid and as will be shown, the error introduced is negligible for all values of βL above $\beta_0 L$ for most values of Z_2 usually encountered. The theory will first be developed for that class of transmission lines for which β does not vary along the taper. The modifications required in the theory when β varies with y will be considered later. When $y=L/2$, $Z(L/2)$ will be made equal to Z_2 and hence $\rho(L/2)$ will vanish when the output line is matched. With the above assumptions it is found that the reflection coefficient at the sending end is given by:

$$\rho = \frac{1}{2}e^{-j\beta L}\int_{-L/2}^{L/2}\frac{d\ln Z}{dy}e^{-2j\beta y}dy. \qquad (12)$$

This integral may be interpreted[12] as the Fourier transform of $d\ln Z/dy$. It may be put into the form studied by Taylor with the following change in variables:

$$p = 2\pi\frac{y}{L}, \qquad \frac{\beta L}{\pi} = \frac{2L}{\lambda} = u. \qquad (13)$$

The co-ordinate u is a measure of the taper length in half wavelengths. With these transformations the integral becomes

$$\rho = \frac{1}{2}e^{-j\beta L}\int_{-\pi}^{\pi}\frac{d\ln Z}{dp}e^{-ipu}dp. \qquad (14)$$

Let $d\ln Z/dp = g(p)$ and define $(F(u)$ by

$$F(u) = \int_{-\pi}^{\pi} g(p)e^{-ipu}dp. \qquad (15)$$

Thus

$$\rho = \tfrac{1}{2}e^{-j\beta L}F(u). \qquad (16)$$

When the function $g(p)$ satisfies the following conditions:

1) It is an even and continuous function of p with continuous derivatives for $|p| < \pi$;
2) It is uniformly bounded for all p within $-\pi < p < \pi$;
3) As p approaches $\pm\pi$, it is asymptotic to $K(p \pm \pi)^\alpha$, where $\alpha > -1$,

then the function $F(u)$ has, among others, the following properties:

(a) It is an even and entire function of u,
(b) As u approaches infinity it is asymptotic to

$$K_1 \frac{\cos\pi\left(u - \dfrac{1+\alpha}{2}\right)}{|u|^{1+\alpha}}$$

for $I_{mag.}\ u=0$, and

$$K_1 \frac{e^{\pi|u|}}{2|u|^{1+\alpha}} \quad \text{for} \quad \mathrm{Re}\ u = 0,$$

(c) The members of the nth zero pair tend to the positions $\pm[n+(\alpha/2)]$ as n approaches infinity.

The values of the constants K and K_1 depend on the particular choice of $g(p)$. Proofs of the above properties are given[13] in the paper by Taylor.

From the asymptotic form of $F(u)$ it is seen that the minor lobes decay as $|u|^{-(1+\alpha)}$. In order to have non-decaying minor lobes α must equal -1. However, since $d\ln Z/dp$ must be finite at $p=\pm\pi$, for a continuously smooth taper α must be greater/or equal to zero. In this case ideal reflection coefficient characteristic cannot be fully realized. Analysis following is based on the assumption that the taper is continuous at $p=\pm\pi$.

Specification of Reflection Coefficient Characteristic

The ideal reflection coefficient characteristic when ρ^2 is neglected in the integration of the differential (11) is proportional to $\cos\sqrt{\beta^2L^2-\beta_0^2L^2}$ since neglecting ρ^2 is equivalent to considering only the first order reflections in the theory of the n section transformer. This characteristic cannot be obtained for reasons already mentioned and hence the ideal characteristic will be approximated by the following function:

$$F(u) = C\cos\pi\sqrt{\frac{u^2}{\sigma^2} - \left(\frac{\beta_0 L}{\pi}\right)^2} \cdot \frac{\prod_{n=\bar{n}}^{\alpha}\left(1 - \dfrac{u^2}{n^2}\right)}{\prod_{n=\bar{n}}^{\infty}\left(1 - \dfrac{u^2}{\sigma^2\left[\left(\dfrac{\beta_0 L}{\pi}\right)^2 + \left(n - \dfrac{1}{2}\right)^2\right]}\right)} \qquad (17)$$

where σ is a constant slightly greater than unity and C is as yet an arbitrary constant. The choice of this particular form was governed by the following considerations. Since $d\ln Z/dp$ must be finite, the smallest value of

[12] F. Bolinder, "Fourier transforms in the theory of inhomogeneous transmission lines," Proc. IRE, vol. 38, p. 1354; November, 1950.

[13] T. T. Taylor, *loc. cit.*

α allowed is zero and hence the minor lobes must decay as $|u|^{-1}$ as u tends to infinity. Also the nth zero pair must tend to the position $\pm n$ for $\alpha=0$. Such a characteristic is given by the function $(\sin \pi u)/\pi u$. The infinite product term in the numerator is equal to this function with the first $\bar{n}-1$ zeroes cancelled out. In order to maintain uniform minor lobes for part of the range of u, the cosine term is introduced. For $u>\bar{n}$ the infinite product term in the denominator cancels the zeroes of the cosine function and allows the sine function to take control of the characteristic. For $|u|<|\bar{n}|$ the zeroes of $F(u)$ occur at

$$u = \pm \sigma\sqrt{\left(\frac{\beta_0 L}{\pi}\right)^2 + \left(n - \frac{1}{2}\right)^2}$$

while for $|u|>|\bar{n}|$ the zeroes occur at $u=\pm n$. For $|u|<|\bar{n}|$ the minor lobes are of very nearly equal amplitude while for $|u|>|\bar{n}|$ the minor lobes decay as $|u|^{-1}$. Thus the integer $\bar{n}$ divides the range of uniform minor lobes and decaying minor lobes. The zeroes of

$$\cos \pi\sqrt{u^2 - \frac{\beta_0^2 L^2}{\pi^2}}$$

are spaced closer than a unit distance apart for small values of u. As u becomes greater the spacing approaches unity and the zeroes tend to the positions $n-\frac{1}{2}$. In practice $\beta_0^2L^2$ would rarely exceed $3\pi^2$ and consequently for $u>5$ the location of the zeroes is within 0.1 of a unit from $n-\frac{1}{2}$, being located at u slightly greater than $n-\frac{1}{2}$. Thus for $\bar{n}>5$ the $(\bar{n}-1)$th zero of $F(u)$ is located approximately at $\bar{n}-3/2$ while the $\bar{n}$th zero is located at $\bar{n}$. This gives a spacing of 1.5 units between these two zeroes and consequently a higher minor lobe maximum in between. For this reason the constant σ is introduced so that the $\bar{n}$th zero of the cosine term will coincide with $\bar{n}$ and no where will the spacing between the zeroes be excessive. Therefore σ is given by

$$\sigma = \frac{\bar{n}\pi}{\sqrt{\beta_0^2 L^2 + (\bar{n} - \frac{1}{2})^2\pi^2}} \tag{18}$$

Considered as a function of $\bar{n}$, σ increases with $\bar{n}$ up to a point and then rapidly decreases asymptotically to unity. The value of $\bar{n}$ must be chosen large enough that any further increase in $\bar{n}$ will decrease σ. This condition is met if $(d\sigma/dn)<0$ or

$$\bar{n} > 2\frac{\beta_0^2 L^2}{\pi^2} + \frac{1}{2}, \tag{19}$$

If this condition is not met, then the first few zeroes of

$$\cos \pi\sqrt{\frac{u^2}{\sigma^2} - \frac{\beta_0^2 L^2}{\pi^2}}$$

for $u>\bar{n}$ occur for values of u slightly less than n. Since these zeroes are cancelled out by the infinite product term in the denominator and replaced by zeroes occurring at integral values of n, the net result is an increase in the spacing of some of the zeroes of $F(u)$ as compared with those of the cosine factor alone. This results in a major to minor lobe ratio smaller than $\cosh \beta_0 L$ and should be avoided. The condition (19) is readily satisfied in practice, usually for $\bar{n}$ not exceeding 5 or 6.

The function $F(u)$ differs very little from

$$C \cos \pi\sqrt{\frac{u^2}{\sigma^2} - \frac{\beta_0^2 L^2}{\pi^2}}$$

for $u<n$. The main difference is a slightly smaller minor lobe maximum and a small decay of the minor lobes. For this reason some of the properties of the reflection coefficient characteristic can be derived from the cosine term alone. The cutoff frequency is obtained from the value of u when the cosine term equals unity or $\pi u = \sigma\beta_0 L$ and hence

$$\beta_c L = \sigma\beta_0 L. \tag{20}$$

It is seen that the cutoff frequency is greater than that for the ideal characteristic by the factor σ, which, however, can be made as nearly equal to unity as desired by increasing $\bar{n}$. The ratio of major to minor lobe maxima is

$$\eta = \cosh \beta_0 L \tag{21}$$

which is identical to that for the ideal characteristic. As $\bar{n}$ is increased, the function $F(u)$ approaches the ideal characteristic asymptotically. In practice values of $\bar{n}$ around 5 or 6 give a characteristic that departs by only a few per cent from the ideal. Eqs. (20) and (21) show that the cutoff frequency and major to minor lobe ratio are related. When one is specified, the other is determined and vice versa. This property is also characteristic of the optimum n section transformer and emphasizes the similarity between the two.

The following equivalent form for $F(u)$ is readily derived by using the infinite product expansion of the sine and cosine functions.

$$F(u) = C \cosh \beta_0 L \frac{\sin \pi u}{\pi u} \cdot \frac{\prod_{n=1}^{\bar{n}-1}\left\{1 - \dfrac{u^2}{\sigma^2\left[\dfrac{\beta_0^2 L^2}{\pi^2} + \left(n - \dfrac{1}{2}\right)^2\right]}\right\}}{\prod_{n=1}^{\bar{n}-1}\left(1 - \dfrac{u^2}{n^2}\right)} \tag{22}$$

This form clearly shows the decaying property of the minor lobes as u becomes large. This form is also the most convenient for use in numerical computation.

Determination of Taper Impedance

In this section the required functional form of $d \ln Z/dy$ will be determined to give a reflection coefficient $\rho = \frac{1}{2}e^{-j\beta L}F(u)$ where $F(u)$ is the function defined by (17).

Let $g(p)$ be expanded into a Fourier series as follows:

$$g(p) = \begin{cases} \sum_{m=0}^{\infty} B_m \cos mp & |p| \le \pi, \\ 0, & |p| \ge \pi. \end{cases} \quad (23)$$

Substituting into the integral (15) and performing the integrations gives:

$$F(u) = \sum_{m=0}^{\infty} 2B_m \frac{u \cos m\pi \sin \pi u}{u^2 - m^2}. \quad (24)$$

To determine the coefficients B_m let u approach an integer s; thus

$$\lim_{u\to s} 2 \sum_{m=0}^{\infty} B_m \frac{u \cos m\pi \sin \pi u}{u^2 - m^2} = \begin{cases} \pi\epsilon_{0m}^{-1} B_m, & s = m, \\ 0, & s \ne m \end{cases}$$

where

$$\epsilon_{0m} = \begin{cases} \frac{1}{2} & \text{for } m = 0, \\ 1 & \text{for } m \ne 0. \end{cases}$$

Therefore the coefficients B_m are given by

$$B_m = \frac{\epsilon_{0m} F(m)}{\pi}. \quad (25)$$

Since $F(m)$ vanishes for all $m \ge \bar{n}$

$$g(p) = \sum_{0}^{\bar{n}-1} \frac{\epsilon_{0m} F(m)}{\pi} \cos mp$$

and

$$\frac{d \ln Z}{dy} = \frac{2}{L} \sum_{0}^{\bar{n}-1} \epsilon_{0m} F(m) \cos 2 \frac{\pi m}{L} y. \quad (26)$$

Integrating this equation gives the taper impedance as a function of y. The function $F(u)$ contains a constant C which is determined as follows. When $y = L/2$, $Z(L/2) = Z_2$ and hence

$$\int_{-L/2}^{L/2} \frac{d \ln Z}{dy} dy = \ln Z_2 = \frac{2}{L} \sum_{0}^{\bar{n}-1} \epsilon_{0m} F(m) \int_{-L/2}^{L/2} \cos 2 \frac{m\pi}{L} y dy$$

$$= F(0) = C \cosh \beta_0 L.$$

Thus

$$C = \frac{\ln Z_2}{\cosh \beta_0 L}. \quad (27)$$

The final form of the reflection coefficient is

$$\rho = \frac{1}{2} e^{-j\beta L} \frac{\ln Z_2 \cos L\sqrt{\frac{\beta^2}{\sigma^2} - \beta_0^2}}{\cosh \beta_0 L} \cdot \frac{\prod_{\bar{n}}^{\infty} \left(1 - \frac{\beta^2 L^2}{n^2\pi^2}\right)}{\prod_{\bar{n}}^{\infty} \left(1 - \frac{\beta^2 L^2}{\sigma^2[\beta_0^2 L^2 + (n - \frac{1}{2})^2\pi^2]}\right)}. \quad (28)$$

Design Procedure

The above analysis is summarized below in convenient form for use in designing a matching taper. Let the characteristic impedance of the output line, normalized with respect to the impedance of the input line, be Z_2. Let the maximum tolerable value of the reflection coefficient in the pass band be ρ_m. The major to minor lobe ratio η is given by $\cosh \beta_0 L$ where

$$\cosh \beta_0 L = \frac{\ln Z_2}{2\rho_m} \quad (29)$$

This relation determines the ideal cutoff value $\beta_0 L$. The practical cutoff value $\beta_c L$ is equal to $\sigma\beta_0 L$ where

$$\sigma = \frac{\bar{n}\pi}{\sqrt{\beta_0^2 L^2 + (\bar{n} - \frac{1}{2})^2\pi^2}} \quad (30)$$

and $\bar{n}$ must be chosen greater than

$$\frac{2\beta_0^2 L^2}{\pi^2} + \frac{1}{2}.$$

For a cutoff wavelength of λ_c the physical length of taper required is

$$L = \frac{\sigma\lambda_c}{2\pi} \operatorname{arc\,cosh} \frac{\ln Z_2}{2\rho_m}. \quad (31)$$

The variation in taper impedance required is

$$Z(y) = \exp\left(\sum_{0}^{\bar{n}-1} \frac{\epsilon_{0m}}{m\pi} F(m) \sin \frac{2m\pi}{L} y + \frac{F(0)}{2}\right) \quad (32)$$

where

$$F(m) = -\ln Z_2 \frac{\cos m\pi}{2} \cdot \frac{\prod_{n=1}^{\bar{n}-1} \left(1 - \frac{m^2\pi^2}{\sigma^2[\beta_0^2 L^2 + (n - \frac{1}{2})^2\pi^2]}\right)}{\prod_{n=1 \cdot n\ne m}^{\bar{n}-1} \left(1 - \frac{m^2}{n^2}\right)} \quad (33)$$

and $F(0) = \ln Z_2$. The reflection coefficient is given by (28). The design is somewhat conservative since the major to minor lobe ratio is slightly greater than η.

Comparison with Exponential and Gaussian Tapers

Exponential Taper

If $\bar{n}$ is put equal to unity in the above theory the exponential taper is obtained. The variation in taper impedance is given by

$$\ln Z(y) = \left(\frac{y}{L} + \frac{1}{2}\right) \ln Z_2. \quad (34)$$

The reflection coefficient is given by

$$\rho = \frac{1}{2} e^{-j\beta L} \ln Z_2 \frac{\sin \beta L}{\beta L}. \quad (35)$$

The first minor lobe maximum occurs when $\tan \beta L = \beta L$ or $\beta L = 1.43\pi$. The ratio of the major lobe to first minor lobe maximum is 4.6. The cutoff frequency occurs when $\sin \beta L = 0.217 \beta L$ or

$$\beta_c L = 0.815\pi.$$

For the optimum taper the corresponding cutoff value is $\beta_c L = \sigma \cosh^{-1} 4.6 = 0.702\pi\sigma$. By choosing $\bar{n}$ large enough, σ can be made as close to unity as desired and $\beta_c L$ approaches 0.702π. Thus the optimum taper is 13.9 per cent shorter than the exponential taper for the same pass band tolerance. However, a value of $\eta = 4.6$ usually results in too large a reflection in the pass band unless Z_2 is close to unity. Thus the exponential taper would normally be used only when a narrow band of frequencies in the vicinity of a zero of the reflection coefficient is to be passed.

For the exponential taper $d \ln Z/dy$ is a constant and the differential (11) can be solved rigorously by elementary means. The rigorously correct solution is:

$$\rho = \frac{C \sin \dfrac{RL}{2}}{R \cos \dfrac{RL}{2} + 2j\beta \sin \dfrac{LR}{2}} \tag{36}$$

where $C = \ln Z_2/L$ and $R = \sqrt{4\beta^2 - C^2}$. When $4\beta^2 \gg C^2$, $R \approx 2\beta$ and the reflection coefficient reduces to

$$\frac{1}{2} e^{-j\beta L} \ln Z_2 \frac{\sin \beta L}{\beta L}$$

which is the same result obtained by neglecting ρ^2 in the differential equation. Hence for the approximate theory to be valid it is necessary that $4\beta^2 L^2 \gg (\ln Z_2)^2$. For all values of βL satisfying this condition the analysis neglecting ρ^2 is sufficiently accurate for design purposes.

The exponential taper is a special case of the general theory and therefore the above criterion will also be used to determine the range of validity of the theory for the optimum taper. For the optimum taper the value of ρ at a minor lobe maximum is $\rho = \ln Z_2/2\eta$. The minimum value of βL of interest is $\beta_c L$. The value of $\beta_c L$ is given by $\beta_c L = \sigma \cosh^{-1} \eta$. Thus the condition $4\beta_c^2 L^2 \gg (\ln Z_2)^2$ may be written as:

$$\left(\cosh^{-1} \frac{\ln Z_2}{2\rho_m}\right)^2 \gg \left(\frac{\ln Z_2}{2\sigma}\right)^2 \tag{37}$$

where η has been replaced by $\ln Z_2/2\rho_m$. In practice a voltage standing wave ratio not exceeding 1.2 will be required and hence $\rho_m < 0.1$. For the left side of (37) to be at least ten times greater than the right, Z_2 cannot exceed 7.5. When ρ_m does not exceed 0.05, values of Z_2 up to 12 are allowed. For values of Z_2 greater than this, correspondingly smaller values of ρ_m are required in order for the theory to be accurate. For value of βL less than $\beta_c L$ the theory does not give an accurate value of the reflection coefficient unless Z_2 is close to unity. However this is of no practical consequence since the taper is not used for frequencies below the cutoff value. When βL is equal to zero, the above theory gives a value $\frac{1}{2} \ln Z_2$ for the reflection coefficient whereas the correct value is $(Z_2-1)/(Z_2+1)$. Expanding the logarithm function gives

$$\frac{1}{2} \ln Z_2 = \frac{Z_2 - 1}{Z_2 + 1} + \frac{1}{3}\left(\frac{Z_2 - 1}{Z_2 + 1}\right)^3 + \cdots$$

so it is seen that the expressions agree to within 4 per cent for $Z_2 < 2$.[14]

Gaussian Taper

The Gaussian taper is obtained when the number of sections in the binomial transformer is increased without limit while the total length of the transformer is kept constant. For this taper the characteristic impedance is given by:[15]

$$Z = \begin{cases} \sqrt{Z_2} \exp\left[\dfrac{2}{L} \ln Z_2 \left(y + \dfrac{y^2}{L}\right)\right], & -\dfrac{L}{2} \le y \le 0 \\ \sqrt{Z_2} \exp\left[\dfrac{2}{L} \ln Z_2 \left(y - \dfrac{y^2}{L}\right)\right], & 0 \le y \le \dfrac{L}{2} \end{cases} \tag{38}$$

The reflection coefficient is given by:

$$\rho = \frac{1}{2} e^{-j\beta L} \ln Z_2 \left[\frac{\sin \beta \dfrac{L}{2}}{\beta \dfrac{L}{2}}\right]^2. \tag{39}$$

The first minor lobe maximum occurs at $\beta L = 2.86\pi$. The ratio of major to first minor lobe maximum is 21.1. The cutoff frequency occurs when $\beta_c L = 1.63\pi$. The Gaussian taper has a cutoff frequency twice as high as the exponential taper but the first minor lobe maximum is only 4.75 per cent compared with 21.8 per cent. This behavior has been achieved by making all the zeroes of ρ double zeroes. The spacing between zeroes has consequently also been doubled. For a value of $\eta = 21.1$ the cutoff frequency for the optimum taper occurs when $\beta_c L = 1.19\pi\sigma$. For sufficiently large $\bar{n}$ the optimum taper is 27 per cent shorter for the same cutoff frequency and pass band tolerance. For the same cutoff frequency and taper length the value of η can be increased to a maximum of 84 for the optimum taper compared with 21.1 for the Gaussian taper. This illustrates the point that superior performance cannot be obtained with multiple zeroes in the pass band.

Curves of reflection coefficient against βL for the Gaussian taper and the optimum taper with $\eta = 21.1$ are plotted in Fig. 3. For the Gaussian and exponential tapers the choice of η is not arbitrary, being fixed at

[14] Conditions similar to those given here may be applied to determine the allowed range of Z_2 in order for the first order theory for the optimum n section quarter-wave transformer to be valid.
[15] Bolinder, F., *loc. cit.*

the values of 21.1 and 4.6 respectively. A curve of $\frac{1}{2}(|\rho|/\ln Z_2)$ vs βL for the exponential taper is readily constructed since it is equal to the simple function $|\sin \beta L|/\beta L$. The design procedure for the optimum taper is more flexible since it allows the arbitrary choice of η and then determines the shortest possible length of taper which will give this value of major to minor lobe ratio.

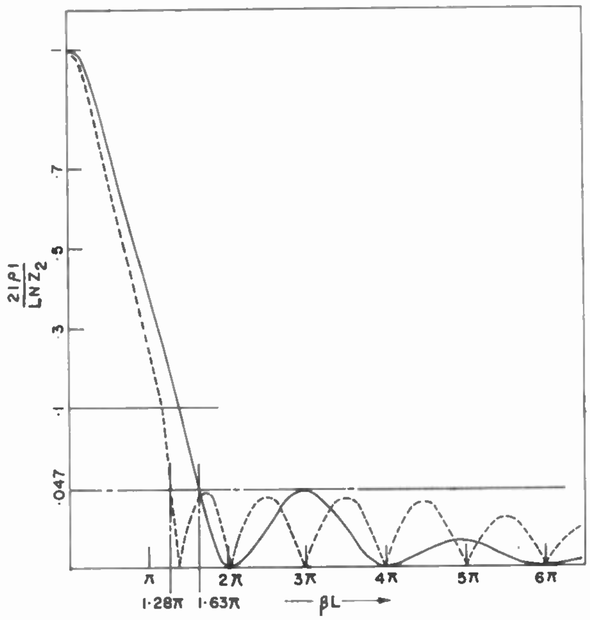

Fig. 3—Reflection coefficient as a function of βL for the Gaussian and optimum taper. ——— Gaussian taper. - - - - - - Optimum taper, $\eta=21.1$, $\bar{n}=5$, $\sigma=1.072$.

The Case When β Varies Along The Taper

As an example of conditions where the propagation constant varies along the taper, consider the problem of matching an infinite lossless dielectric sheet to free space by means of an inhomogeneous dielectric sheet of thickness L and of varying dielectric constant along the y axis as in Fig. 4. Let the relative dielectric constant of the dielectric sheet to be matched by K_2. A *TEM*

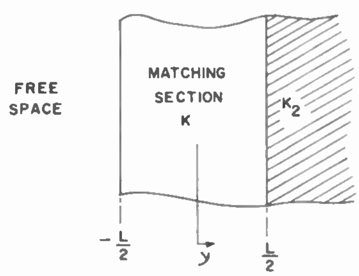

Fig. 4—Matching of a dielectric sheet to free space by means of an inhomogeneous intermediate section.

wave is assumed incident normally on the dielectric sheet. For this type of wave the propagation constant is given by $\sqrt{K}\beta$ where β is the free space propagation constant $\omega\sqrt{\mu\epsilon_0}$ and K is the relative dielectric constant of the taper and is a function of y. The normalized wave impedance of the dielectric may be taken as $K^{-1/2}$. Thus neglecting ρ^2 the differential equation for the reflection coefficient becomes:

$$\frac{d\rho}{dy} = 2j\beta\sqrt{K}\rho + \frac{1}{2}\frac{d \ln \sqrt{K}}{dy}. \tag{40}$$

When $y=L/2$, K will be made equal to K_2 so $\rho(L/2)$ vanishes. Introducing the new phase variable θ defined as

$$\theta = \int_{-L/2}^{y} \sqrt{K}dy - \frac{1}{2}\int_{-L/2}^{L/2} \sqrt{K}dy \tag{41}$$

it is readily found that ρ is given by the following integral:

$$\rho = -\frac{1}{2} e^{-j2\beta\theta_0} \int_{-\theta_0}^{\theta_0} \frac{d \ln \sqrt{K}}{d\theta} e^{-j2\beta\theta} d\theta \tag{42}$$

where

$$\theta_0 = \theta\left(\frac{L}{2}\right) = -\theta\left(-\frac{L}{2}\right) = \frac{1}{2}\int_{-L/2}^{L/2} \sqrt{K}dy. \tag{43}$$

The constant term has been introduced in the phase variable θ in order to obtain a symmetrical range in θ. The previous analysis may now be applied to this integral with θ replacing y and θ_0 replacing $L/2$. In particular it is found that the dielectric constant along the taper must satisfy the following integral equation for an optimum taper design:

$$K = \exp\left(-2\sum_{0}^{\bar{n}-1} \frac{\epsilon_{0m}F(m)}{m\pi} \sin\frac{m\pi\theta}{\theta_0} + \ln\sqrt{K_2}\right) \tag{44}$$

where θ is the function of K and y given by (41). In general the solution to this integral equation must be obtained numerically. For the particular case of the exponential taper this integral equation simplifies to:

$$\ln\sqrt{K} = \frac{\ln\sqrt{K_2}\int_{-L/2}^{y}\sqrt{K}dy}{\int_{-L/2}^{L/2}\sqrt{K}dy} \tag{45}$$

for which

$$\sqrt{K} = \left(\frac{1-\sqrt{K_2}}{L\sqrt{K_2}}y + \frac{1+\sqrt{K_2}}{2\sqrt{K_2}}\right)^{-1} \tag{46}$$

is a solution.

Conclusion

The theory for the optimum design of a line source distribution in antenna pattern synthesis has been applied to the problem of designing a tapered transmission line matching section that optimizes the relationship between cutoff frequency and minimum pass band tolerance. Such an optimum design leads to a considerably smaller pass band reflection as compared with that obtainable from an exponential or Guassian taper of the same physical length and having the same cutoff frequency. The theory is restricted to that class of transmission line for which the propagation constant can be expressed as the product of a frequency varying function and a space varying function. As such it is not generally applicable to waveguide tapers. An exception is the case of the rectangular waveguide supporting H_{0n} modes and which is tapered along the narrow guide dimension only. For this type of taper the guide wavelength does not vary along the taper and the characteristic impedance may be taken proportional to the waveguide height.

The directional coupler employing continuously distributed coupling may be treated in a similar manner.

Appendix I

Infinite Product Expansion of an Integral Function

Let $F(w)$ be an analytic integral function of the complex variable w. The logarithmic derivative

$$\frac{d \ln F}{dw} = \frac{F'}{F}$$

is a meromorphic function whose only singularities in the complex plane are poles located at the zeroes of $F(w)$. Let F have simple zeroes at $w=w_n$, $w_n \neq 0$; $n=1, 2, \cdots \infty$. Thus F'/F has simple poles at $w=w_n$. By integrating the partial fraction expansion of F'/F the following infinite product expansion of F is obtained,[16]

$$F(w) = F(0)e^{wF'(0)/F(0)} \prod_{n=1}^{\infty} \left(1 - \frac{w}{w_n}\right) e^{w/w_n}.$$

If F is an even function of w with simple zeroes at $w = \pm w_n$, $n=1, 2, \cdots \infty$, then $F'(0)=0$ and

$$\begin{aligned} F(w) &= F(0) \prod_{-\infty}^{\infty} \left(1 - \frac{w}{w_n}\right) e^{w/w_n} \\ &= F(0) \prod_{1}^{\infty} \left(1 - \frac{w}{w_n}\right) e^{w/w_n} \prod_{1}^{\infty} \left(1 + \frac{w}{w_n}\right) e^{-w/w_n} \\ &= F(0) \prod_{1}^{\infty} \left(1 - \frac{w^2}{w_n^2}\right). \end{aligned}$$

As an example consider the function $\cos L\sqrt{\beta^2-\beta_0^2}$ where βL is the variable. The zeroes of this function occur at $L\sqrt{\beta^2-\beta_0^2} = \pm(S-\frac{1}{2})\pi$. $S=1, 2, \cdots \infty$, or $\beta^2 L^2 = \beta_0^2 L^2 + (S-\frac{1}{2})^2\pi^2$. When $\beta L=0$ the value of the function is $\cosh \beta_0 L$. Hence using the general formula

$$\cos L\sqrt{\beta^2 - \beta_0^2} = \cosh \beta_0 L \prod_{S=1}^{\infty} \left(1 - \frac{\beta^2 L^2}{\beta_0^2 L^2 + (S - \frac{1}{2})^2\pi^2}\right)$$

which is the derivation of (4) in the text. The expansion of $\sin w$ is obtained from the expansion of the even function $\sin w/w$; *i.e.*,

$$\sin w = w \prod_{n=1}^{\infty} \left(1 - \frac{w^2}{n^2\pi^2}\right).$$

Appendix II

The analysis of Klopfenstein,[17] shows how the ideal reflection coefficient characteristic may be obtained by including a discontinuity at each end of the taper. The part of the reflection coefficient due to the two discontinuities is

$$e^{-j\beta L} \frac{\ln Z_2}{2 \cosh \beta_0 L} \cos \beta L.$$

As βL tends to infinity the reflection from the smooth part of the taper vanishes leaving only that due to the two discontinuities, the latter having constant amplitude minor lobes.

As $\bar{n}$ tends to infinity our results become identical to that of Klopfenstein. A by-product of our analysis is a simple Fourier series expansion of $\ln Z(y)$. When $\bar{n} \to \infty$, $\sigma \to 1$ and

$$F(u) \to \frac{\ln Z_2}{\cosh \beta_0 L} \cos \sqrt{u^2\pi^2 - \beta_0^2 L^2}.$$

Thus (32) becomes

$$\ln Z(y) = \frac{1}{2} \ln Z_2 + \frac{\ln Z_2}{\cosh \beta_0 L} \sum_{m=0}^{\infty} \frac{\epsilon_{0m}}{m\pi} \cos \sqrt{m^2\pi^2 - \beta_0^2 L^2} \times \sin \frac{2m\pi y}{L}.$$

Adding and subtracting a similar series gives

$$\ln Z(y) = \left(\frac{1}{2} + \frac{y}{L}\right) \ln Z_2 + \frac{\ln Z_2}{\cosh \beta_0 L} \sum_{1}^{\infty} \frac{\cos m\pi}{m\pi} \sin \frac{2m\pi y}{L} + \frac{\ln Z_2}{\cosh \beta_0 L} \sum_{1}^{\infty} \frac{1}{m\pi} (\cos \sqrt{m^2\pi^2 - \beta_0^2 L^2} - \cos m\pi) \cdot \sin \frac{2m\pi y}{L}.$$

The second series converges rapidly and the first series

[16] P. M. Morse, H. Feshbach, "Methods of Theoretical Physics," McGraw-Hill Book Company, New York, chap. 4, 1953.

[17] R. W. Klopfenstein, "A transmission line taper of improved design," Proc. IRE, vol. 44, pp. 31–35; January, 1956.

is the Fourier sine series expansion of the sawtooth function $-Y/L$. It is this series which brings in the discontinuous change in $Z(y)$ at $y=\pm L/2$. The value of $\ln Z(y)$ is readily computed from the following result:

$$\ln Z(y) = \left(\frac{1}{2}+\frac{y}{L}\right)\ln Z_2 - \frac{y}{L}\frac{\ln Z_2}{\cosh \beta_0 L} + \frac{\ln Z_2}{\cosh \beta_0 L}\sum_1^\infty \frac{1}{m\pi}(\cos\sqrt{m^2\pi^2-\beta_0{}^2L^2}-\cos m\pi) \cdot \sin\frac{2m\pi y}{L}, \qquad |y| \leqq \frac{L}{2}.$$

$$\ln Z(y) = \ln Z_2,\ y > \frac{L}{2}. \qquad \ln Z(y) = 0,\ y < \frac{-L}{2}.$$

The discontinuity in $\ln Z(y)$ at $y=\pm L/2$ is of magnitude

$$\frac{\ln Z_2}{2\cosh \beta_0 L}.$$

The above result agrees with that obtained by Klopfenstein. It appears that Eq. (12) in Klopfenstein's paper should have a term $-\rho_0/\cosh A$ added to the right-hand side.

A New Annular Waveguide Rotary Joint*

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Summary—**A new waveguide rotary joint has been designed which will permit multiple stacking of similar joints on a common axis. The rotary joint of annular waveguide design will carry high power and permit low vswr and low insertion loss operation throughout the full rotation. The theory, design, and characteristics of such a joint are presented.**

Introduction

ROTARY JOINTS of the multiple coaxial type cannot provide many independent concentric high-power transmission circuits since the large spacings required to prevent breakdown create higher mode problems in the outer spaces. The commonly-used waveguide rotary joint which utilizes two rectangular-TE_{10}-to-circular-TM_{01}-mode transitions is unadaptable to multiple use since the input and output lines would physically interfere.

These difficulties are avoided when an annular type of rotary joint is used. In this configuration each joint consists of a ring split concentrically so that the two sections can rotate about a common axis while transferring power from one to the other. One transmission line would be connected to one ring section, the other to the second ring section. For a stacked array, all the inner transmission lines would then pass through the rings. The annular type of rotary joint will permit multiple stacking of similar joints and complete removal of all electronic components of a radar system from the rotating platform.

Previous work on an annular rotary joint was reported by Coleman[1] in 1948. This joint had a "dead spot" although it was only 3 to 4 degrees wide. Another annular rotary joint was invented by Breetz[2] and it was characterized by low vswr and insertion loss at several extremely narrow evenly-spaced frequency bands.

Basic Considerations of Transvar Coupler

In the design of the new rotary joint, Transvar (TM) couplers[3] are used to obtain essentially uniform low vswr, low insertion loss characteristics (no "dead spots") as a function of joint rotation. Such desirable characteristics are possible at power levels approaching the maximum-power capacity of the waveguide.

For this application the basic Transvar coupler must pass the total input power through three different modified structures and coupling conditions. These will be called single coupler, cascaded couplers, and diagonal-arm condition.

Single Coupler

Referring to Fig. 3 in footnote reference 3, if the design of coupling apertures is such as to couple the total input power, it is possible to make the shutter much

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[1] G. M. Coleman, Unpublished notes on "Around-the-mast radar antenna," U. S. Navy Electronics Lab., San Diego, Calif.

[2] L. D. Breetz, Patent No. 2, 595, 186, issued April 29, 1952. Also L. D. Breetz, "A waveguide rotary joint with waveguide feed," Naval Res. Lab., NRL Rep. 3795; January 18, 1951.

[3] K. Tomiyasu and S. B. Cohn, "The transvar directional coupler," Proc. IRE, vol. 41, pp. 922–926; July, 1953.

longer than d, the total length of the coupling apertures, and still obtain total power coupling provided all apertures are exposed. Sliding the secondary waveguide merely changes the relative phase between E_2 and E_0.

Cascaded Couplers

A total transfer of input power must be obtained when two Transvar couplers are connected in cascade, separated any arbitrary distance, with one coupler of length ϵd and the other of length $(1-\epsilon)d$, where ϵ lies between zero and one (Fig. 1). Mathematically, this can be demonstrated neglecting reflections at discontinuities.

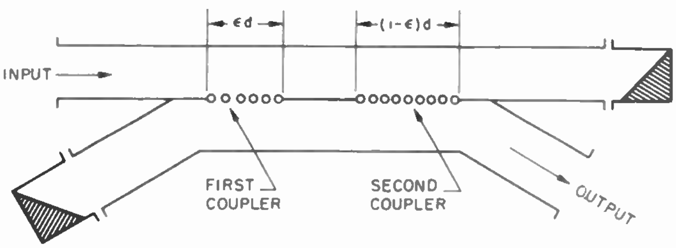

Fig. 1—Cascaded couplers.

Designating the electric fields in the primary guide as E_1 and in the coupled guide as E_2, the normalized electric vectors at the end of the first coupler are

$$(E_1)_1 = 1 + e^{j\epsilon\pi}$$
$$(E_2)_1 = 1 - e^{j\epsilon\pi}. \tag{1}$$

These relationships are based on the $\epsilon\pi$ radian phase difference that exists between the two propagating modes at the end of the first coupling region.

Between the two couplers is a section of uniform waveguide so that the relative phase and amplitude of $(E_1)_1$ and $(E_2)_1$ are preserved at the input of the second coupler. Since for the second coupler the phase difference between the two propagating modes is $(1-\epsilon)\pi$ radians, the total phase difference at the end of the two couplers will be π radians. Hence the total input power can be coupled to the secondary waveguide through cascaded couplers.

Diagonal-Arm Condition

The third condition under which a Transvar coupler must pass the total input power is when one pair of diagonal arms of the coupler are connected together. This condition must be satisfied regardless of the degree of coupling of the isolated coupler and the length of the waveguide connecting diagonal arms.

Referring to Fig. 2, the length of the coupling region $a-b$ can have any arbitrary value. The waves in all arms are propagating in the same direction stipulating no reflections from any discontinuities. The following analysis assumes a lossless coupler. The electric fields in arms 1 and 2 at a, the beginning of the coupler, can be resolved into the two propagating modes E_A the antisymmetrical and E_S the symmetrical to get

$$E_A = \frac{E_1 - E_2}{2}$$
$$E_S = \frac{E_1 + E_2}{2}. \tag{2}$$

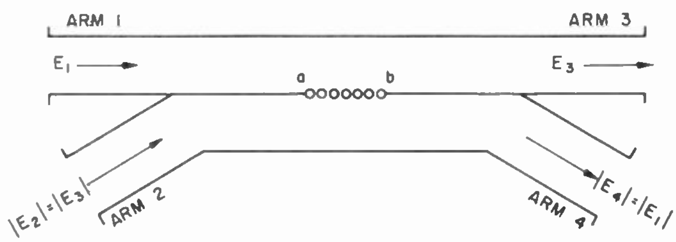

Fig. 2—Diagonal-arm condition.

All electric vectors are complex. If the odd vector E_A is considered the reference vector, the even vector E_S will undergo a relative time phase change of α degrees so that E_3 and E_4 are related by

$$\frac{E_3 + E_4}{2} = \frac{E_1 - E_2}{2} e^{-j\phi}$$
$$\frac{E_3 + E_4}{2} = \frac{E_1 + E_2}{2} e^{-j(\phi+\alpha)}, \tag{3}$$

where ϕ = electrical length of the coupler for the antisymmetrical mode. Let the phase angle between E_3 and E_2 be β. Since one pair of diagonal arms are connected together

$$E_2 = E_3 e^{-j\beta}. \tag{4}$$

Making this substitution in (3) yields the following simultaneous equations:

$$E_3 = \frac{E_4 + E_1 e^{-j\phi}}{1 + e^{-j\phi - j\beta}} \tag{5}$$

$$E_3 = \frac{E_1 e^{-j\phi - j\alpha} - E_4}{1 - e^{-j\phi - j\alpha - j\beta}}. \tag{6}$$

Eliminating E_3 from (5) and (6) and solving yields

$$\left|\frac{E_4}{E_1}\right| = 1. \tag{7}$$

This equality which is independent of the values of α, β, or ϕ proves that if the powers in one pair of diagonal arms are equal then the powers in the other pair of arms must be equal. This is a result of conservation of energy.

By taking small but finite losses into consideration, it can be shown theoretically and experimentally that $|E_4/E_1|$ can become significantly less than unity for certain values of α, β, and ϕ. This is called a "pseudoresonant" condition in the rotary joint and is discussed

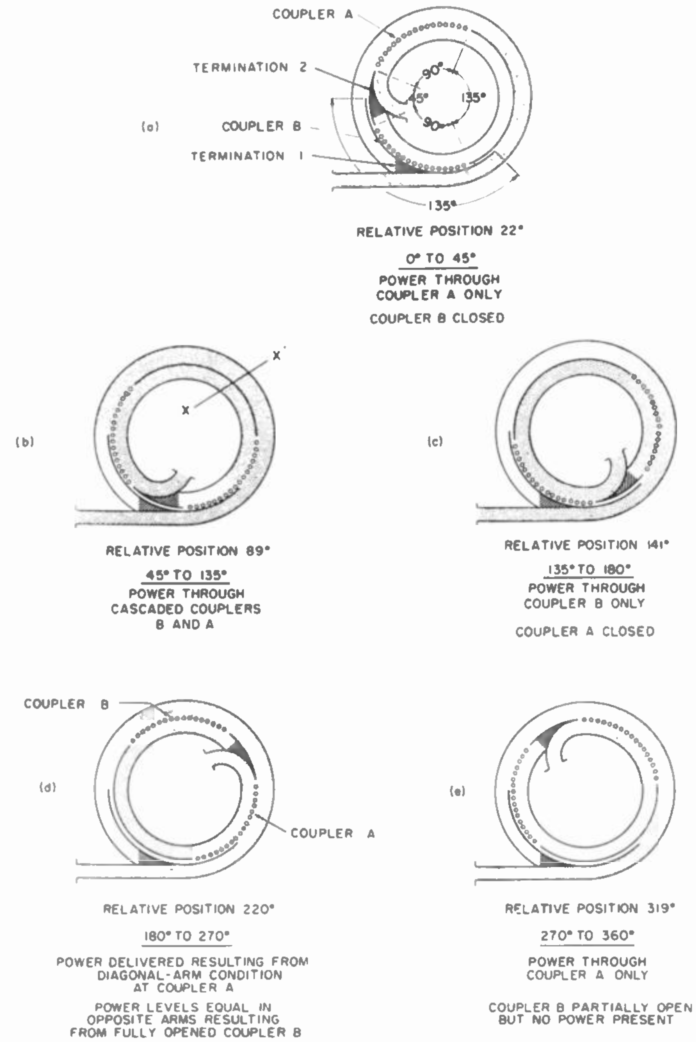

Fig. 3—Schematic diagram showing principle of operation of the annular waveguide rotary joint.

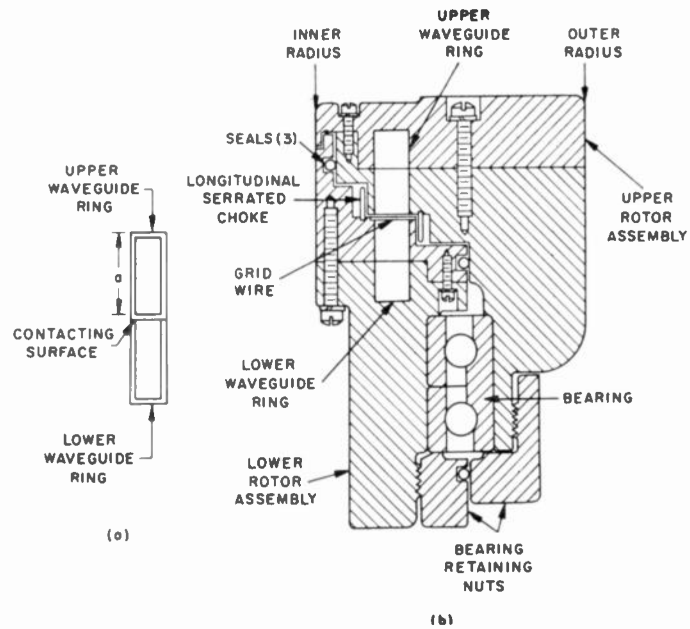

Fig. 4—Cross section of waveguide ring.

in the Appendix. Under this condition, however, it can be shown that $|E_3/E_1|$ will be much greater than unity.

Principle of Operation of the Rotary Joint

The schematic diagram and theory of operation of the rotary joint are illustrated in Fig. 3. This joint actually consists of two E-plane waveguide rings which have equal diameters and are stacked concentrically with a common narrow wall. The diagram shows a schematic plan view for clarity. The ring with the outer terminal has on one narrow wall an opened section 225 degrees long. The ring with the inner terminal has two 90-degree long Transvar couplers spaced 45 degrees apart. These angular dimensions on the rotary joint are arbitrarily chosen for convenience. By proper design a wide selection of angles may be used. When the rings have the relative position, arbitrarily chosen to be approximately 22 degrees, as shown in Fig. 3(a), the transfer of power from one ring to the other is accomplished through coupler A. In Fig. 3(b), the transfer of power is accomplished through the cascaded couplers A and B. In Fig. 3(c), power is transferred only through coupler B since coupler A is closed. In Fig. 3(d), the power is delivered because the powers in diagonal arms of a Transvar coupler are equal. In Fig. 3(e), the power passes through coupler A only. Coupler B is partially open but no power is present. Thus all of the input power appears at the output terminal for any rotational position of the joint.

In order to reduce the effects due to reflections within the rotary joint, terminations were used as shown in Fig. 3(a). Termination 1 absorbs the small amount of power which is not transferred to the coupled ring. Termination 2 absorbs the reflected power due to the finite directivities of the couplers.

In order to preserve the guide wavelength within the opened section of the outer terminal ring (225 degrees long), it is necessary to maintain the guide width constant by soldering an additional thickness of metal against the opposite narrow wall.[3]

An unusual choke problem arose in developing the new rotary joint. Referring to Fig. 3(b), a cross sectional view at $X-X'$ is shown in Fig. 4(a). The input ring has an open narrow wall in contact with the output ring. Relative motion occurs at this contacting surface. The contact was eliminated by using a very low impedance non-contacting "longitudinal" choke as shown in Fig. 4(b).[4] The bearings necessary to support the weight of the assembly is placed outside of this region.

Mechanical Description of the Joint

The aluminum rotary joint shown in Figs. 5 to 9 is composed of two major subassemblies, namely upper and lower rotor assemblies. The bearing retaining nuts are not separately shown. Each rotor assembly has an incomplete ring waveguide comprised of two halves mating at the center of its broad wall. The mean di-

[4] K. Tomiyasu and J. J. Bolus, "Characteristics of a new serrated choke," Trans. IRE, vol. MTT-4, pp. 33–36; January, 1956.

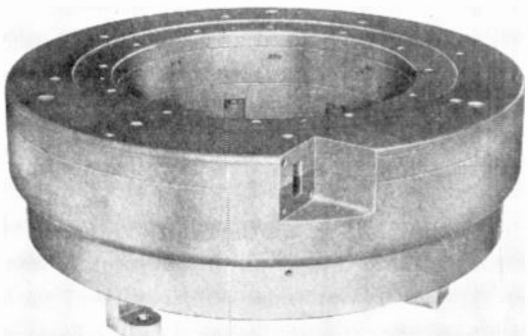

Fig. 5—Annular waveguide rotary joint.

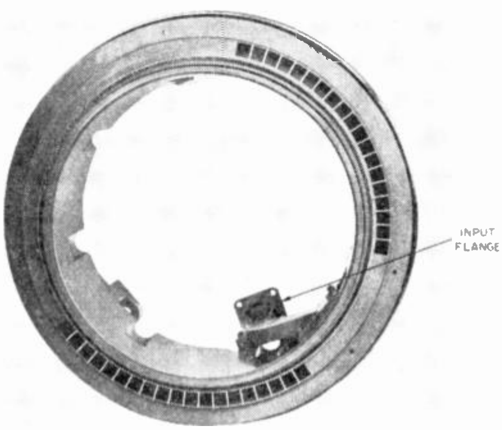

Fig. 6—Top view of lower rotor assembly.

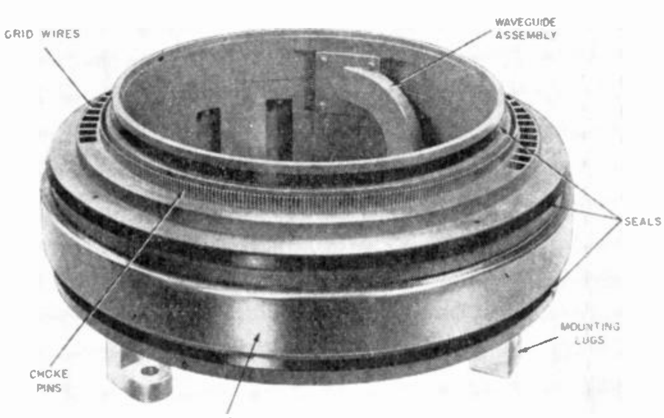

Fig. 7—Side view of lower rotor assembly.

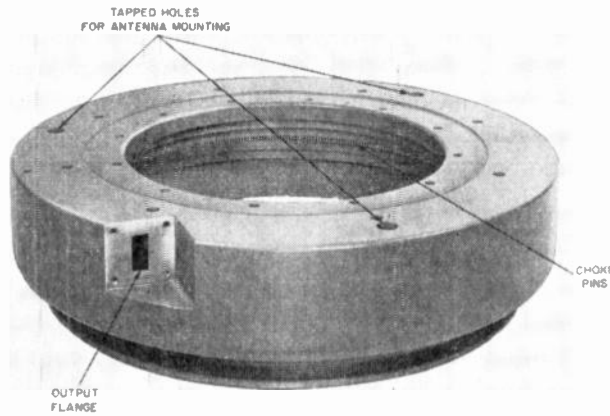

Fig. 8—Side view of upper rotor assembly.

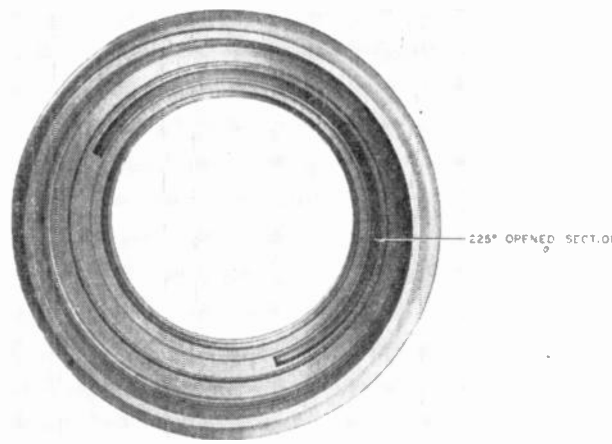

Fig. 9—Bottom view of upper rotor assembly.

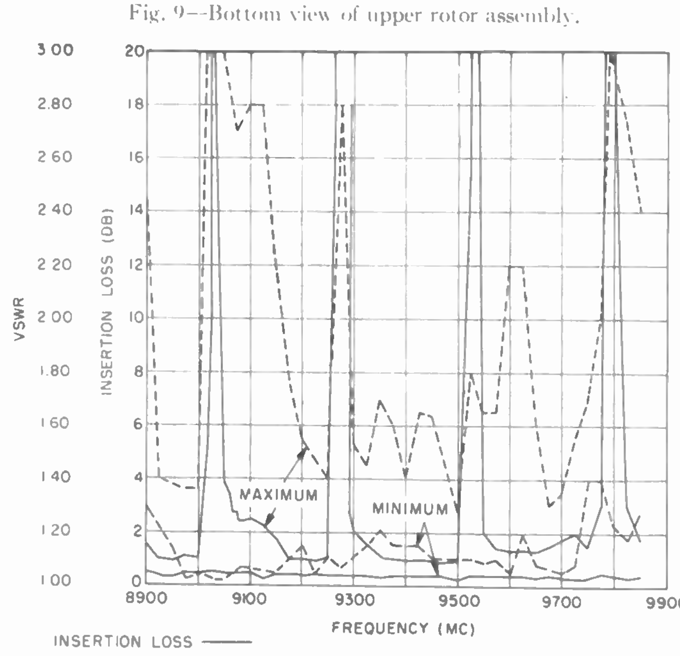

Fig. 10—Maximum and minimum insertion loss and vswr as a function of frequency.

ameter of the 0.400×0.900 inch id waveguide rings is 10.712 inches. The 720 choke pins (0.050-inch diameter) were press fitted every 1 degree on two reference diameters. The weight of this joint is 50 pounds.

Electrical Characteristics

The insertion loss and input vswr of the rotary joint as a function of rotation were measured every 25 mc from 8,900 to 9,850 mc. It was found that the insertion loss and the vswr fluctuated with a periodicity of about 10 degrees of rotation suggesting finite reflections and coupler directivities. Most of these fluctuations were measured in detail, but it was felt that only the maximum and minimum vswr and the maximum and minimum insertion loss data throughout the rotation are significant. These data (Fig. 10) show that low vswr and low insertion loss is obtainable throughout the full rotation for particular frequencies. For example, for all frequencies between 9,350 and 9,500 mc, the maximum insertion loss does not exceed about 1 db for any rotational position.

Pseudo-resonances were observed at 9,027, 9,274, 9,527, and 9,793 mc for rotational positions ranging from 233 to 247 degrees. These resonances were very narrow both in frequency and rotational position. Accurate measurements have shown that at these conditions the insertion loss ranges from about 28 to 50 db.

It is interesting to note that at these resonant frequencies the mean circumference ($C_m=33.653$ inches) of the circular ring waveguide divided by the waveguide wavelength yields the results shown in Table I.

TABLE I

Frequency (mc)	Guide Wavelength (inches)	$C_m/(\lambda_g)_0$
	Calculated for 0.4×0.9 inches Inside Dimension Waveguide	
9,027	1.9023	17.691
9,274	1.7996	18.700
9,527	1.7076	19.708
9,793	1.6222	20.745

In making this comparison it is assumed that the guide wavelength can be accurately computed. However, this is not justified since part of the ring waveguide has an open narrow wall which is choked. There is reason to believe that the phase velocity of the waves traveling in the "choked" waveguide may be smaller by as much as 2 per cent compared to that in an unperturbed waveguide of otherwise identical inside dimensions.[4] Further consideration of these resonances will be made in the Appendix.

In this type of rotary joint the pseudo-resonant conditions cannot be avoided; however, usable bands exist between resonant frequencies. Improvement in the electrical characteristics such as lower insertion loss and vswr in the usable bands can undoubtedly be achieved by increasing the directivity and decreasing the vswr of the couplers under all degrees of power transfer conditions.

High-power measurements were made at 9,375 mc. These have shown that for any rotational position the rotary joint can handle the full-power output from a QK-221 magnetron measured to be about 210 kw. Reference to Fig. 10 indicates that the joint should not be troubled with pseudo-resonances at this frequency.

Appendix

Pseudo-Resonance Investigation

When the rotary joint is operating in the diagonal arm condition, at particular positions and frequencies, resonances will occur that will result in large insertion losses. These resonances are called "pseudo-resonances" because they are due to an actual increase in field strength of the unidirectional waves in the diagonal arm of the rotary joint, and do not theoretically result in high standing waves within the joint.

Referring to Figs. 3(d) and 2, codirectional waves E_1 and E_2 at a in Fig. 2, as well as codirectional waves E_3 and E_4 at b, can be resolved into two other modes: one symmetrical and the other antisymmetrical with respect to the plane of symmetry in the coupler. Due to a difference in phase velocity of the resolved modes, a transfer of energy will occur in the coupler.

The antisymmetrical mode will undergo a phase change and attenuation through the coupler in accordance with

$$\frac{E_3 - E_4}{2} = [fe^{-i\phi}]\frac{E_1 - E_2}{2}. \tag{8}$$

By referring the symmetrical mode to the asymmetrical mode,

$$\frac{E_3 + E_4}{2} = [ge^{-i(\phi+\alpha)}]\frac{E_1 + E_2}{2}. \tag{9}$$

By virtue of the ring waveguide and Transvar coupler B in Fig. 3(d), E_3 and E_2 are related in accordance with

$$E_2 = [ke^{-i\beta}]E_3. \tag{10}$$

The scalar terms f, g, and k account for any attenuation. By substituting (10) in (8) and (9) to eliminate E_2 and then equating E_3, the following equation relating E_4 and E_1 is obtained:

$$\frac{E_4}{E_1} = \frac{e^{-i\phi}e^{-i\alpha}}{e^{-i\theta}} \frac{g - fe^{+i\alpha} + 2fgke^{-i\theta}}{2e^{+i\theta} - gke^{-i\alpha} + fk}, \tag{11}$$

$$\text{where} \quad \theta \equiv \phi + \beta.$$

For maximum insertion loss, (11) approaches zero. Because the absolute magnitude of the denominator cannot exceed a value of 4 and is always greater than zero for finite losses, the numerator must approach zero to obtain maximum insertion loss. For convenience, part of the numerator can be redefined as

$$N \equiv g + fe^{i(\pi+\alpha)} + 2fgke^{-i\theta}. \tag{12}$$

A polar representation of (12) is illustrated in Fig. 11. By virtue of its definition θ is dependent only upon the frequency and not on α in any way. Similarly, α is dependent only on the rotational position of the joint and is independent of θ. For N to equal zero, two conditions must be simultaneously satisfied: one for α_0 and the other for θ_0. The condition for α is given by:

$$\cos\alpha_0 = (1/2)\left[\frac{g}{f} + \frac{f}{g}\right] - 2fgk^2. \tag{13}$$

The range of α_0 is from 0 to 180 degrees.

The rotational position, γ, of rotary joint in diagonal arm condition for resonance is given by the relation

$$\gamma_0 = \pi + \frac{\alpha_0}{2} \cdot \tag{14}$$

Second condition to be satisfied for (12) to equal zero;

$$\cos \theta_0 = \frac{f}{4g^2k} - \frac{1}{4fk} - fk. \tag{15}$$

For pseudo-resonance to occur both (13) and (15) must be satisfied simultaneously.

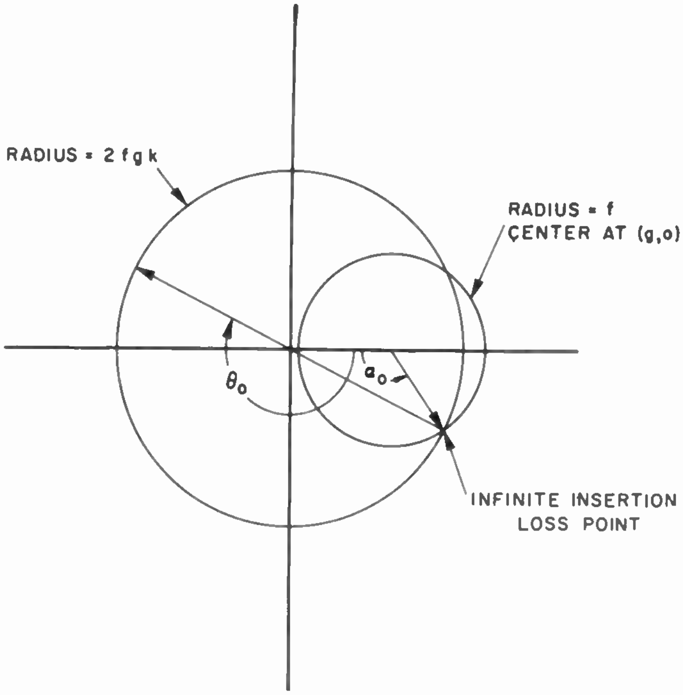

Fig. 11—Polar representation of $g+fe^{j(\pi+\alpha)}+2fgke^{-j\theta}$.

By assuming $f=g$, pseudo-resonance will occur when

$$\alpha_0 = 2 \sin^{-1}(fk) \tag{16}$$

and

$$\theta_0 + \pi + \cos^{-1}(fk) + 2\pi n \tag{17}$$

where n is any integer. Referring to Fig. 11,

$$90° > \cos^{-1}(fk) > 0°. \tag{18}$$

Equating (16) and (17) results in

$$\theta_0 = \frac{3\pi}{2} - \frac{\alpha_0}{2} + 2\pi n. \tag{19}$$

Eqs. (17) and (19) reveal that the pseudo-resonant frequencies are related by integral-wavelength differences in the electrical lengths of the ring circumference θ. Because coupler "B" in Fig. 3(d) introduces a π radian phase lag, the mean ring circumference C_m is related to θ_0 by the following:

$$\frac{C_m}{(\lambda_g)_0} = \frac{\theta_0 - \pi}{2\pi} = n + \frac{1}{2\pi}\cos^{-1}(fk). \tag{20}$$

Note that if the losses in the joint approach zero, $C_m/(\lambda_g)_0$ approaches n.

In the rotary joint, the following Table II was obtained at the pseudo-resonant conditions:

TABLE II

Frequency (Measured)	γ_0 (Meas.)	α_0 [From (14)]	k db [From (16)]	θ_0 [From (19)]	$\frac{C_m}{(\lambda_g)_0}$ [From (20)]
9,027 mc	233°	106°	1.55*	217°	18.103
9,274 mc	241°	122°	1.0†	209°	19.081
9,527 mc	247°	134°	0.5†	203°	20.064
9,793 mc	237°	114°	1.1*	213°	21.092

* Assuming $f=g=.9550$ (0.4 db).
† Assuming $f=g=.9772$ (0.2 db).

In measuring the rotational position γ_0 for maximum insertion loss, it should be pointed out that because of multiple reflections within the joint, there may be a few degrees error in γ_0. Such a small error could materially change the calculated value of k. The order of magnitude of k was measured at $\gamma=180$ degrees to be about 0.6 to 1.0 db; these values appear to be in substantial agreement with the calculated values.

As to the mean circumference of the ring waveguide, a 0.4 wavelength difference exists between the above calculated data and the data shown in the section on Electrical Characteristics based upon a theoretical phase velocity. The difference in electrical length of the mean circumferences can be attributed to an average difference in phase velocity of only 2 per cent. Such a difference, due to the "choked" waveguide section, appears entirely plausible.

Acknowledgment

The author wishes to acknowledge Gary Silverman's aid in the mechanical design of the rotary joint and the assistance by Walter B. Dennen, Jr. in preparing the illustrations.

A Double-Slab Ferrite Field Displacement Isolator at 11 KMC*

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Summary—**A double-slab ferrite field displacement isolator has been constructed for the 10.7–11.7 kmc band with the following performance: reverse loss of 70 db from 10.8–11.7 kmc and 64 db at 10.7 kmc; forward loss less than 1 db, except 1–1.2 db between 11.6 and 11.7 kmc; vswr less than 1.15, except 1.15–1.20 between 11.6 and 11.7 kmc; variation in forward loss less than 0.1 db for any 20 mc channel.**

Introduction

THE PRINCIPLES of field displacement isolators in rectangular waveguide have been discussed in the literature.[1] A theoretical and experimental study of a single-slab isolator has been made[2] for the 6,000 mc region. This device utilizes a single slab of ferrite, transversely magnetized, and suitably positioned in the waveguide. Over an eight per cent band a 30 db reverse loss, a 0.2 db forward loss, and a vswr of 1.05 are obtained.

While an isolator with this performance will undoubtedly be quite satisfactory for many applications, some particular applications can be visualized where different characteristics are desired. Considerations of this type have led us to explore the possibility of constructing a broadband isolator with a much larger reverse loss. The double-slab isolator described here is a result of this investigation in the 10.7–11.7 kmc band. This type of isolator is preferable to several single-slab isolators in tandem from the point of view of compactness and economy.

The operation of the double-slab isolator depends upon the use of two slabs, symmetrically disposed with respect to the center line of a waveguide. The slabs have the same thickness and magnetic properties, and the dc magnetic fields are equal in magnitude but oppositely directed in the two slabs. The basis for such a geometry can be seen from the calculations of electric field distributions for forward and reverse directions of propagation made by Lax, Button, and Roth[3] for both the single-slab and double-slab cases. Typical calculated field distributions are shown in Figs. 1 and 2 in the next column.

* Original manuscript received by the IRE, October 28, 1955.
† Bell Telephone Laboratories, Incorporated, Murray Hill, New Jersey.
[1] A. G. Fox, S. E. Miller, M. T. Weiss, "Behavior and applications of ferrites in the microwave region," *Bell Syst. Tech. Jour.*, vol. 34, pp. 65–76; January, 1955.
[2] S. Weisbaum and H. Seidel—Submitted for publication in *Bell Syst. Tech. Jour.*
[3] B. Lax, K. J. Button, and L. M. Roth, "Ferrite phase shifters in rectangular waveguide," M.I.T. Lincoln Laboratory, TM No. 49; November 2, 1953.

The difference in electric fields at the resistance strips for opposite directions of propagation results in a large difference in attenuation for the forward and reverse waves.

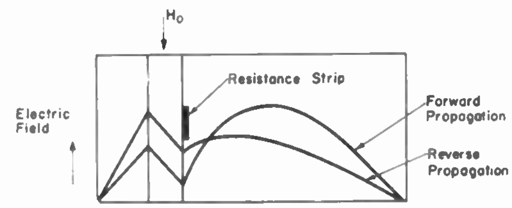

Fig. 1—Electric field distributions in single-slab isolator.

In Fig. 2, the rf polarizations on the two sides of the guide where the slabs are located have opposite senses of rotation, and hence interact equally with the electron magnetic moments which are spinning oppositely in the two slabs. This enhances the field displacement effect indicated in Fig. 1.

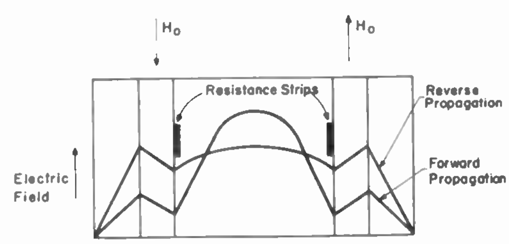

Fig. 2—Electric field distributions in double-slab isolator.

Design Considerations and Results

We have constructed a double-slab field displacement isolator, using partial height slabs for purposes of better match. The geometry of the model is shown in Fig. 3.

The ferrite slabs were supplied by L. G. Van Uitert of Bell Telephone Laboratories and are a modified nickel ferrite of saturation magnetization 3,850 gauss. We varied H_0 until best results were found. This occurred at 1,045 oersteds. As yet, the effects of variations of slab position, ferrite thickness, and ferrite height, and resistivity of the resistance material have not been studied in detail. The parameters used in the model are indicated in Fig. 3.

Starting with the 6 kmc single slab isolator, which gave nearly optimum results, we applied the scaling principle[2] to determine the best parameters for an 11 kmc single-slab isolator. The double-slab isolator then follows from the principles discussed in the introduction and has been found to yield forward and reverse losses slightly more than twice as large as observed in the corresponding single-slab isolator. A possible explanation for this enhancement in loss is that the symmetric field configuration of the double-slab isolator eliminates

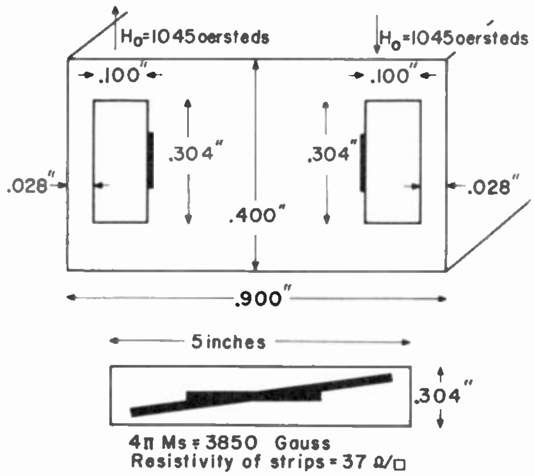

Fig. 3—Geometry of double-slab field displacement isolator and disposition of resistance material. Performance of this model is shown in Fig. 4.

odd modes, leaving symmetric field distributions which can interact favorably with the resistance strips. In the single-slab case the asymmetric higher modes have smaller electric fields at the ferrite face and therefore are not as readily absorbed as symmetric mode fields by the resistance strip. Another possibility is that the partial height, double-slab geometry is even more favorable to the existence of longitudinal components of electric field than is the single-slab case. It is believed[2] that, in the partial height, single-slab case, these longitudinal components are responsible for a portion of the reverse loss.

At 6 kmc the optimum performance was obtained when an electric field null was made to exist at the ferrite face for the forward propagation direction.[2] The scaling principle maintains this null at 11 kmc for the single-slab case. The addition of a second slab will, in general, negate this null. However, Mr. H. Seidel of Bell Telephone Laboratories has shown theoretically that, under proper operating conditions, a new position of the slabs may be found for which a double null exists.

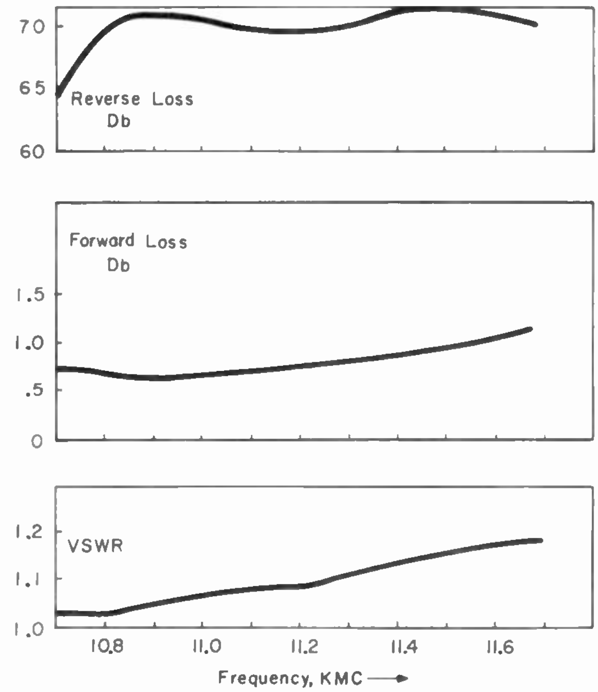

Fig. 4—Performance of double-slab isolator of Fig. 3.

The performance of the isolator is shown in Fig. 4. Reverse loss is of the order of 70 db over the band with a minimum of 64 db at 10.7 kmc. Forward loss is of the order of 1 db or less, except at the end of the band (11.6–11.7 kmc) where it increases to slightly more than 1 db. VSWR is low except at the end of the band where it approaches 1.15–1.20. The variation in forward loss in any 20 mc channel is less than 0.1 db. As yet, the double null predicted theoretically has not been looked for experimentally.

Acknowledgment

We should like to thank E. F. O'Neill for making the measurements reported in Fig. 4, E. H. Turner for access to some unpublished notes, and H. Seidel for valuable discussions.

Correspondence

A Note on Sidebands Produced by Ferrite Modulators*

In devices employing ferrite rotators as modulators, the usual equation for amplitude modulation is valid when the plane of polarization is rotated through a small angle. However, when large values of Faraday rotation are required to increase sideband energy, this relationship no longer holds. In this case, a sinusoidal current applied to the magnetizing coil produces a modulation envelope containing all the harmonics of the modulation frequency. For example, consider the vector A (Fig. 1) which represents the polarization of the microwave energy. When the coil is energized with a sine wave of frequency w', the rotation of the plane of polarization (θ) will be given by:

$$\theta = K \sin w't^{1}$$

where K = maximum rotation angle.

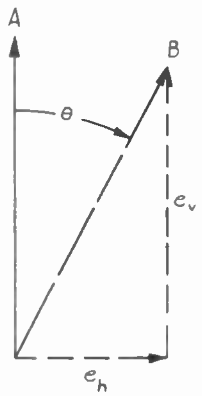

Fig. 1—Representation of a vector exhibiting Faraday rotation.

Analyzing the vertical and horizontal components of the rotated vector B, the following is obtained:

Horizontal Polarization:

$$e_h = E \sin (K \sin w't)$$

Vertical Polarization:

$$e_v = E \cos (K \sin w't)$$

where E is the magnitude of the vector B.

Because of the orthogonal properties of the Bessel function, the above expressions can be expanded in the following manner:

$$\begin{aligned} e_v &= E \cos (K \sin w't) \\ &= E[J_0(K) + 2J_2(K) \cos 2w't \\ &\quad + 2J_4(K) \cos 4w't + \cdots] \\ e_h &= E \sin (K \sin w't) \\ &= 2E[J_1(K) \sin w't \\ &\quad + J_3(K) \sin 3w't + \cdots]. \qquad (1) \end{aligned}$$

Since these equations are the modulating functions of the microwave, the vertical component will contain the carrier frequency w, plus the sidebands $(w \pm 2w')$, $(w \pm 4w')$, etc. Examining the expression for e_v, it is seen that the first two terms give the modulating function for sine wave amplitude modulation.[2] Similarly the first term of e_h is the amplitude function for carrier suppressed modulation. In both cases the presence of the other sidebands represents a deterioration of the modulation. Eq. (1) shows that the amplitudes of the undesired sidebands are functions of the maximum rotation (K). The relative amplitude of both the wanted and unwanted sidebands for e_h are plotted in Fig. 2 as a function of K.

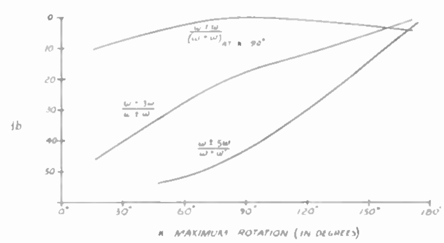

Fig. 2—Relative amplitudes of sidebands in e_h.

Fig. 3 shows the ratio in decibels of the unwanted sidebands to the desired sidebands for e_v as a function of K. Also plotted is the ratio of desired sideband level to carrier level. From the figure it can be seen that at $K = 90°$ the desired sideband energy is approximately 6 db down from the carrier energy and the unwanted sideband is 25 db below the desired sideband. Thus for this and lesser values of K, the modulating wave approximates, to a good degree, sine wave amplitude modulation.

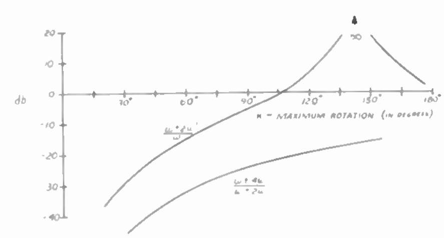

Fig. 3—Relative amplitudes of sidebands in e_v.

It is interesting to note that the carrier level is also a function of K, and that when $K = 140°$, the carrier is reduced to zero

$$\left(i.e., \frac{w \pm 2w'}{w} \to \infty \right),$$

and only sidebands are generated.

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* Received by the IRE, October 21, 1955.

[1] This equation assumes that the ferrite is operated in the linear region of its magnetization curve.

[2] Note that the envelope has twice the frequency of the modulating signal.

Principles of Communication Systems*

The book review section of the PROCEEDINGS often presents enlightening material. No exception is Mr. Marchand's review of my book "Principles of Communication Systems" which appeared in the November issue.[1] It would appear that I have written a book with a fairly good table of contents, no contents to speak of, yet the book is found interesting. If true, a neat trick!

The point is made that some of the material could have been covered in radio courses. The book, as the preface points out, is in part intended for those who take no radio courses. However, they are expected to have a prior knowledge of, or junior courses in, circuit theory and basic electronics. Also, the reviewer is dissatisfied with the limited use I make of information theory. It must be a somewhat sobering experience to those who have become enamored of the newer trend and terminology in information theory to find that most of the communication systems now in use not only were invented but were refined to the extent that their basic limitations were quite well understood before the advent of the new emphases. This is not to say that the new developments would not have been useful had they arrived earlier. In the development of new systems, there still is no substitute for human inventiveness and ingenuity. This is a point which I attempted to emphasize in the book.

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* Received by the IRE, December 22, 1955.

[1] W. D. Hershberger, "Principles of Communication Systems," reviewed by Nathan Marchand, PROC. IRE, vol. 43, p. 1679; November, 1955.

The Equivalent Characteristics of the Cascode Amplifier*

An error in the equivalent characteristics of the Cascode Amplifier on page 534 of the "Radiotron Designer's Handbook" has been found by Dr. L. B. Hedge of Washington, D. C. The amplification as correctly given by (3) may, by simple manipulation, be put into the form:

$$A' = \frac{1}{\dfrac{\mu + 2}{g_m(\mu + 1)R_L} + \dfrac{1}{\mu(\mu + 1)}}. \qquad (4)$$

By making R_L infinite we determine μ' which is $\mu(\mu+1)$. Hence $g_m' = g_m(\mu+1)/(\mu+2)$ and $r_p' = (\mu+2)r_p$.

Eq. (4) as printed in the "Handbook" is numerically correct, but not in a form suitable for the application made.

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* Received by the IRE, January 16, 1956.

Fourier Transforms and Tapered Transmission Lines*

Because of the rather general interest in tapered transmission lines, as indicated by numerous articles in this journal [1–4] and in other journals [5–6], the following comments may be appropriate. In a letter to the Editor [7] I pointed out that Fourier transforms can be used for an approximate theory of tapered transmission lines. The results quoted in the letter were drawn from a thesis "Fourier Transforms in the Theory of Inhomogeneous Transmission Lines," which was published in the Transactions of the Royal Institute of Technology, Stockholm, Sweden, in 1951 [8]. Since this thesis may not be readily available in the United States, the work may have escaped the attention of many fellow workers in the tapered-line field. I should therefore like to give a brief summary of the work here.

Pierce [9] has shown that the well-known second-order differential equations for the voltage and for the current in a tapered line, valid under the assumption of a plane field, can be compressed into a single first-order nonlinear differential equation in impedance. This differential equation, a Riccati equation, can be written in the form

$$Z_0 \frac{dz}{du_1} + Z_0{}^2 - z^2 = 0 \tag{1}$$

where

$$u_1 = \int_0^x \gamma dx \tag{2}$$

and Z_0 is the characteristic impedance varying with the distance x, z is the ratio of voltage and current, and γ is the propagation constant.

Now the problem is: How can this nonlinear differential equation be linearized so that the approximate solution obtained after the linearization process has been performed is an accurate one? It is obvious that substitutions have to be made so that the nonlinear squared term will be negligible compared to the other terms. This can be done by the two substitutions

$$\frac{z}{Z_0} = \psi \tag{3}$$

and

$$\psi = \frac{1+\rho}{1-\rho}. \tag{4}$$

Here ρ is the complex reflection coefficient. (The differential equation originating after the first substitution (3) has been performed was used by Burkhardtmaier [10].) Then (1) transforms into

$$\frac{d\rho}{du_1} + \frac{1}{2}\frac{d \ln Z_0}{du_1}(1-\rho^2) - 2\rho = 0. \tag{5}$$

Just before submitting the manuscript I found that (5) had been derived by Walker and Wax [11] and a reference was included in the thesis. Eq. (5) fulfills the condition looked for because, for tapered lines with small tapers, $\rho^2 \ll 1$. If ρ^2 is neglected in (5) the Riccati differential equation is transformed into a first-order linear differential equation

$$\frac{d\rho}{du_1} - 2\rho + \frac{1}{2}\frac{d \ln Z_0}{du_1} = 0. \tag{6}$$

This constitutes the main idea of the thesis. The solution of (6) is

$$\rho_{x=0} = \int_0^l \frac{1}{2}\frac{d \ln Z_0}{dx} \exp\left(-2\int_0^x \gamma dx\right) dx \tag{7}$$

which, if $\gamma = \alpha + j\beta$ is constant ($\beta = 2\pi/\lambda$, λ = wavelength) and if there are no losses, simplifies to

$$\rho_{x=0} = \int_0^l \frac{1}{2}\frac{d \ln Z_0}{dx} e^{-2j\beta x} dx \tag{8}$$

where l is the length of the tapered line.

A physical picture of (8) can be obtained by building it up in the following way. If the reflections from a length dx of the tapered line is $P(x)$ at a distance x from the sending end, then

$$d\rho_{x=0} \approx P(x)e^{-2j\beta x}dx. \tag{9}$$

Integrating

$$\rho_{x=0} = \int_0^l P(x)e^{-2j\beta x}dx. \tag{10}$$

Now

$$P(x)\Delta x = \frac{Z_{0(x+\Delta x)} - Z_{0x}}{Z_{0(x+\Delta x)} + Z_{0x}} \approx \frac{\Delta Z_{0x}}{2Z_{0x}} \tag{11}$$

or

$$P(x)\Delta x \approx \tfrac{1}{2}\Delta \ln Z_{0x}. \tag{12}$$

If Δx approaches 0 we get

$$P(x) = \frac{1}{2}\frac{d \ln Z_0}{dx}. \tag{13}$$

A combination of (10) and (13) yields (8).

Shifting the origin of the coordinate system

$$y = x - \frac{l}{2} \tag{14}$$

and considering that there are no reflections for $-\infty < y < -l/2$ and $l/2 < y < \infty$, (8) can be written

$$\rho\left(\frac{2}{\lambda}\right) = \int_{-\infty}^{\infty} \frac{1}{2}\frac{d \ln Z_0}{dy} e^{-j2\pi(2/\lambda)y}dy$$
$$= \int_{-\infty}^{\infty} P(y)e^{-j2\pi(2/\lambda)y}dy. \tag{15}$$

This is a Fourier integral. Its mate is

$$P(y) = \int_{-\infty}^{\infty} \rho\left(\frac{2}{\lambda}\right) e^{j2\pi(2/\lambda)y} d\left(\frac{2}{\lambda}\right) \tag{16}$$

This Fourier pair (15) and (16) makes it possible to determine the reflection coefficient as a function of the wave number $k = 1/\lambda$, when the variation of the characteristic impedance Z_0 is known along the line, and conversely. The calculations may be simplified by using tables and by applying results from other fields where Fourier transforms have found applications; for instance, in the theory of communication and the antenna theory.

Besides a discussion of other approximate theories of tapered transmission lines—for example, theories by Burkhardtmaier, Frank, Hansen, and Smith—numerous examples are treated in the thesis, both for continuous tapered lines and so-called step lines. In the thorough study of step lines main interest was focused on three special cases: 1) uniform distributions of point reflections, 2) binomial distributions of the point reflections, and 3) reflection distributions that optimize the relationship between passband width and reflection-coefficient level in the passband. In the third example the well-known results of Dolph, who by using Tchebycheff polynomials obtained optimized broadside-array current distributions in the antenna theory, were interpreted in a nonuniform line sense. Optimized step lines have also been calculated by Burkhardtmaier [10].

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High-Frequency Shot Noise in *P-N* Junctions*

The conductance of a p-n junction increases at high frequencies; so does the shot noise current. A quantitative relation between these effects will be derived on the basis of a simple physical mechanism.

The assumptions are the usual ones of p-n junction analysis. Carriers are supposed to traverse the space-charge region instantaneously and in mutually independent fashion (because the charge of the carriers is negligible in comparison to the fixed charge in the space-charge region). Because of this independence, no "space-charge smoothing" is involved. In neutral p and n regions,

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minority carriers diffuse, drift, and recombine, and are also thermally generated—each independently of all others.

There are three kinds of behavior to be considered. First, thermally-generated minority carriers that find their way to the space-charge region give rise to a saturation current i_s in the reverse direction. Second, majority carriers cross the junction in the forward direction (thereby becoming minority carriers) and do not return; a current i_0+i_s must flow by this means, where i_0 is the net direct current (taken positive in the forward direction).

Third—and responsible for the high-frequency effects—carriers cross the junction in the forward direction and return after spending some time as minority carriers. The number of these carriers must be proportional to the number that do not return, because the fate of the emitted carriers is not predetermined.

At any frequency, the third type of behavior adds to the conductance and shot noise current. The conductance may be written

$$G_\omega = \frac{q}{kT}(i_0 + i_s)(1 + F_\omega), \qquad (1)$$

where F_ω represents the effect of carriers of the third kind. The magnitude of the charge of a single carrier is denoted by q. Similarly, the mean-square noise current may be written:

$$\frac{\langle i_\omega^2\rangle}{B} = 2qi_s + 2q(i_0 + i_s)(1 + H_\omega), \qquad (2)$$

where H_ω represents the effect of carriers of the third kind. At equilibrium, thermal noise must prevail; *i.e.*, $\langle i_\omega^2\rangle/B = 4kTG_\omega$ for $i_0 = 0$ in both (1) and (2). From this relation one finds that

$$H_\omega = 2F_\omega. \qquad (3)$$

Thus, the theoretical shot noise can be calculated easily from measured or theoretical conductances.

Above equations when rearranged give

$$\frac{\langle i_\omega^2\rangle}{B} = 4kTG_\omega + 2qi_0, \qquad (4)$$

a result obtained in a recent one-dimensional analysis of the details of diffusion and volume recombination.[1] The present derivation shows that this result is valid also for nonplanar junctions and in cases where drift in a steady electric field is important.

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[1] A. van der Ziel, "Theory of shot noise in junction diodes and junction transistors," Proc. IRE, vol. 43, pp 1639–1646; November, 1955.

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J. H. Burnett (M'51) received his bachelor's degree in engineering in 1932 and his electrical engineering degree in 1933, both from Princeton University.

He has published several technical articles and is the author of Thyratron Amplifiers, a chapter in the forthcoming book "Servomechanism and Feedback Control Handbook," to be published by McGraw-Hill Book Co. He holds several patents on control tubes and circuits.

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C. J. Creveling was born in Bloomsbury, N. J., on March 5, 1918, and received the B.E.E. degree from the University of Florida in 1941.

While serving in the United States Army Signal Corps from 1941 to 1946, he attended the Military College of Science and served at the Telecommunications Research Establishment in England where he was engaged in design of microwave test equipment. Later he was attached to the Naval

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D. G. Fink (A'35–SM'45–F'47) was born on November 8, 1911, at Englewood, N. J. He received the B.S. degree in electrical engineering from the Massachusetts Institute of Technology in 1933 and the M.S. degree in electrical engineering from Columbia University in 1942.

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During 1933–1934, Mr. Fink was a research assistant in the M.I.T. departments of electrical engineering and geology. From 1934 to 1952 he was on the editorial staff of *Electronics*, becoming Editor-in-Chief in November, 1946. Since 1952 he has been associated with the Philco Corp. as Director of Research for radio, television and appliances.

Mr. Fink designed the standard loran transmitter in 1941–1943, when on leave to the Radiation Laboratory at M.I.T. In 1946, he was a civilian consultant of the Commander, Joint Task Force One, in charge of preparing damage reports on all electronic material and test facilities for the Bikini atomb bom tests.

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He received the B.S. and M.S. degrees in electrical engineering from the University of California, Berkeley in 1950 and 1955. From 1948 to 1949, and 1950 to 1951 he was employed as a design engineer at the San Francisco Naval Shipyard. From 1951 to 1952 he worked on the development of instruments for the biology section of the U. S. Naval Radiological Defense Laboratory, San Francisco. From 1952 to 1955 he worked on the development of instruments for the Radioisotope Research Unit, V. A. Hospital, San Francisco. He is now working on the development of instruments for Ames Aeronautical Laboratory, NACA, Moffett Field, Calif., and is studying at Stanford.

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J. J. Kleimack was born in Bayonne, N. J., on November 26, 1912. He became a member of the Bell Telephone Laboratories in 1930 and joined the Research Department as a Technical Assistant working on copper oxide rectifiers. From 1938 to 1949 as technical assistant and later as a member of the technical staff, his major effort was devoted to the development of high speed thermistors and thermistor bolometers, partly in connection with N.D.R.C. contracts for the army and the navy. Since then he has engaged in exploratory development of transistors, mainly point contact and alloy junction types.

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From 1948 to 1952, Dr. Le Caine studied at the University of Birmingham, England, on a National Research Council grant, where he obtained his Ph.D. From 1952 to date, Dr. Le Caine has been back in the laboratories of the National Research Council where he has devoted most of his time to the development of electronic musical instruments.

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A. J. Wahl

He became a member of the technical staff at Bell Telephone Laboratories in 1950, but was recalled to active duty in the Air Force Air Research and Development Command in 1951, and spent the following two years on the Air Force transistor development and application program. For his services in this capacity he received the Legion of Merit. After release from his second tour of active duty he returned to the Bell Telephone Laboratories, where he is now engaged in the exploratory development of transistors.

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IRE Awards, 1956

Medal of Honor Award

JOHN V. L. HOGAN

For his contributions to the electronic field as a founder and builder of the Institute of Radio Engineers, for the long sequence of his inventions, and for his continuing activity in the development of devices and systems useful in the communications art.

Morris Liebmann Memorial Prize

KENNETH BULLINGTON

For his contributions to the knowledge of tropospheric transmission beyond the horizon, and to the application of the principles of such transmission to practical communications systems.

Browder J. Thompson Memorial Prize

JACK E. BRIDGES

For his paper entitled, "Detection of Television Signals in Thermal Noise," which appeared in the September, 1954 issue of the PROCEEDINGS OF THE IRE.

Harry Diamond Memorial Award

WILBUR S. HINMAN, JR.

For his contributions to the electronic art in the fields of meteorology and proximity fuzes.

Vladimir K. Zworykin Television Prize

FRANK J. BINGLEY

For his contributions to colorimetric science as applied to television.

New Fellows

S. N. Alexander

For contributions to the development and application of digital computers.

N. G. Anton

For contributions to the design and production of power, counter, and voltage regulator tubes.

W. S. Bachman

For contributions to the recording and reproducing of sound.

G. W. Bailey

For sustaining service to amateur radio, and administrative leadership.

W. J. Barkley

For pioneering and management in the field of electronic communication.

H. E. M. Barlow

For contributions to engineering education, telecommunication, and high-frequency techniques.

L. E. Barton

For contributions to radio engineering, including inventions in Class-B amplification.

R. E. Beam

For contributions to education and research in the fields of microwave theory and techniques.

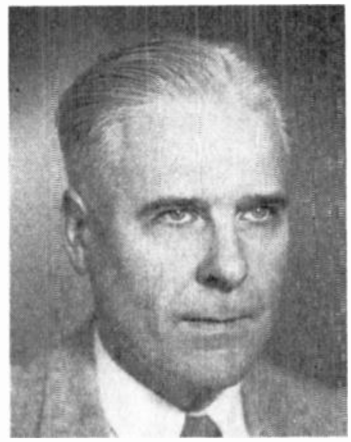
J. E. Beggs

For contributions to the development of new designs of radio tubes.

New Fellows

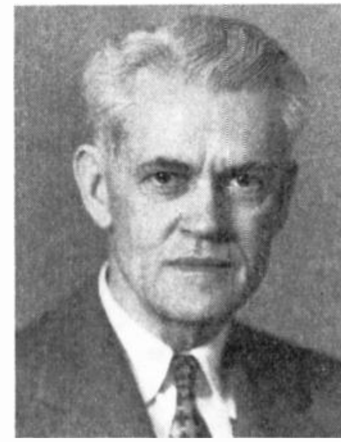
W. H. BELTZ (deceased)

For leadership in improving the reliability of military electronic systems.



W. R. BENNETT

For contributions in the fields of circuit and transmission theory.



E. M. BOONE

For contributions as an educator and research investigator in the field of electronics.

W. P. BOOTHROYD

For contributions to the development of microwave relays, multiplex equipment, and television receivers.

P. N. BOSSART

For contributions to railway safety and operating efficiency through electronic communications.

A. B. BRONWELL

For contributions to radio science as teacher and author.

A. S. BROWN

For contributions to, and leadership in, military electronics.

H. T. BUDENBOM

For contributions to electronic navigation and to precision military radar systems.

R. D. CAHOON

For contributions to international shortwave services and to Canadian broadcasting.

New Fellows

H. J. CARLIN

For advances in microwave network synthesis.

A. B. CLARK (deceased)

For early development and leadership in the field of telephonic transmission systems.

G. F. CORCORAN

For contributions to electrical engineering education and to the associated literature.

T. M. DAVIS

For contributions in the field of military radio communication.

E. N. DINGLEY, JR.

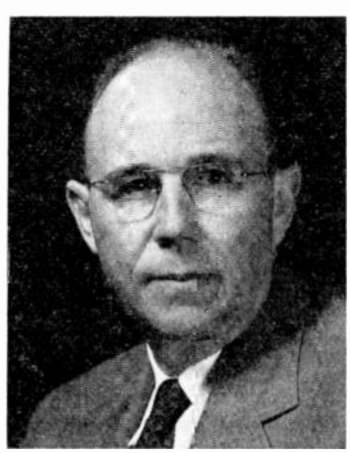
For contributions in the fields of electronic guidance and detection systems.

O. S. DUFFENDACK

For contributions to spectroscopy and gaseous electronics, and for leadership in electronic research.

J. P. ECKERT, JR.

For contributions to electronic digital computation.



H. E. EDGERTON

For contributions in the application of electronic techniques to high-speed stroboscopic photography.

G. A. ESPERSEN

For contributions in the fields of thermionic emission and electron tubes.

New Fellows

C. E. Fay

For contributions to the development of high-power vacuum tubes.

W. G. H. Finch

For contributions in the fields of record and radio communications.

A. G. Fox

For research and invention in microwave waveguide techniques.

A. M. Glover

For contributions to the development of photo tubes.

S. Goldman

For contributions to signal and circuit theory and to electro-medical research.

L. Goldstein

For contributions in the field of microwave gaseous electronics.

J. V. N. Granger

For leadership in research and contributions to aircraft radiation systems.

N. I. Hall

For contributions to the measurement of the velocity of radio waves and the design of radar and guidance systems.

D. B. Harris

For contributions to telephonic communication practices and to the organization of electronic research and development.

New Fellows

R. C. HERGENROTHER

For contributions in electron optics and storage tube development.

J. E. HOBSON

For leadership in organization for research.

J. C. JENSEN

For research in the field of static charges on aircraft and their relation to radio communication.

L. C. JESTY

For leadership and personal contributions in the development and evaluation of television systems.

H. P. KALMUS

For contributions in the fields of electromechanical devices and electronic measurement instruments.

M. E. KENNEDY

For contributions in flood control and civil defense communication systems.

G. KOEHLER

For contributions to engineering education and educational broadcasting.

N. I. KORMAN

For contributions in the field of radar fire control and missile guidance.

K. LEHOVEC

For basic work in the field of semiconduction.

New Fellows

H. W. LEVERENZ

For contributions to the field of luminescence as applied to electronic devices.

H. F. MAYER

For contributions in the development of airborne radar systems.

G. MCELRATH

For contributions in the development of broadcasting and television operating practices and techniques.

M. D. MCFARLANE

For contributions in the development of facsimile and radar.

J. Z. MILLAR

For administrative contributions to military communication, and for commercial application of microwave systems.

B. F. MILLER

For improvements in sound recording techniques, and for the application of electronics in isotope separation.

ROLF MOLLER

For contributions to the development of television in Germany.

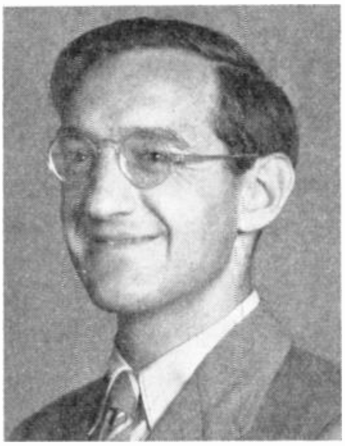
R. C. NEWHOUSE

For his work in the fields of terrain clearance indicators, airborne communications and military weapons systems.

W. B. NOTTINGHAM

For basic studies, teaching, and leadership in the field of physical electronics.

New Fellows



C. H. Page

For contributions to military electronic research and development.

W. Palmer

For contributions to the theory and practice of radio navigation.

S. E. Petrillo

For contributions in the development, production, supply, and quality control of military signal equipment.

W. J. Poch

For contributions in the development and design of television studio equipment.

J. Rabinow

For contributions in the fields of electronic ordnance and automatic control.

G. Rappaport

For contributions and leadership in electronic countermeasures research and development.

A. G. Richardson

For contributions and leadership in the development of radio direction finders.

Lothar Rohde

For contributions to electronic measurement techniques and for technical administration.

C. E. Scholz

For contributions to international radio communication.

New Fellows

J. L. Sheldon

For contributions in the field of glass technology of electron devices.

A. M. Skellett

For contributions to the physics of the upper atmosphere and to radial beam switching tubes.

J. J. Slattery

For contributions and administration in the radar and electronic countermeasure fields.

J. B. Smyth

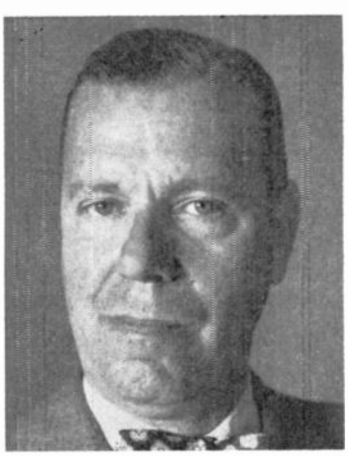
For contributions and leadership in electromagnetic propagation research.

H. A. Snow

For contributions to electronic circuitry, measurement techniques, and the development of the variable mu tube.

E. E. Spitzer

For contributions and leadership in the design of power tubes.



M. J. O. Strutt

For contributions to the knowledge of electron tubes and associated circuits, particularly at high frequencies.

C G. Suits

For leadership in industrial research.

G. K. Teal

For advances in electronics, particularly in the field of semiconductors.

New Fellows

W. A. Tolson

For invention and development in the field of television receivers and military target tracking.

A. van der Ziel

For research leadership and for studies of fluctuation phenomena in electron devices.

R. L. Wallace, Jr.

For contributions in the field of transistor technology and applications.

IRE News and Radio Notes

Committee Headed Symposium on Communication

Standing (left to right)—J. J. Renner, Technical Program; W. W. Balwanz, George Washington University representative; R. I. Cole, Finance; and C. E. McGinnis, Local Arrangements. *Seated (left to right)*—J. D. Wallace, Treasurer; E. N. Dingley, Jr., Advisory; K. S. Kelleher Chairman of the Symposium; and C. Goatley, Secretary. Principal symposium speakers were E. M. Webster and A. B. Dumont.

Scatter Techniques Symposium Held at George Washington University

The Professional Groups on Antennas and Propagation, and Communications Systems recently held a Symposium on Communication by Scatter Techniques in Washington, D.C.

The symposium was opened with a welcoming address by M. A. Mason, Dean of the School of Engineering at George Washington University. Commissioner E. M. Webster of the Federal Communications Commission, in the keynote address which followed, stressed the importance of scatter techniques in extending the frequency spectrum from which additional channels could be allocated to broadcasting stations.

Twenty-one papers were presented in the two-day symposium, and student members of the IRE chapter at George Washington University also participated.

At the end of the first day of the symposium, an informal dinner was held. A. B. Dumont delivered a dinner address to the members of the two Professional Groups and the Washington Section on recent developments and potential applications of scatter techniques in extending line-of-sight distances over which present television transmission systems now operate.

H. F. Mayer Feted by Rome-Utica Section

At the January recognition dinner honoring H. F. Mayer, recently elected an IRE Fellow, were *(left to right)*: M. V. Ratynski, dinner chairman; P. J. Schenk, General Electric Co.; H. F. Mayer, Rome-Utica Section Chairman and H. A. Carlson, Arrangements. The award went to a Section member for the second successive year.

Calender of Coming Events

Symposium on the Application of Ferrite Devices to Microwaves, Harvard Univ., Cambridge, Mass., Apr. 2–4

PGIE-AIEE-ISA Conference on Magnetic Amplifiers, Syracuse, N. Y., Apr. 5–6

Seventh Regional Technical Conference and Trade Show, Hotel Utah, Salt Lake City, Utah, Apr. 11–13

Tenth Annual Spring Television Conference of the Cincinnati Section, Engineering Society of Cincinnati Building, Cincinnati, Ohio, Apr. 13–14

International Electronics Conference and Show of the El Paso Section, El Paso Coliseum, El Paso, Tex., Apr. 13–15

National Industrial Research Conference, Hotel Sherman, Chicago, Ill., Apr. 18–19

New England Radio Engineering Meeting, Sheraton Plaza Hotel, Boston, Mass., Apr. 23–24

PGCT-PIB Symposium on Nonlinear Network Theory, Engineering Society Building, New York City, Apr. 25–27

Fourth Annual Semiconductor Symposium of Electrochemical Society, Mark Hopkins Hotel, San Francisco, Calif., Apr. 29–May 3

URSI Spring Meeting, National Bureau of Standards, Washington, D. C., Apr. 30–May 3

IRE-RETMA-AIEE-WCEMA Electronic Components Symposium, U. S. Department of Interior, Washington, D. C., May 1–3

National Aeronautical and Navigational Conference, Hotel Biltmore, Dayton, Ohio, May 14–16

Symposium on Reliable Applications of Electron Tubes, University of Pennsylvania, Philadelphia, Pa., May 21–22

Second Annual Radome Symposium, Ohio State Univ., Columbus, Ohio, June 4–6

National Telemetering Conference, Biltmore Hotel, Los Angeles, Calif., Aug. 20–21

IRE-West Coast Electronic Manufacturers' Association, WESCON, Biltmore Hotel, Los Angeles, Calif., Aug. 21–24

Boulder, Colorado Is Site of URSI Twelfth General Assembly, Aug. 22–Sept. 5

The International Scientific Radio Union will hold its Twelfth General Assembly at Boulder, Colorado, August 22-September 5, 1957. This will be the first Assembly held in the United States since 1927, and will be sponsored by the U. S. A. National Committee of URSI, the National Academy of Sciences, the University of Colorado, and the Boulder Laboratories of the National Bureau of Standards. J. H. Dellinger of Washington, D. C. is chairman of the General Arrangements Committee for this Assembly.

URSI is an international society which holds meetings, called General Assemblies, usually at three-year intervals, in different countries. These meetings, which last about two weeks, are devoted to papers and symposiums in which the latest developments in radio science are reviewed. URSI's research programs are carried out by its technical commissions on radio measurements and standards, radio and troposphere, ionospheric radio, radio noise of terrestrial origin, radio astronomy, radio waves and circuits, and radio electronics.

The first session of the program will pay special recognition to the beginning phase of the International Geophysical Year. Thirteen half-days , devoted to technical sessions, follow. Field trips will be taken to the Echo Lake Laboratories, the High Altitude Observatory, and the NBS Boulder Laboratories. There will also be an excursion to Rocky Mountain National Park and a ladies' program of activities.

Participation in the Assembly is open to representatives officially designated by the National Committees of the member countries belonging to the Union. Persons in any country who wish to attend should contact their National Committee during 1956.

URSI Convenes April 30–May 3

The URSI Spring Meeting is scheduled for the National Bureau of Standards in Washington, D.C. April 30–May 3. The cosponsors this year will include the IRE Professional Groups on Antennas and Propagation, and Microwave Theory and Techniques. A combined technical session of interest to all participants is scheduled for the morning of May 1, to be followed by one or more sessions in each of the following fields: *Commission 1*—On Radio Measurements and Standards; *Commission 2*—On Radio and Troposphere; *Commission 3*—Ionospheric Radio; *Commission 4*—On Radio Noise of Terrestrial Origin; *Commission 5*—On Radio Astronomy; *Commission 6*—On Radio Waves and Circuits; and *Commission 7*—On Radio Electronics.

The Commission and Professional Group representatives are as follows: *Commission 1 and PGMTT*—E. Weber, Head of Dept. of Electrical Engineering, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.; *Commission 2*—J. B. Smyth, Head of U. S. Navy Electronics Laboratory, San Diego, California; *Commission 3*—M. G. Morgan, Thayer School of Engineering, Dartmouth College, Hanover, N. H., *Commission 4*—A. W. Sullivan, Engineering and Industrial Experimental Station, University of Florida, Gainesville, Florida; *Commission 5*—F. T. Haddock, Naval Research Laboratory, Washington, D. C.; *Commission 6*—E. C. Jordan, Department of Electrical Engineering, University of Illinois, Urbana, Illinois; *Commission 7*—W. G. Shepherd, Department of Electrical Engineering, University of Minnesota, Minneapolis, Minn.; and *PGAP*—D. C. Ports, Jansky & Bailey, Washington 7, D. C.

Three-Day Simulation Conference Attracts 400

J. Tou, Univ. of Pennsylvania; C. C. Calvin, Chance Vought Aircraft, Inc.; and H. E. Blanton, Hycon Eastern, Inc. (*left to right*) were three of the thirty-five speakers at the National Simulation Conference held in Dallas recently. Over four hundred persons attended technical sessions and toured the Convair plant at Fort Worth.

Radio Club of America, Inc. Elects New Slate of Officers

The Radio Club of America, Inc. elected Frank A. Gunther, a Senior Member of the IRE, president of the organization for 1956. Serving with him are Walter A. Knoop, Jr., Vice-President; O. James Morelock, Corresponding Secretary; Joseph J. Stantley, Treasurer; and John H. Bose, Recording Secretary.

F. A. Gunther

Directors for the coming year include: Ernest V. Amy, Ralph R. Batcher, George E. Gurghard, Paul F. Godley, Harry W. Houck, Fred A. Klingenschmitt, O. F. Masin, Renville H. McMann, Jr., Jerry B. Minter, Harry Sadenwarer, Francis H. Shepard, Jr., Joseph J. Stantley, Jr., and Craig Walsh.

The Radio Club of America was organized in New York City on January 2, 1909, which makes it the oldest group of its kind in this country. Membership includes outstanding men in the field of radio engineering and invention both in this country and abroad. Its monthly meetings constitute a forum for new ideas. From time to time members whose achievements entitle them to that distinction are awarded the Armstrong Medal, established by the directors in 1935.

Mr. Gunther is vice-president of Radio Engineering Laboratories, Inc., in Long Island City, N. Y. Besides his business contact with radio for thirty years, he has been a licensed amateur with the call letters W2ALS since 1919. A retired Major, USAF Reserve, Mr. Gunther is a licensed commercial pilot and member of the Civil Air Patrol Reserve.

Walter A. Knoop is a professional engineer in the firm of Gawler-Knoop Co. O. James Morelock is a radio consultant in Millington, N. J. Joseph J. Stantley is with Continental Sales Co., Inc., of Newark, N. J. John H. Bose is engaged in research at the Electronics Research Laboratories of Columbia University.

Call for Papers Issued for PGED Technical Meeting

The Second Annual Technical Meeting of the IRE Professional Group on Electron Devices will be held October 25–26, 1956, at the Shoreham Hotel in Washington, D. C., it has been announced by T. M. Liimatainen of Diamond Ordnance Fuze Laboratories, general chairman of the meeting.

A call for titles and abstracts of 100–200 words on papers to be offered for presentation is being issued this month by R. L. Pritchard of the Research Laboratory, General Electric Co., Schenectady, N. Y., who is in charge of the meeting's technical program. The deadline for receiving titles and abstracts is August 1, but earlier submission of titles to facilitate planning is requested. Papers should deal with electron devices and be of applied or developmental nature.

Convention on Ferrites Set for October in London

The Institution of Electrical Engineers will sponsor a Convention on Ferrites during the week of October 29, in London, England. There will also be exhibits shown.

The tentative program has been divided into: *Session I*—Opening Session and Introductory Lecture; *Session II*—Theory, Preparation and Properties of Ferrites; *Session III*—Microwave Applications; *Session IV*—Square Loop Applications; *Session V*—Radio and Television Applications; and *Session VI*—Carrier Frequency Applications.

Further information may be obtained from W. K. Brasher, Secretary, Institution of Electrical Engineers, Savoy Place, London W.C. 2, London, England.

Washington Section Honors 22 at Annual February Banquet

More than four hundred members of the Washington D.C. Section of the IRE and their guests attended the annual banquet on February 11. The event marked the 42nd anniversary of the Section; it was chartered in 1914, just two years after the national organization was founded in New York. From an initial charter membership of 41, it has grown to a membership of nearly two thousand scientists.

E. M. Webster

For the occasion, the Washington Section created a special "citation for distinguished service." The first awards were made by the chairman, H. I. Metz of the Civil Aeronautics Administration, to the following past and present members of the Section: F. W. Albertson, senior partner of the law firm of Dow, Lohnes and Albertson; J. H. Dellinger, radio consultant and adviser; H. G. Dorsey, retired; F. P. Guthrie, assistant manager, Washington office, RCA Frequency Bureau; E. H. Rietzke, President, Capitol Radio Engineering Institute; A. H. Taylor, retired; J. D. Wallace, Naval Research Laboratory; E. M. Webster, Commissioner, Federal Communications Commission; and L. C. Young, Naval Research Laboratory.

Commissioner Webster also served as toastmaster for the banquet, at which the annual student awards were presented to four university seniors from the George Washington University and the University of Maryland, locations of the two student branches of the Section.

The winners and runners-up of the 1956 student awards were: D. B. Keever and H. K. Morlock, (runner-up) of George Washington University, and J. P. Sussman and D. R. Gouge, Jr. (runner-up), of the University of Maryland. W. R. Ferris, chairman of the Section's Special Affairs Committee, made the awards.

Fellow awards were presented by Stuart L. Bailey of the Washington Section, acting for IRE President A. V. Loughren, to the following: S. N. Alexander, National Bureau of Standards; G. F. Corcoran, Professor and Chairman, Electrical Engineering Department, University of Maryland; T. M. Davis, Naval Research Laboratory; E. N. Dingley, Jr., National Security Agency; H. P. Kalmus, Diamond Ordnance Fuze Laboratories; C. H. Page, National Bureau of Standards; J. Rabinow, President, Rainbow Engineering Company. The late Captain W. H. Beltz, USN, (ret.) and A. B. Clark were also made Fellows.

Mark Swanson, an ex-chairman of the Washington Section, and in charge of the arrangements for the banquet, discussed briefly some of the early history of the Section.

February Regional Conference Featured Tours

The IRE Southwestern Conference at Oklahoma City featured tours of Tinker and Will Rogers Fields, a medical research foundation, and television stations. Shown (*left to right*) are: H. V. Byler, Director, Civilian Training, Tinker Field, USAF; C. L. Farrar, Technical Program; C. E. Harp, Conference Chairman; C. O. Hart, Publicity.

Armour Research Foundation Held Automation Conference

A two-day conference was sponsored by Armour Research Foundation of Illinois Institute of Technology, and held at Chicago, Ill., Feb. 14–15. The conference, titled "Automation—A Conference for Executives," included case histories and discussions of sound automation programs.

The conference opened with welcoming remarks by H. A. Leedy, Armour Research Foundation director. A critical examination of automation as it exists today was the topic of a talk by J. R. Bright, Harvard University Graduate School of Business Administration. G. C. Ensign, director of research, Elgin National Watch Company, then presented a review of the developments which have led to the present surge in automatization. He described its dependence on traditional production engineering principles. Management's responsibility to weigh the various factors which influence decisions whether to introduce automation in operations, the effect on company earnings and progress, stockholders, and employees, was discussed by Arthur Studt, manager of education and training, Hotpoint Corporation.

The first day's afternoon session dealt with the essentials of a successful automation program.

Speakers included M. M. Lavin, senior engineer, Caywood-Schiller and Associates, who spoke on *Operations Research and Automation;* G. A. Nothmann, manager of mechanisms and dynamics research, Armour Research Foundation, who talked on *Design Tools for Automation*, and David Rubinfien, consultant, Arthur Young and Company, who addressed the conference on *A Survey of Clerical Automation in the Oil Industry.*

An address by D. J. Gimpel, research engineer in the electrical engineering research department of Armour Research Foundation, on the role of control and instrumentation in automation opened the second day's program.

The rest of the conference was devoted to automation case histories and included the following speakers and subjects: Wayne Larson, service engineer, Sciaky Brothers, *Welding of Automobile Wheels;* David Hinkel, assistant cashier, First National Bank of Chicago, *Automatic Processing of Checks;* Mark Morgan, project engineer, International Business Machines Corporation, *Digitally Controlled Cam Milling Machines;* and Martin Sheridan, public relations director, Admiral Corporation, *Automatic Assembly System for Electronic Components.*

WESCON Papers Deadline Set for April 15

Authors wishing to present papers at the 1956 WESCON Convention, to be held at Los Angeles on August 21–24, should send abstracts of their papers by April 15 to the alternate chairman, W. H. Ware, The Rand Corp., 1700 Main Street, Santa Monica, Calif., for evaluation by the technical program committee.

Authors will be notified whether or not their papers are accepted by May 15. Authors of accepted papers will be expected to submit a complete manuscript by July 1 so that preprints may be distributed at the meeting.

Third Information Theory Symposium Was Held at London

(Because of the international interest evoked by the London Symposium on Information Theory last fall and the wide range of topics covered, at the suggestion of the IRE Professional Group on Information Theory there is reprinted below from "Nature" an account of the meeting prepared by E. Colin Cherry of the Imperial College of Science and Technology, London.—The Editor)

The mathematical theory of information has undergone considerable broadening and clarification during the past few years, and it is being studied and applied to the design of experiments by people in very varied fields—in particular, communication engineering, taxonomy, linguistics and neurophysiology. Two symposia were previously held in London, in 1950 (*Nature*, 167, 20; 1951) and in 1952 (*Nature*, 170, 1051; 1952); and, as a further experiment, another interdisciplinary and international meeting was held recently at the Royal Institution during September 12–16, 1955. This meeting was attended by two hundred and fifty representatives from fifteen different countries, including a large body from the United States and representatives of the Academy of Sciences of the U.S.S.R. The meeting was organized by Colin Cherry, of the Electrical Engineering Department, Imperial College of Science and Technology, London, England.

The symposium was opened by Balth van der Pol of Geneva. From the start, the discussions were numerous and lively, showing that, in spite of their variety of background, the participants enjoyed a considerable degree of mutual understanding through their common study of information theory. The subject is, of course, mathematical and forms a part of scientific method; it does not relate solely to telecommunications (whence it arose) but concerns all phenomena which are *communicative* in nature, for which representations or scientific models are made, to which true measures can be applied precisely and numerically. The early promise offered, that information theory would provide a valuable unifying force between various specializations, is certainly being fulfilled.

The first day of the meeting was devoted to a review of the fundamental concepts and the various measures which may be applied to information rates. J. L. van Soest of the Netherlands discussed the essentially discrete or quantal nature of communicable information and some of its consequences to the theory. A paper by G. Spencer Brown on the difficulties underlying definition of *randomness* raised discussion to the level of debate—as might be expected of that thorny subject. But the concepts of *randomness* of a stochastic source and of the degree of *disorder* of finite sequence are of great importance to the whole theory; information theory may yet resolve some of the difficulties.

On the same day, Dr. Gabor, and W. Meyer-Epper of Germany extended the application of the theory to physics in particular, to optics. Finally, the relevance of *information* in the theory of games was considered by Dr. Farquharson of Great Britain.

On the second day, questions of coding of information-sources were discussed—the transformations of message representations from one form to another, and the possibility of improving rates of flow of information in noisy channels. A. E. Laemmel and P. Elias of the United States considered the theoretical bases of encoding so as to reduce the probability of error in, respectively, low and high noise-levels. Their colleague, D. A. Huffman, then showed ways in which such encodings might be carried out automatically in practice.

A related field of work was surveyed in the afternoon, which included the study of preferred ways of cataloguing, filing or classifying data in, for example, libraries—the theory of clerical operations. R. A. Fairthorne of Great Britain has worked in close association with Calvin N. Mooers of the United States, and gave a most humorous exposition of what a layman might guess to be a dry subject.

Real-life, human language was the topic on the third day, with accounts of some of the remarkable statistical facts about it. W. Fucks, of Aachen, Germany disclosed statistical data relating to nine different languages, ancient and modern, Eastern and Western, showing that they have certain forms in common. Prof. Fucks is well known for his claims of the authenticity of certain controversial classical texts by the use of discriminating-functions applied to the

Chicago Section Forms Publicity Committee of PG Publicity Chairmen

Standing, left to right—T. S. Pryst, PG on Audio; B. S. Schwartz, Section Chairman of Publicity; P. D. Huston, PG on Nuclear Science; R. M. Soria, Section Vice-Chairman; D. G. Haines, Section Secretary; K. Karrow, PG on Electronic Computers. *Seated, left to right*—H. C. Rett, PG on Communications Systems; J. S. Brown, Section Chairman; E. G. Book, PG on Antennas and Propagation; H. Brauer, PG on Engineering Management; J. Cunningham, PG on Broadcast and TV Receivers. The committee aims to better publicize the activities of the IRE and the electronics profession through newspapers, national magazines, and engineering periodicals.

words. A. S. C. Ross and D. A. Bell, of Birmingham, England, continued their earlier work (reported in the 1952 Symposium) of computing the degree of redundancy in a written language. This time they chose written Welsh because it is essentially a phonetic writing, and so their results bear at least some reasonable approximation to a spoken language (statistical data about spoken languages is extremely difficult to gather). B. Mandelbrot from Geneva and Grenoble continued the discussion of statistical laws which relate to language. There are close relations between statistical laws of language and other biological laws, in particular, those first exposed by J. C. Wills. The reasons for the similarity are by no means clear, but the suggestion was offered by Margaret Meade that it arises from the fact that the various descriptive categories (phonemes, say, in language, and species or genera in biology) are set up by groups of people in society. The laws may not be so much properties of nature, but of ourselves who set up descriptions and classifications.

The afternoon of this same day was devoted to mechanical translation. The principal contributors were A. D. Booth and J. Cleave, of Birkbeck College, London, England, V. H. Yngve, of the Massachusetts Institute of Technology, the United States, and S. Ceccato and E. Maretti from Italy. The distinction between translation and transcription (or encoding) was quickly drawn, and discussion largely concentrated upon the parts played by the syntactic and semantic rules in translation. The machines, as at present constructed, operate necessarily upon syntactic rules, and it is indeed surprising to what an extent successful conversions of a text from one language to another may be accomplished, even on a word-by-word conversion basis. The speakers emphasized the practical achievements of their work with examples of translation of scientific texts. The semantic rules may come in with the pre- and post-editing processes using reasonably intelligent human beings. As might be expected, it is scientific texts which offer most promise because they are to a large extent intercultural. The two Italian speakers gave a brief account of the approach of their operational school to problems of language and translation, which led to discussion of some of the philosophical difficulties and pitfalls.

During the afternoon, speakers from Britain, France and the United States outlined modern views upon the structure of physical speech signals, and upon aural and visual perception.

The final day was devoted to psychological and neurophysiological studies. Models representing certain aspects of brain functioning were discussed which, in their main attributes, are nowadays far removed from the digital computer analogies which so plagued us a few years ago. The stress today is upon self-organizing models in initial random connection. A paper by W. K. Taylor, of University College, London England, is likely to become a classic; he considered the properties essential to the elements of a mechanism which can *learn by association*, based upon known properties of neurons in living nerve cells and sense receptors—correct firing-rates, mutual interaction, facilitation and inhibition, accommodation—which determine the whole pattern of activity when associated. His analogue mechanisms are of the self-organizing type which, experimentally, exhibit primitive learning and adaptive behavior.

Later in the same day, some varied uses of information theory were demonstrated in human behavior studies, especially quantitative measures for experiments upon learning, motor skills, visual recognition and human operator capacity.

It was a happy sign of the times to hear, finally, a short report on information theory developments in the U.S.S.R. from Prof. Siforov, the leader of the Russian delegation. He brought news of a number of well-known workers in the field, including the pioneers Khinchin and Kolmogorov.

The *Proceedings* of the Third London Symposium on Information Theory are to be published by Butterworth's Scientific Publications, London, England.

Engstrom Wins Ericsson Medal

E. W. Engstrom, Senior Executive Vice-President of the Radio Corporation of America and an IRE Fellow, has been elected a Foreign Member of the Royal Swedish Academy of Engineering Sciences.

The announcement followed the presentation to Dr. Engstrom of the John Ericsson Medal by the American Society of Swedish Engineers at its annual dinner in New York on February 11. The medal was presented for "distinguished achievements in science and engineering." Established to honor the Swedish-American engineer among whose many outstanding developments was the "Monitor" of Civil War fame, the John Ericsson Medal is given at four-year intervals in this country to a person of Swedish descent. At intervening four-year intervals, it is awarded to a citizen of Sweden.

E. W. Alexanderson, Swedish-born pioneer in radio and electrical engineering, presented the medal, praising Dr. Engstrom for "ability and creativeness" and for his leadership of important team research in the field of electronics.

Obituary

Cornelius D. Ehret (F'16) died recently. He had been a patent lawyer in the field of radio from 1902 until his retirement in 1948. He was the holder of a degree in electrical engineering from Cornell University and the L.L.B. degree from the National Law School, Washington, D. C.

Professional Group News

Three New Chapters Approved

The Executive Committee, at its meeting of February 1, approved the formation of three new chapters. They are: PG on Communications Systems, Syracuse Section; PG on Aeronautical and Navigational Electronics, Akron Section; PG on Medical Electronics, Joint Metropolitan Area (New York, Northern New Jersey and Long Island Sections).

Technical Committee Notes

Chairman H. Jasik presided at a meeting of the **Antennas and Waveguides** Committee at IRE Headquarters on January 11. The definitions of *port* were discussed at some length. It was recognized that: (a) a port relates to a single waveguide mode; (b) port is equivalent to *terminal pair*; (c) a port has no particular physical configuration; and (d) the energy coupled through a port does not necessarily propagate beyond it. A tentative definition was prepared, and will be sent to the Circuits Committee requesting their comments. The committee reviewed the Proposed Standards on Measurements prepared by Subcommittee 2.4 on Waveguide and Waveguide Component Measurements.

The **Electron Tubes** Committee met at IRE Headquarters on February 10 with Chairman P. A. Redhead presiding. The Proposed Standards on Electron Tubes: Definitions of Non-Transit-Time Tubes was discussed, amended and unanimously approved. This proposed standard will be submitted shortly to the Definitions Coordinator for review and comment. The following proposed standards were reviewed, amended and referred back to the originating subcommittee for further work: Proposed Standard On Electron Tubes: Definition of Terms Related to Camera Tubes, and Proposed Standard on Electron Tubes: Physical Electronics Definitions.

Chairman K. R. McConnell presided at a meeting of the **Facsimile** Committee at the Times Building on January 13. The committee discussed the IRE Facsimile Test Chart, which should be ready shortly for distribution through the Radio-Electronics Television Manufacturers Association. The chairman reported that the Proposed Standards on Facsimile: Definitions of Terms has been approved by the Standards Committee. The committee made a final review of the approved Definitions of Terms. The next project that the committee will start is a review of the Facsimile Test Standards.

The **Radio Frequency Interference** Committee met at IRE Headquarters on February 6 with Chairman R. M. Showers presiding. The major portion of this meeting was devoted to the review of the supplement to IRE Standard 54 IRE 17.S1 prepared by Subcommittee 27.3 on Radio and Television Receivers. This supplement is to permit the extension of conductor interference measurements defined in Standard 54 IRE 17.S1 from a range of 300 to 10,000 kc to a range of 300 kc to 25 mc and replaces sections 3.2.2 and 3.2.5 of that standard. This proposed supplement was amended and approved, and will now be distributed to the chairmen of all technical committees and the chairmen of measurements subcommittees for their review and comment.

Chairman P. J. Herbst presided at a meeting of the **Radio Transmitters** Committee at IRE Headquarters on February 2. The committee decided that in the near future they will start working on a revision of the Transmitters Section of IRE Standards on Antennas, Modulation Systems and Transmitters: Definitions of Terms, 1948. H. R. Butler was appointed chairman of Subcommittee 15.1 on FM Broadcast Transmitters. Mr. Herbst announced that H. Goldberg had accepted the appointment as chairman of the Radio Transmitters Committee for the term May 1, 1956—April 30, 1957. Subcommittee 15.3 on Double Sideband AM Transmitters submitted a Proposed Standard on Double Sideband Transmitters: Methods of Test for the comments of the committee. This proposed standard will be reviewed at the next meeting of the main committee. A. Brown, Chairman of Subcommittee 15.5 on Single Sideband Transmitters, announced that a Proposed Standard on Power Output Measurement will be completed by the subcommittee in June, 1956. The formation of a new Subcommittee on Microwave Communication Systems will be discussed at the next meeting.

The **Standards** Committee met at IRE Headquarters on February 9 with Chairman E. Weber presiding. A. G. Jensen, Standards Coordinator, announced that a new policy had been instituted regarding proposed standards on definitions. In the future, preliminary definitions will be forwarded to ASA Sectional Committee C42 on Definitions of Electrical Terms for their review and comment before they are approved IRE Standards. The scope of the Piezoelectric Crystals Committee was amended to include Ferroelectricity. Due to the fact that foreign IRE members have been appointed recently to technical committees, the quorum rule governing technical Committees was amended to read as follows: "One-third of the number of members resident on the North American continent will constitute a quorum." At the request of the Facsimile Committee further consideration was given to the revisions made in the Proposed Standards on Facsimile: Definitions of Terms. After review this proposed standard was unanimously approved. The Proposed Standard on Symbols for Semiconductor Devices was discussed, amended and unanimously approved. The committee started review of the Proposed Standard on Methods of Testing Transistors, which will be continued at the next meeting.

Books

Electronic Motion Pictures by Albert Abramson

Published (1955) by University of California Press, Berkeley 4, California. 177 pages+6 page index+iv pages+8 page glossary+20 pages of notes. 92 figures. 9¼×6¼. $5.00.

This book traces the history of the motion picture camera through early film cameras and principles, and television cameras, to the present day image orthicon camera and color transmission.

The treatment is simplified, but thorough in a general way. The author has gone to considerable lengths to obtain a large selection of very interesting photographs of early and recent television equipment. It is not likely that there is much in this book that the average television engineer will not know already, but it is contained in a readily accessible form. Readers interested in obtaining a *general* impression of the development of the television camera will find this book worth reading, as will the student in a radio or television course. The footnotes and notes are copious and very comprehensive. Quite obviously the author has gone to a lot of work to compile this reference volume, but this reviewer does not feel that it was worth it from the point of view of the average television engineer.

J. H. Battison
KAVE, KAVE-TV
Carlsbad, New Mexico

Principles of Guided Missile Design: Guidance by A. S. Locke et al

Published (1955) by D. Van Nostrand Co., Inc., 257 Fourth Ave., N. Y. 3, N. Y. 713 pages+14 page index+xvii pages. Illus. 9¼×6¼. $12.50.

The book, written by a number of competent scientists chiefly associated with the U. S. Navy, represents an admirable compilation of fundamentals which must be thoroughly understood by the guided missile designer. The first two chapters are devoted to considerations of the guided missile as a weapon and a brief description of the portion of prior art from which security restrictions have been removed. The next five chapters treat the fundamentals of associated subjects such as navigation, radio propagation, infrared transmission, servo systems theory, mathematics and stability criteria. These subjects may be found in other texts, but are conveniently handled here with special orientation toward guided missile problems. The remaining thirteen chapters treat missile guidance and related subjects in varying detail and include system considerations, guidance components, classification of missile missions, prelaunch handling, target problems and even economic considerations.

This book is the first of a series on guided missile design. Five other volumes of the series are planned but are not yet written. The cover page indicates subjects to be treated in the entire series and not the contents of this first volume. Other subjects yet to be treated are problems of propulsion, boosters, combustion chamber materials, choice of fuels, range testing, external instrumentation, armament and operations research. It is hoped that the evaluation of target misses, analysis of errors and the location of malfunctions will be included.

The text will be of interest to people with engineering, physics and mathematics backgrounds who are engaged in guided missile development. It is not primarily designed for the undergraduate student, but may well be considered a text for a graduate course in the fundamentals of guided missile design.

C. H. Hoeppner
Stavid Engrg., Inc.
Plainfield, N. J.

Basic Processes of Gaseous Electronics by L. B. Loeb

Published (1955) U. of California Press, Berkeley 4, California. 966 pages+37 pages index+xii pages. Illus. 9¼×6¼. $13.50.

This very excellent and valuable book is well named. The basic processes covered are: ionic mobilities, the diffusion of carriers in gases, the velocities of electrons in gases, the distribution of energy of electrons, the formation of positive ions and of negative ions, and recombination of ions. It is not a treatise on gaseous conduction in the general sense of the term, because it deals only to a very limited extent with the gross aspects of gas discharges. For example it does not discuss the dimensions taken by an unconfined discharge, volt-ampere characteristics, or ambipolar diffusion except in a very limited way.

The book is a critical review of the correlation between theory and experiment regarding detail processes occurring in gas discharges. It represents in essence a permanent record of the life work of Dr. Loeb, who is probably the outstanding contributor to the scientific knowledge and understanding of gaseous conduction processes in the present generation. Particularly valuable is the emphasis on the techniques of experimentation, brought out in the comparative evaluation of experimental information obtained under various conditions and by various groups of workers throughout the world.

It does not contain to any considerable extent the detail derivations of the theoretical equations whose correlations with experiment comprise to a large degree the content of this book. There are extensive references to the original papers where the derivations are given, and to various review papers. In most cases the author gives rather detailed discussions of the hypotheses on which the theoretical expressions are based, and comparative criticisms of the logic and assumptions underlying conflicting theories, as obtained by various workers.

It is refreshing to find a book on this subject in which the emphasis is placed on techniques of instrumentation that have been developed since 1950. From this standpoint, there is some merit in reading the book backwards, that is, beginning with the last chapter, because it is in the last chapter or two that the dominating influence of recent instrumentation techniques becomes apparent, particularly pulse techniques. In this book, as in his earlier related one, *Fundamental Processes of Electrical Discharge in Gases* (published before the Second World War and now out of print), Dr. Loeb has a great deal to say regarding misinterpretations given to work by earlier experimenters, because of the inadequacies of their vacuum and instrumentation techniques.

The book somewhat belabors the historical approach. To an engineer whose professional needs call for a full scientific understanding of basic gaseous electronic processes it is the present state of the art that is important; when and how the knowledge has been obtained is of little consequence except as necessary to permit proper evaluation of items of conflicting information. In this respect the present work is an improvement over its predecessor, in that less emphasis is placed on early concepts that have since been discarded. However, the author still goes through the laborious operation of deriving voltage breakdown relations on the basis of ionization by ionic collision, then showing this process to be nearly nonexistent.

The major grouping of the subject matter according to various detail processes is satisfactory and convenient as far as the reader is concerned. Within each chapter, the book tends to consist of running commentaries on work done by various individuals and groups, always with the expectation that the reader will refer to the original papers for ultimate information. Thus the book is not conveniently arranged as a direct source of quantitative information. Actually, much directly useful information is contained in the curves and tables presented, but they are arranged in the order called for by the need for correlation with theory, rather than in a pattern of probable value to the reader.

However, the overriding point here is that this particular type of treatment is the one that Dr. Loeb has been in position to give, that represents as complete as possible a record of his knowledge and understanding of the field. A more completely digested treatment of the subject, better organized from the standpoint of utility for problem-solving purposes, can perhaps be the subject of someone else's book. If Dr. Loeb had given more time to this aspect of organization, he would have given less time to the other aspects which he is peculiarly well qualified to handle.

It is important to start any use of this book by reading the preface, to gain a proper perspective on what the author is trying to do. Probably the best review comments that can be made are the following quotations from the preface: "obviously, the problem [of the present need] could not have been solved by writing a new edition of the earlier work The old book laid many ghosts of the past which require no comment today. Furthermore, the new material is so extensive that it must be approached *de novo* in many instances. In order to be of value, the present work must serve not only for use in instruction in the subject, but, even more, for the benefit of the engineer or physicist who is not conversant with the field of gaseous electronics, there must be a simplified, phenomenological or kinetic theoretical introduction to each topic, as there was in the earlier book. Then on the basis of experimental findings and later theoretical development, there will be presented the more esoteric advances. Obviously, it is impossible to include in detail the elaborate mathematical physical analyses, but enough of the analytical approach can be given to indicate the physics underlying the main trend of the study, together with its logical conclusions."

In summary, Dr. Loeb's book is an extremely valuable and important addition to the literature of applied physics, of a type that will be of increasing value to the electrical engineering profession in the future.

W. G. Dow
Univ. of Michigan
Ann Arbor, Mich.

Transistors and Other Crystal Valves by T. R. Scott

Published (1955) by Essential Books, Inc., Fairlawn, New Jersey. 236 pages+4 page index+17 page bibliography+xvi pages. 65 figures. 8¾×5¾. $7.20.

This is an excellent book for electrical engineers and for students who want an understanding of transistors and semiconductor diodes without becoming experts in quantum mechanics, crystallography and metallurgy. The material is largely descriptive and results rather than derivations of theory are used. After a brief historical introduction the author treats the basic characteristics of crystals and *p-n* junctions. From these he goes on to describe junction transistors, point contact devices, circuit applications, and special high frequency transistors. Throughout the book there are copious references to original works. Many of the diagrams are reproduced from original articles. There is an excellent collection of references at the back of the book, though as a bibliography it is far from complete.

The book is meant to be timely, and recognition must be made of the fast pace in the semiconductor field. Nevertheless there are some curious gaps in the book, as parts of the manuscript were apparently prepared in 1952, other parts in 1954, while the book itself was published in 1955. For example, on page 90 reference is made that as of May, 1952 only one company was making junction transistors and these at a rate of less than 100 a month. Nowhere in the book is it pointed out that by June, 1954 one company alone had manufactured and sold over one million junction transistors. In attempting to be critical of the available products the book necessarily dates itself.

There are a number of typographical errors in the book which one can assume will be corrected in future printings. The reader will also be somewhat distraught over the small size of many of the diagrams. From the standpoint of content the treatment on surface effects is negligible. The author recognizes that the problem exists and brushes over the solution as essentially trade secrets. There is only casual mention of the problem of noise to the point that noise is relegated to a future volume to be produced by the author. The circuit treatment is on the whole perfunctory.

In spite of the omissions the book is quite useful considering its size. It is easy to read and will undoubtedly find a place among the current books on transistors.

I. A. Getting
Raytheon Mfg. Co.
Waltham, Mass.

Noise by Albert van der Ziel

Published (1954) by Prentice-Hall Inc., 70 Fifth Ave., N.Y. 11, N.Y. 443 pages+6 page index+xi pages. 97 figures. 8¾×5¾. $10.35.

This book is a theoretical text on noise in electronic circuits. In general, it reduces the solutions of most noise problems to an analysis of simple networks.

The book begins with a consideration of basic concepts of thermal noise, noise in electrical circuits and components, noise figure, and noise measurements. Tube noise at low and high frequencies is then analyzed in detail, and practical low-noise circuits are considered in terms of gain, bandwidth, and noise figure. The circuits include grounded cathode, grounded grid, grounded anode, cascode, cathode coupled, distributed amplifiers. A general treatment is then given of thermal, shot, and excess noise in semiconductors, with application to crystal diodes, transistors, and photoconductive cells. Noise in tube and crystal diode mixers and mixer circuits, and noise in various forms of feedback circuits are also considered. The book concludes with a rather complete mathematical treatment of statistical methods, Fourier analysis of fluctuating quantities, noise in linear, quadratic, and phase sensitive detector circuits, vacuum tube electronics, and space-charge waves in electron beams.

Since the author's objective is to make a rather complete analysis of the noise problem in electronic circuits, the treatment is necessarily a mathematical one, and the book is a text, not a handbook. It could be used at the college senior level, but it will serve its purpose best in a graduate course on noise. The practicing engineer will find it useful as a reference for the theoretical treatment of almost any noise problem, particularly at low or medium frequencies, in which he is interested. Unfortunately, with the exception of Chapter 7 on low-noise circuits, there are few illustrative examples, nor is there a sufficient number of charts and graphs illustrating the results of the mathematical treatment, and this will hinder the electronic engineer in his use of the book as a reference.

The treatment of noise in microwave circuits is not entirely up-to-date due to the fact that almost all references cited are for 1952 or earlier. This applies to the discussion on crystal diode mixer noise, for example, where the work of Strum and others has not been included. The noise figure of a representative microwave crystal mixer is given as 12 db, which is high for present-day mixers. Local oscillator noise gets a very brief treatment, and although klystron oscillator noise is described as large, no data are presented. The treatment of the fluorescent lamp as a noise source at microwaves includes the original Mumford data on the temperature dependence on the excess noise (−.055 db per degree centrigrade above 40°C). Mumford and Schafersman have shown recently that this relationship is not always valid, even for a lamp that once obeyed it. All of these data have been made available only in the last year or two, and thus too late for the book. The microwave engineer should use the book with this in mind.

Despite these reservations, the book is an authoritative contribution to the literature on the subject of noise, particularly, as indicated, in low or medium frequency circuits. It is recommended as a text for the student or engineer in the field interested in a comprehensive coverage of the theory of noise in such circuits.

W. E. Fromm
Airborne Instruments Laboratory, Inc.
Mineola, New York

Recent Books

Hoelscher, R. P. and Springer, C. H., *Engineering Drawing and Geometry*. John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. $8.00.

Johnson, J. R., *Television, How It Works*, 2nd ed. John F. Rider Publisher, 480 Canal St., N. Y. 13, N. Y. $4.60.

Kopal, Zdenek, *Numerical Analysis*. John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $12.00.

Platt, Sidney, *TV Repair Questions and Answers—Deflection and H-V Circuits*. John F. Rider Publisher, 480 Canal St., N. Y. 13, N. Y. $2.10.

Proceedings of the RETMA Symposium on Automation. Engineering Publishers, GPO Box 1151, N. Y., N. Y. $5.00. Contains all material presented at the symposium held September 26–27, 1955 at the University of Pennsylvania, Philadelphia, Pennsylvania.

Schure, Alexander, *A-M Detectors*. John F. Rider Publisher, 480 Canal St., N. Y. 13, N. Y. $1.25.

Schure, Alexander, *Limiters and Clippers*. John F. Rider Publisher, 480 Canal St., N. Y. 13, N. Y. $1.25.

Swaluw, H. L. and van der Woerd, J., *Introduction to TV Servicing (For 525 and 625 Line Receivers)*. Elsevier Press, Inc., 2330 Holcombe Blvd., Houston 25, Texas. $5.50.

The 400 American Standards in the Electrical Field. American Standards Association, 70 E. 45th St., N. Y. 17, N. Y. Gratis.

U.R.S.I. Proceedings of the XI General Assembly: Vol. Ten, Part Five. General Secretariat of U.R.S.I., 42 Rue des Minimes, Brussels, Belgium. $2.50.

U.R.S.I. Proceedings of the XI General Assembly: Vol. Ten, Part Six. General Secretariat of U.R.S.I., 42 Rue des Minimes, Brussels, Belgium. $3.00.

White, H. E., *Modern College Physics*, 3rd ed. D. Van Nostrand Co., 257 Fourth Ave., N. Y. 10, N.Y. $6.75.

1956 IRE CONVENTION RECORD

All available papers presented at the 1956 IRE National Convention will appear in the IRE Convention Record to be published in July. The Convention Record will be issued in nine Parts, with each Part devoted to related subjects. The papers for each session are listed on pages 384–417 of the March issue.

Instructions on Ordering

1. If you are a member of a Professional Group and have paid the group assessment by April 30, you will automatically receive, free of charge, that Part of the Convention Record pertaining to the field of interest of your group, as indicated in the chart below.

2. If you are not a member of an IRE Professional Group, Convention Record, Parts may be purchased at the prices listed in the chart below. Orders must be accompanied by remittance, and to assure prompt delivery, should be sent immediately to The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y.

Convention Record

Part	Title	Free to Paid Members of Following Professional Groups	Prices for Members (M) Colleges and Libraries (L) Non-Members (NM)		
			M	L	NM
1	Telemetry, Antennas & Propagation **Sessions: 5, 14, 22, 24, 28, 33, 38, 40**	Antennas & Propagation Telemetry and Remote Control	$3.00	$7.20	$ 9.00
2	Circuit Theory **Sessions: 30, 41, 49**	Circuit Theory	1.25	3.00	3.75
3	Electron Devices & Receivers **Sessions: 16, 23, 29, 37, 43, 50**	Broadcast & Television Receivers Electron Devices	2.50	6.00	7.50
4	Computers, Information Theory, Automatic Control **Sessions: 7, 10, 32, 39, 42, 46, 53**	Automatic Control Electronic Computers Information Theory	3.50	8.40	10.50
5	Microwaves & Instrumentation **Sessions: 1, 26, 34, 47, 48, 54**	Instrumentation Microwave Theory & Techniques	2.75	6.60	8.25
6	Manufacturing Electronics **Sessions: 6, 8, 17, 27, 35, 44, 45, 52**	Component Parts Engineering Management Industrial Electronics Production Techniques Reliability & Quality Control	3.25	7.80	9.75
7	Audio & Broadcast **Sessions: 12, 13, 20, 21, 25, 55**	Audio Broadcast Transmission Systems	2.25	5.40	6.75
8	Aeronautical, Communications & Military Electronics **Sessions: 3, 4, 11, 15, 19, 31, 36**	Aeronautical & Navigational Electronics Communications Systems Military Electronics Vehicular Communications	2.75	6.60	8.25
9	Ultrasonics, Medical & Nuclear Electronics **Sessions: 2, 9, 18, 51**	Medical Electronics Nuclear Science Ultrasonics Engineering	1.50	3.60	4.50
	Complete Convention Record (All Nine Parts)		$22.75	$54.60	$68.25

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 & Navigational Electronics	Vol. ANE-2, No. 4	$1.40	$2.10	$4.20
Circuit Theory	Vol. CT-2, No. 4	1.85	2.75	5.55
Electronic Computers	Vol. EC-4, No. 4	.90	1.35	2.70
Engineering Management	Vol. EM-3, No. 1	.95	1.40	2.85
Information Theory	Vol. IT-1, No. 3	1.55	2.30	4.65
Microwave Theory and Techniques	Vol. MTT-4, No. 1	1.65	2.45	4.95

*** Public libraries and colleges may purchase copies at IRE Member rates.**

Circuit Theory

Vol. CT-2, No. 4, December, 1955

Abstracts of Papers in This Issue

The Papers in This Issue—S. Darlington

A Survey of Network Realization Techniques—Sidney Darlington

This paper is a brief survey of network realization techniques, from a general point of view. By realization techniques is meant methods of finding explicit networks (configurations and element values) from prescribed characteristics (such as impedance functions) which can be realized exactly. The statement of conditions under which prescribed characteristics are realizable is included in the definition.

Important ingredients of realization techniques are first discussed. Then various past trends in the development of realization techniques are noted. Finally, some as yet unsolved problems are formulated.

A New System of Two-Terminal Synthesis—Fusachika Miyata

This paper describes methods of synthesizing prescribed two-terminal impedances, of the RLC sort. While these methods are not entirely general, when they do apply they lead to an economy in elements, relative to the Bott and Duffin method, without requiring the mutual inductances of the Brune method.

Particular attention is paid to the even part of the impedance function. The even part determines the entire function uniquely, except for a possible added term representing the impedance of a network of reactances only. For purposes of network synthesis, the even part of the prescribed function is decomposed into a sum of parts, such that each part may be realized in a relatively simple way. The simplicity of the realization is achieved by suitably locating the zeros of the real parts of the partial impedances.

Special Synthesis Techniques for Driving Point Impedance Functions—E. S. Kuh

An important problem in network design is the synthesis of driving-point impedance functions. As is well known, O. Brune was the first to state the necessary and sufficient conditions for physical realizability. Unfortunately, the synthesis technique which he proposed leads in general to perfectly coupled transformers. This is true also in the case of the contributions made later by S. Darlington. Perfect transformers were eliminated by R. Bott and R. J. Duffin. However, their solution is, in general, expensive in terms of the number of elements that are required. Since the publication of their letter, many attempts have been made to find a solution that would lead to networks containing a number of elements closer to the minimum specified by Brune. An advance in this direction has been made by F. Miyata for a restricted class of positive real functions. He has centered attention on the even part of the impedance function. The following paper exploits this point of view and amplifies some of the ideas given by Miyata. In addition, several new ideas are described relative to methods of decomposing the even part of the impedance function in such a way as to obtain a network without perfect transformers.

Synthesis of Three-Terminal RC Networks—H. M. Lucal

Several methods are available for synthesizing three-terminal RC networks when only two of the three short-circuit admittances or two of the three open-circuit impedances are specified. This paper treats the synthesis problem in which all three functions are given; that is, in which terminal behavior of network is entirely prescribed.

There is a discussion of necessary conditions which must be satisfied by a set of short-circuit admittances or open-circuit impedances in order that it correspond to a three-terminal RC network. A synthesis method is then developed by means of which networks can be found for a wide variety of specified functions.

By means of well-known correspondences between networks of different kinds of elements, similar synthesis methods are easily derived for LC and RL network, without mutual inductances.

The Concept of the One in Voltage Transfer Synthesis—P. M. Lewis II

The concept of the one is introduced and found to give a physical interpretation to the constant multiplier in voltage transfer synthesis. It places in evidence some constraints placed on the performance of physical systems by the configuration used in the synthesis. Several theorems are derived concerning the ultimate limits of performance obtainable when a given configuration is used to realize a given voltage transfer function.

Realizability Theorem for Mid-Series or Mid-Shunt Low-Pass Ladders Without Mutual Induction—T. Fujisawa

This paper treats the synthesis of nondissipative low-pass ladders without mutual induction. Necessary and sufficient conditions for physical realizability of mid-series or mid-shunt nondissipative low-pass ladders without mutual induction are presented here and the complete proof is given. The essence of the conditions imposed upon the input impedance of the network terminated in a pure resistance is a relation between the finite frequencies at which the loss is infinite and the roots of a polynomial which appears in the rational fractional representation of the input impedance. The method used is the ladder development of a two-terminal impedance which is elementary in network synthesis. Furthermore, a sufficient condition for physical realizability of general low-pass ladders without mutual induction, which is wider than the above conditions, is also given.

This paper is concerned with the realization problem of a dissipative two-terminal impedance by means of a nondissipative two-terminal-pair network without mutual induction terminated in a pure resistance. The above-mentioned conditions are also necessary and sufficient in order that a dissipative two-terminal impedance can be realized by means of a mid-series or mud-shunt nondissipative low-pass ladder without mutual induction terminated in a pure resistance.

Synthesis of Transmission-Line Networks and the Design of UHF Filters—H. Ozaki and J. Ishii

Recently, Richards proved the remarkable fact that the synthesis of networks consisting of resistors and lossless transmission-lines with finite electrical length may be treated exactly the same way as in the case of conventional linear passive networks. This paper offers a new synthesis of this kind without mutual induction and its application to the design of uhf filters having Tchebyscheff characteristics in both pass and attenuation bands. The synthesis problem treated here is, in other words, the analog of the so-called Darlington problem in conventional lumped constant networks. Approximation procedures for filter characteristics are based on the well-known potential analogy.

Topics in the Design of Insertion Loss Filters—V. Belevitch

This paper combines the theory of scattering matrices and the design of filters on an insertion loss basis. It is not intended to add substantially new results. Its purposes are to simplify the presentation of the insertion-loss theory, with the help of the scattering parameters, and to study a number of particular cases serving as a guide in practical design and as a starting point of approximate formulas for numerical work. The particular cases and approximate formulas give the designer an appreciation of the influence of various parameters, before the design is crystallized and the accurate computations are begun.

The first part of the paper develops properties of general reactance networks, terminated in resistances, and of the special class of such networks which are either "symmetric" or "antimetric." Then the general theory is applied to the design of filters with Tchebycheff behavior in the pass band.

A New Approach to the Problem of Cascade Synthesis—E. A. Guillemin

Several familiar procedures in the synthesis of driving-point and transfer impedances yield a network in the form of a cascade of two-terminal-pair networks, each of which is responsible for a set of zeros of the even part of the prescribed impedance. In each of these procedures, the computation of the typical component network begins with the determination of the equivalent tee-circuit; and then a remainder function (from which other component networks are to be found) is determined by subtracting the tee's branch impedances (or admittances) from previously determined functions. In general an equivalent circuit of some other configuration must be determined from the tee-circuit, to achieve physical realizability.

These computational procedures are rather laborious, and the subtractions may cause serious degradations of numerical accuracy.

In this paper, a more direct approach is suggested. It leads to the same ultimate networks in a more logical way, and obviates much of the computational tedium. The general approach is developed here primarily in terms of RC transfer impedances, but applications to LC networks are also included.

Topological Considerations in the Design of Driving-Point Functions—S. Seshu

As contrasted with the conventional discussions of network synthesis which concern themselves with the analytic character of the network functions, this paper presents the algebraic topological considerations in the realization of the driving point functions. Abstractly a network may be considered as a linear graph with certain "weights" attached to the elements of the graph. These weights may be either impedance or admittance functions of the elements. If the network contains no mutual inductances, the network functions are expressible simply in terms of the topology of this weighted graph. On the other hand if the function is specified in terms of these weights it should be possible to deduce the topology. The paper begins with the assumption that the given driving point admittance function $Y_d(s)$ is expressed as a ratio of homogeneous polynomials $V(y_i)/W(y_i)$ in the weights or the elementary admittance functions of resistors, inductors and capacitors. Just how the function $Y_d(s)$ is to be expressed as $V(y_i)/W(y_i)$ is left as an unsolved problem. The properties of the polynomials $V(y_i)$ and $W(y_i)$ with regard to their realizability are the main results of the paper. A procedure for exhibiting the network from the realizable functions $V(y_i)$ and $W(y_i)$ is also given. Relationships between the vertex and circuit matrices of the network and the polynomials $V(y_i)$ and $W(y_i)$ are derived for this purpose. A generalization of Brune's theorem I and a sufficient criterion for a matrix to be a circuit matrix are two auxiliary results obtained.

The value of the paper is primarily academic—in providing results which are useful in a discussion of basic ideas and in suggesting problems in the rich and unexplored field of topological methods of network synthesis.

On the Response of Linear Systems to Signals Modulated in Both Amplitude and Frequency—F. Zweig, P. M. Schultheiss and C. A. Wogrin

The treatment by Carson and Fry of linear circuit responses to frequency modulated signals is extended to include the case of simultaneous amplitude and frequency modulation of a carrier wave by two arbitrary continuous signals. The response of a network to such signals is given in terms of an envelope function multiplying the unmodulated carrier.

Reviews of Current Literature Correspondence

PGCT Activities

Electronic Computers

Vol. EC-4, No. 4, December, 1955

Fast Carry Logic for Digital Computers—Bruce Gilchrist, J. H. Pomerene, and S. Y. Wong

Existing large scale binary computers typically must allow for the maximum full length carry time in each addition. It has been shown that average carry sequences are significantly shorter than this maximum, on the average only five stages for a 40 digit addition. A method is described to realize the implied 8 to 1 time saving by deriving an actual "carry completion" signal. Experimental results verify this saving.

Bit Storage via Electro-Optical Feedback—Alfred Milch

An electro-optical binary storage device has been built which consists of a vacuum diode containing a photocathode and a phosphor-anode. The device is capable of storing both optical and electrical information pulses. The present paper comprises a description of the behavior and construction of the prototype diode, an empirical derivation of a criterion for the conditions of stable feedback, and a numerical calculation for the case of two electrode pairs. Briefly mentioned are the possibility of high speed storage of digital information, the problem of self-triggering, the applicability of the criterion to solid state devices, and the analogous bistable behavior of a radio frequency triggered neon diode.

Ternary Counters—R. S. Mackay and R. MacIntyre

Counter stages having three stable states, and using no more components than standard binary counters, have been built by properly using the already present nonlinearity of grid current. Problems of stability, cascading, and decoding have been worked out.

A Logarithmic Voltage Quantizer—E. M. Glaser and H. Blasbalg

This paper describes an analog to digital converter which converts voltage into a number which is proportional to the logarithm of the voltage. The device is completely automatic. It can handle input data at the rate of 10,000 voltage samples per second. The accuracy of conversion or quantization is determined by the designer's selection of the circuit parameters. The Radiation Laboratory quantizer has a maximum quantization error of 5 per cent. Samples of duration greater than .5 microsecond can be quantized.

High Density Williams Storage—S. Y. Wong

In this study, an investigation was made to discover methods of storing more bits on a Williams tube than can be stored by conventional methods. Read-around was ignored because the object was to explore the Williams tube as a secondary memory for non-random access operations. Proceeding on this basis, it was found that a four-fold increase in packing is possible with established techniques, and an even greater increase with other methods described in this paper. Such a memory is more versatile than a magnetic drum, as shown by the example of machine organization presented in this paper.

Contributors

Correspondence

PGEC News

Reviews of Current Literature

Engineering Management

Vol. EM-3, No. 1, January, 1956

The Problems of Engineering Management—L. J. Fogel

Small Engineering Company Organization: A Philosophy and Method—T. W. Jarmie

Market Development—The Neglected Companion of Product Development—A. D. Ehrenfried

Management Techniques and Controls for Engineering Writing Organizations—Irving Hirsch

Methodology for Reliable Failure Reporting from Maintenance Personnel—F. A. Hadden and L. W. Sepmeyer

Correspondence

Information Theory

Vol. IT-1, No. 3, December, 1955

L. A. De Rosa

In Which Fields Do We Graze?—L. A. De Rosa

Theory of Noise in a Correlation Detector—M. Horowitz and A. A. Johnson

The problem of detecting a signal that has both magnitude and sign is considered. A new type of correlation device, which employes the derivative of the correlation function, is proposed for measuring target range and position, and an analysis to determine the most useful waveform is made. The effects of white noise accompanying the input signal are minimized when an input waveform of long duration, wide bandwidth, and high derivative power is chosen.

Optimum Sequential Detection of Signals in Noise—J. J. Bussgang and D. Middleton

A device which performs a sequential test on a mixture of signal and noise is called a *Sequential Detector*. With such a device, two thresholds are introduced, each of which is associated with a terminal decision. The length of the detection process (integration time) is not fixed in advance of the experiment but is a random variable, depending on the progress of the test. An optimum form of such a test exists and is characterized by the fact that detection is performed *on the average* faster than with conventional; *i.e.*, fixed sample size (optimum or non-optimum), devices. The sequential analysis developed by A. Wald is fully applied in this paper, but an important new feature is the treatment of correlated samples and its application to continuous sampling processes.

In the introduction, the problem is presented within the framework of Wald's Statistical Decision Theory, and the optimum properties of sequential detectors are discussed accordingly. It is pointed out that a sequential detector is defined in terms of *conditional* probabilities and hence its operation is essentially independent of *a priori* information, although the average risk or cost of detection necessarily depends on the *a priori* signal data. The general theory is illustrated with some cases of special interest.

The simplest example of detection involves independent, discrete observations; *e.g.*, the case of a pulsed carrier in normal noise. Here the optimum detector still has the well-known $\log I_0$ structure, but it is shown that the square law approximation for weak signals requires a bias correction due to the fourth order term. Coherent sequential detection of causal signals in normal noise provides another illustration of the theory. An interesting result is that the probabilities of error do not depend on the shape of the filter, provided the proper computer is used. The use of RC-filtered noise illustrates the treatment of continuous detection processes. Finally, the reduction in minimum detectable signal level resulting from the use of a sequential detector is computed. A third example is the sequential detection of random signals in normal noise. It is shown that, although the optimum computer involves the knowledge of the inverted correlation matrix, the average length of the test does not. Hence a curious result is obtained that in this instance detection can be performed in an arbitrarily short time. The paper concludes with a discussion of the practical necessity of truncating the detection process and exact expressions for the error probabilities of such truncated tests are derived and compared with Wald's original approximations.

On Binary Channels and Their Cascades—R. A. Silverman

A detailed analysis of the general binary channel is given, with special reference to capacity (both separately and in cascade), input and output symbol distributions, and probability of error. The infinite number of binary channels with the same capacity lie on double-branched equicapacity lines. Of the channels on the lower branch of a given equicapacity line, the symmetric channel has the smallest probability of error and the largest capacity in cascade, unless the capacity is small, in which case the asymmetric channel (with one noiseless symbol) has the smallest probability of error and the largest capacity in cascade. By simply reversing the designation of the output (or input) symbols, we can decrease the probability of error of any channel on the upper branch of the equicapacity line and increase the capacity in cascade of any asymmetric channel on the upper branch.

In a binary channel neither symbol should be transmitted with a probability lying outside the interval $[1/e, 1-(1/e)]$ if capacity is to be achieved. The maximally asymmetric input symbol distributions are approached by certain low-capacity channels. For these channels, redundancy coding permits an appreciable fraction of capacity in cascade if sufficient delay can be tolerated.

Minimum Energy Cost of An Observation—F. P. Adler

The minimum energy expenditure required in performing basic observations and measurements is analyzed. The energy cost, in ergs per binary unit (bit) of information, is found for three fundamental cases using idealized experimental procedures: 1) the determination of the presence (or absence) of an input signal on an indicating instrument, 2) the measurement of a time interval and 3) the measurement of a distance. The variation of energy cost with the reliability and accuracy of the experiment is determined; it is found that with a suitable procedure the minimum value of kT ln2 ergs per bit predicted by the Second Law (interpreted so as to include informational entropy) can be approached arbitrarily closely under conditions of small reliability and high accuracy. The present results are compared with those derivable from C. E. Shannon's equation for the capacity of a communication channel.

Some Remarks on Statistical Detection—W. L. Root and T. S. Pitcher

A particular type of communications detection problem is considered: the problem of specifying a detector to decide which one of two sure signals is being transmitted when the signals are perturbed randomly both by Gaussian noise and multipath transmission. If the delays in the various channels are known, but the strengths are random, a maximum likelihood detector may be specified by methods which are a simple extension of known methods. If the delays are random, the problem is more difficult. One possible solution is first to estimate certain channel parameters from the received signal and then to use these estimates in a likelihood test. It is shown how to make consistent unbiased estimates for appropriate channel parameters under certain assumptions on the nature of the signal.

Correction

Contributors

IRE Professional Group on Information Theory Membership Directory

Microwave Theory and Techniques

Vol. MTT-4, No. 1, January, 1956

Message from the Editor

A. G. Clavier

"Wire" vs "Wireless" Communication—A. G. Clavier

The Characteristic Impedance of a Slotted Coaxial Line—R. E. Collin

The propagation of a second type of TEM mode in a slotted coaxial line is analyzed. The characteristic impedance of the slotted line is evaluated by means of variational expressions giving upper and lower bounds to the true value. A two term approximation to the charge distribution and a one term approximation to potential distribution give results accurate to within ±2 per cent. Curves of characteristic impedance against angular slot width are presented.

Broadband Ferrite Microwave Isolator—P. H. Vartanian, J. L. Melchor, and W. P. Ayres

A new type broadband transmission line has been built utilizing the difference in energy distribution between two counter-rotating circularly polarized waves in a circular waveguide containing a ferrite. This principle of isolation is different from those which have been used previously.

A large difference is observed in the energy distribution of two counter-rotating TE_{11} modes in a ferrite loaded circular waveguide. A ferrite rod magnetized along its axis presents an effective rf permeability of approximately two for the mode rotating in a negative screw sense with respect to the direction of magnetization. For the positive sense of rotation the effective rf permeability becomes very small and negligible energy is transmitted through the ferrite rod.

Unidirectional transmission characteristics were achieved by adding quarter wave plates before and after the ferrite rod and inserting an absorber into the ferrite. For the direction of propagation for which the quarter wave plate converts from a linear input to a positive circular rotation the positive wave tends to go around the ferrite with small loss. For the other direction of propagation the quarter wave plate converts the linear input wave to a negative wave which tends to concentrate in the ferrite and is absorbed.

Based on the principles described, an isolator was constructed which gives better than 30 db isolation over the range 8 to 11 kmc. The insertion loss is less than 2 db from 8 to 10.5 kmc and increases to 3 db at 11 kmc. The complete unit is $10\frac{1}{2}$ inches long and weighs $2\frac{1}{4}$ pounds.

The main advantage of this isolator over present transverse field rectangular waveguide isolators and Faraday rotation isolators is its improved bandwidth. Other advantages are that the isolator is not sensitive to changes in magnetic field and it operates with a readily obtainable ferrite at low magnetic fields. Its vswr over the band is less than 1.2. The principle of this isolator is applicable to other frequency bands.

An Approximate Analysis of Coaxial Line with a Helical Dielectric Support—J. W. E. Griemsmann

Improved Microwave Noise Measurements Using Ferrites—C. H. Mayer

The ferrite isolator and the ferrite circulator have been applied separately to improve the accuracy of measuring small microwave noise powers or small power differences. Either the isolator or the circulator effectively isolated the input circuit of a microwave receiver from the impedance of the source. As a result, the measurement errors introduced by mismatched source impedances were reduced by as much as 98 per cent. The added input circuit losses of the ferrite components reduced the receiver sensitivity by only about 10 per cent. Since the accuracy of measuring small noise power differences was limited principally by impedance errors, the addition of ferrite isolation to the receiver input circuit increased the sensitivity of measurement to near the theoretical limit.

The ferrite isolator was used as a passive transmission element in these experiments. The ferrite circulator, however, was used as an electrically-operated, microwave switch. This switch was used to replace the mechanical chopper in a Dicke-type radiometer. In addition to impedance isolation, the ferrite switch makes possible rapid comparison measurements of the microwave noise powers from any two sources, or of the noise powers from the same source in two different polarizations.

The Characteristic Impedance of the Shielded Slab Line—R. H. T. Bates

The characteristic impedance of the shielded slab line is worked out exactly in terms of elliptic functions. A design graph is given to cover most practical applications.

Characteristics of a New Serrated Choke—Kiyo Tomiyasu and J. J. Bolus

A new type of serrated choke will permit cuts or gaps anywhere on the walls of a rectangular waveguide. The low gap impedance is provided essentially by closely spaced, quarter-wavelength, open-ended, two-wire-line stubs. Low power and high power characteristics of many designs are presented.

Microwave Filters Utilizing the Cutoff Effect—P. A. Rizzi

Two band-rejection microwave filters employing the waveguide cutoff effect are discussed. One type utilizes the cutoff property in the series arm of an E plane tee to improve the filter's characteristics, while the other utilizes this property in the E and H arms of a magic tee. Experimental data for both single and multistage filters are presented. Methods of obtaining low standing wave ratios over a broad pass-band are also presented.

Technique of Pulsing Low Power Reflex Klystrons—J. I. Davis

Very little published information is available on pulsing low power reflex klystrons. Since low power reflex klystrons have been generally designed for cw operation as local oscillators, a minimum of effort has been directed toward the development of specific low power pulse reflex klystrons.

This paper summarizes an effort that has been directed toward pulsing typical low power reflex klystron with a description of the techniques evolved and a summary of the limitations and merits of each technique. Included also is a description of a pulse klystron "priming" technique that minimizes the effect of pulse shortening and leading edge jitter associated with typical pulse operation.

Some Properties of Image Circles—H. F. Mathis

Some properties of image circles for four-terminal networks are discussed. A procedure for correcting an image circle, obtained with a lossy short, is presented. Indirect methods for determining the open and short circuit impedances of a symmetrical four-terminal network are discussed.

Determination of the Parameters of Cavities Terminating Transmission Lines—R. A. Lebowitz

Correspondence

Roster of PGMTT Members

Contributors

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 *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.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Index to the Abstracts and References published in the PROC. IRE from February, 1954 through January, 1955 is published by the PROC. IRE, April, 1955, Part II. It is also published by *Wireless Engineer* and included in the March, 1955 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534.2+538.566 641

On the Focusing Effect of Reflection and Refraction in a Velocity Gradient—Noble. (See 729.)

534.2+538.566 642

Dispersion and Simple Harmonic Point Sources in Wave Ducts—Tolstoy. (See 730.)

534.2-14 643

Theory of Continuous-Tone [underwater] Reverberation—G. A. Klotzbaugh. (*J. acoust. Soc. Amer.*, vol. 27, pp. 956–961; September, 1955.)

534.2-14 644

Continuous-Tone Underwater Reverberation—P. Conley. (*J. acoust. Soc. Amer.*, vol. 27, pp. 962–966; September, 1955.)

534.23-8:546.431.824-31 645

Ultrasonic Shutter—C. L. Darner and R. J. Bobber. (*J. acoust. Soc. Amer.*, vol. 27, pp. 908–912; September, 1955.) The transmission of ultrasonic radiation by a $BaTiO_3$ transducer can be controlled by connecting a suitable electrical network to the electrodes. An intensity range of 40 db can be covered at resonance conditions. The network can be constituted by a carbon microphone, when the ultrasonic beam can be modulated by speech.

534.232 646

Directed Sound—K. Feik. (*Hochfrequenztech. u. Elektroakust.*, vol. 64, pp. 35–62; September, 1955.) Comprehensive analysis is presented leading to the derivation of the radiation characteristics of arrays with linear, ring, spherical, and other arrangements. 40 references.

534.232:537.228.1 647

On the Determination of Electromechanical Coupling Coefficients in a Polyresonant Piezoelectric Vibrator—E. A. G. Shaw. (*Canad. J. Phys.*, vol. 33, pp. 504–508; September, 1955.) The well-known relation between the electromechanical coupling coefficient and the separation σ of resonance and antiresonance frequencies requires modification where several significant modes occur within a small frequency interval. An approximate expression is given for the values of σ which would arise if each mode existed alone, in terms of the many-moded system.

534.613-8 648

Experimental Studies on Acoustic Radiation Pressure—E. M. J. Herrey. (*J. acoust. Soc. Amer.*, vol. 27, pp. 891–896; September, 1955.) Experiments are discussed in which a beam of ultrasonic radiation in water is intercepted by an absorbing or a reflecting disk. Evaluation of the normal and shearing components of the radiation pressure indicates that the latter is properly represented by an anisotropic stress tensor. The technique provides a simple means for measuring the power of a transducer or the reflection, transmission and absorption coefficients of materials. Results of some measurements on cork, Al, and steel plates are presented.

534.75 649

Localization of Sound from Single and Paired Sources—T. T. Sandel, D. C. Teas, W. E. Feddersen, and L. A. Jeffress. (*J. acoust. Soc. Amer.*, vol. 27, pp. 842–852; September, 1955.) Experiments are described in which a listener indicates the direction of (*a*) a single loudspeaker source, (*b*) a pair of phase-aiding loudspeakers, or (*c*) a pair of loudspeakers in phase opposition, by pointing a further loudspeaker in the direction of the source. The loudspeakers to be located emit tone signals alternating with the wide-band noise emitted by the pointer loudspeaker.

534.75 650

Listening to Differentially Filtered Competing Voice Messages—W. Spieth and J. C. Webster. (*J. acoust. Soc. Amer.*, vol. 27, pp. 866–871; September, 1955.) Continuation of investigations described previously [2837 of 1954 (Spieth, et al.)]. In nearly all cases, differential filtering produced an improvement in the ability of an air-control-tower operator to attend to two messages overlapping in time. Details are given of results for different cutoff prequencies.

534.78 651

Vowel Synthesis by means of Resonant Circuits—E. S. Weibel. (*J. acoust. Soc. Amer.*, vol. 27, pp. 858–865; September, 1955.) "Various ways of creating artificial vowels are discussed and compared from the viewpoint of coding efficiency. A method which requires seven parameters to specify any vowel is then introduced and described in detail. It is based on the realization of the transfer impedance of a nonuniform transmission line by lumped elements."

534.78 652

Phonemic Confusion Vectors—H. M. Moser and J. J. Dreher. (*J. acoust. Soc. Amer.*, vol. 27, pp. 874–881; September, 1955.) Experiments are described on the recognition of words used for indicating the letters of the alphabet in telecommunications, with special reference to air-traffic procedure.

534.78 653

Auditory Testing of a Simplified Description of Vowel Articulation—A. S. House and K. N. Stevens. (*J. acoust. Soc. Amer.*, vol. 27, pp. 882–887; September, 1955.)

534.79 654

The Measurement of Loudness—S. S. Stevens. (*J. acoust. Soc. Amer.*, vol. 27, pp. 815–829; September, 1955.) Data regarding the relation between subjective loudness and objective sound intensity are reviewed. The evidence suggests that for the typical listener a loudness ratio of 2:1 corresponds to a pair of sound signals differing by 10 db. At levels below about 50 db the loudness of white noise grows more rapidly than that of a 1 kc tone; above that level the two increase more nearly proportionally.

534.833.1 655

New Method of Recording the Sound Transmission Loss of Walls as a Continuous Function of Frequency—R. V. Waterhouse and R. K. Cook. (*J. acoust. Soc. Amer.*, vol. 27, pp. 967–969; September, 1955.) Voltages derived from microphones in the "loud" and "quiet" rooms separated by the wall under test are applied to a potentiometer recorder which balances itself by driving an attenuator. The difference between the sound levels in the two rooms is nearly identical with the loss in the wall. Results for several types of partition are presented graphically.

621.395.61+621.395.623.7 656

Limits of Quality of Electroacoustic Transducers—F. Spandöck. (*Elektrotech. Z., Edn A*, vol. 76, pp. 598–604; September 1, 1955.) The efficiency of transducers is considered; the value for good-quality microphones is about 0.2 per cent, and for loudspeakers about 5 per cent. On the basis of the reciprocity theorem, a general formula is derived for the transmission factor of all reversible transducers; this formula is used to examine the individual importance of the various factors involved. For transducers of known types, no appreciable improvement is to be expected except by development of the magnetic, piezoelectric, and mechanical properties of the materials used. 62 references.

621.395.61.089.6:534.612 657

Reversion to the Theory of the Gold-Leaf Thermophone—P. Riéty. (*Ann. Télécommun.*, vol. 10, pp. 169–178; July/August, and pp. 195–201; September, 1955.) The problem of the absolute calibration of microphones is considered. Past work on the thermophone, on which the CCIT method is based, has left unexplained discrepancies between the calculated and observed values of the acoustic pressure produced. Analysis is given for static and oscillating conditions; the acoustic pressure is calculated from the value of the estimated temperature close to the gold leaf. Formulas derived are compared with those published previously. For early work see, e.g., *Phys. Rev.*, vol. 10, pp. 22–38; July, 1917, (Arnold and Crandall).

621.395.623.52 658

Analysis of Hypex Horns—R. I. Ścibor-Marchocki. (*J. acoust. Soc. Amer.*, vol. 27, pp. 939–946; September, 1955.) Impedance transformations for hyperbolic-function horns [829 of 1946 (Salmon)] are derived using the Smith chart.

621.395.623.7:537.523.3 659

Corona-Wind Loudspeaker—D. M. Tombs. (*Nature, Lond.*, vol. 176, p. 923; November 12, 1955.) Brief note of experimental results obtained using arrays of triodes comprising two point electrodes and an interposed ring electrode which controls the corona wind without requiring af electrical power. The frequency response is good up to 15 kc; the radiation pattern is adjustable.

ANTENNAS AND TRANSMISSION LINES

621.315.212 660

Extremely Uniform Wide-Band Cables—L. Krügel. (*Telefunken Ztg.*, vol. 28, pp. 107–115; June, 1955. English summary, p. 134.) In order to achieve adequate uniformity in Type-2.6/9.5 cable, the outer conductors are ribbed to prevent deformation, and diamond dies are used for drawing the inner conductors; the spacers are rigid-dielectric disks.

621.315.212:621.372.8 661

Investigations on a Wide-Band Junction between a Coaxial Line and a Waveguide—E. Belohoubek. (*Arch. elekt. Übertragung*, vol. 9, pp. 432–440; September, and pp. 469–474; October, 1955.) Junctions are discussed of the type in which the inner conductor of the coaxial line serves as an antenna inside the waveguide, the latter being provided with a short-circuiting piston. The dependence of bandwidth on antenna length, the eccentricity of the coupling, the position of the piston, and antenna diameter is examined. Bandwidth is greatest for mid-band wavelengths about 1.4 a, where a is the width of the waveguide. A description is given of an arrangement for measuring the bandwidth at different frequencies; bandwidths of 20 per cent are realizable at 10-cm λ, with a reflection coefficient <2 per cent.

621.372.2 662

Reflections in Helical Lines at a Change in the Helix Pitch: Part 2—Mathematical Treatment of the Problem—G. Piefke. (*Arch. elekt. Übertragung*, vol. 9, pp. 402–410; September, 1955.) Part 1: 322 of February.

621.372.2.029.64:538.6 663

Application of Electron Plasma in the Production of Nonreciprocal Systems—A. L. Mikaelyan. (*Bull. Acad. Sci. U.R.S.S., tech. Sci.*, no. 7, pp. 23–33; July, 1955. In Russian.) A study of the propagation of em waves between parallel planes or in a coaxial line containing a layer of electron plasma in a steady magnetic field. Nonreciprocal systems for TEM waves are discussed generally; a brief note suggests that an artificial dielectric could be used in place of the plasma or ferrite; the Hall effect would then be made use of, and no resonance effects would occur in the medium. A similar account is given in *C.R. Acad. Sci. U.R.S.S.*, vol. 104, pp. 72–75; September 1, and pp. 233–236; September 11, 1955; also in Russian.

621.372.22:621.372.51 664

Nonuniform Transmission Lines as Impedance-Matching Sections—J. Willis and N. K. Sinha. (Proc. IRE, vol. 43, p. 1975; December, 1955.) Design analysis is given for lines having better matching properties than the parabolic line described by Yang (3148 of 1955).

621.372.8 665

Degenerate Oscillations in Waveguides—B. Z. Katsenelenbaum. (*Bull. Acad. Sci. U.R.S.S., tech. Sci.*, pp. 9–22; July, 1955. In Russian.) The perturbation of oscillations in a waveguide due to slight deformations in the walls is considered theoretically.

621.372.8 666

Dielectric Transformers for X-Band Waveguide—I. D. Olin. (*Electronics*, vol. 28, pp. 146–147; December, 1955.) Teflon sections are described for (*a*) matching circular to rectangular waveguide, and (*b*) providing a pressure seal for circular waveguide.

621.372.8:621.318.134 667

Magnetically Controlled Microwave Directional Coupler—R. W. Damon. (*J. appl. Phys.*, vol. 26, pp. 1281–1282; October, 1955.) The coupling effected by a Ni-Zn ferrite cylinder linking two stacked rectangular waveguides is controlled by changes in the sense and magnitude of an applied magnetic field parallel to the electric vector. The coupling in db is shown plotted against the magnetic field for a typical device operating at 8.4 kmc.

621.396.67:621.317.3 668

Measurements on Receiving Aerials: Part 2—I. Grosskopf. (*Arch. tech. Messen*, no. 236, pp. 195–196; September, 1955.) Part 1: 3492 of 1955.

AUTOMATIC COMPUTERS

681.142 669

On Starting Routines for the C.S.I.R.O. Mark I Computer—G. W. Hill and T. Pearcey. (*Aust. J. Phys.*, vol. 8, pp. 412–416; September, 1955.) This computer was noted in 640 and 641 of 1954 (Pearcey and Hill).

681.142 670

An Assessment of the System of Optimum Coding used on the Pilot Automatic Computing Engine at the National Physical Laboratory—J. H. Wilkinson. (*Phil. Trans. A*, vol. 248, pp. 253–281; October 20, 1955.) Simple examples are given of programs prepared for the pilot ACE, and an assessment is made of the gain in speed resulting from the use of optimum coding in general. The design of the full-scale ACE is described.

681.142 671

A Criterion for the Operational Stability of Analogue Mathematical Machines—M. Parodi. (*C.R. Acad. Sci., Paris*, vol. 241, pp. 1104–1105; October 24, 1955.)

681.142:512.831 672

Matrices in Analogue Mathematical Machines—P. M. Honnell and R. E. Horn. (*J. Franklin Inst.*, vol. 260, pp. 193–207; September, 1955.) Discussion indicates the usefulness of matrix methods for solving linear differential equations by means of electronic analog machines.

681.142:621.372 673

The Use of Electronic Analogue Computers with Resistance Network Analogues—W. J. Karplus. (*Brit. J. appl. Phys.*, vol. 6, pp. 356–357; October, 1955.) In the solution of partial differential equations by resistance-network analogs the appropriate voltage required at the nodes of the network may be provided automatically by connecting standard electronic-analog-computer units to each node. This procedure combines the speed of the computer method of solution with the inherent accuracy of the resistance-network analog.

681.142+621.318.5](47):016 674

List of Russian and Translated Literature on the Theory of Relay-Contact [switching] Circuits for 1950–1954—Povarov. (See 680.)

681.142:517 675

Approximations for Digital Computers [Book Review]—C. Hastings, Jr. Publishers: Princeton Univ. Press, Princeton, N.J., 201 pp.; 1955. (*Science*, vol. 122, p. 602; September 30, 1955.) The adaptation of special functions for purposes of machine computing is described and illustrated; Tchebycheff approximations are used.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.012:537.312.6 676

The A.C. Admittance of Temperature-Dependent Circuit Elements—R. E. Burgess. (*Proc. phys. Soc.*, vol. 68, pp. 766–774; October 1, 1955.) The general relation between the isothermal and steady-state I/V characteristics of a temperature-dependent device are deduced. Analysis is presented for the small-signal admittance in terms of the thermal admittance. The reactive components due to the thermal inertia and the turnover characteristics exhibited when the ac resistance is negative are examined with reference to equivalent circuits. Where heat loss from the device is by a path which is not "thermally short," the thermal admittance is determined from the diffusion equation. This is done for the case of one-dimensional conduction along the supporting wires of a thermistor or lamp, and for the case of radial conduction from the contact in a point-contact rectifier.

621.314.2:621.375.227.029.3 677

UL ["ultralinear"] Output Transformers—D. M. Leakey and R. B. Gilson. (*Wireless World*, vol. 62, pp. 29–32; January, 1956.) Factors affecting the stability, particularly at high frequencies, of "ultralinear" push-pull output stages are discussed and a practical circuit is described. Avoidance of cross-coupling between the push-pull tubes and correct design of the output transformer, details of which are presented, will usually ensure stability without additional circuitry.

621.316.825:621.3.011.21 678

Impedance of a Thermistor at Low Frequencies—F. J. Hyde. (*J. Electronics*, vol. 1, pp. 303–313; November, 1955.) The small-signal impedance of a directly heated thermistor has been measured at frequencies between 0.116 and 2.34 cps, and the effective circuit time-constant of the same thermistor has been measured as a function of the magnitude of an external series resistance. Assuming the cooling process is Newtonian, the results are in accord with the equivalent circuit proposed by Burgess (676 above).

621.318.5:621.318.134 679

New Type Ferrite Microwave Switch—R. F. Sullivan and R. C. LeCraw. (*J. appl. Phys.*, vol. 26, pp. 1282–1283; October, 1955.) Discussion of a switch comprising a ferrite rod arranged axially in a cylindrical cavity fed and terminated by rectangular waveguides. When a longitudinal magnetic field is applied the cavity is detuned for certain modes and the plane of polarization is rotated; the output waveguide acts as a polarization filter. Good isolation is obtained over a wide range of magnetic-field values. Advantages over the rotation-type switch are indicated.

621.318.5+681.142](47):016 680

List of Russian and Translated Literature on the Theory of Relay-Contact [switching] Circuits for 1950–1954—G. N. Povarov. (*Avtomatika i Telemekhanika*, vol. 16, pp. 411–420 and correction sheet; July/August, 1955.) Bibliography on computers is included.

621.318.57:621.383.4:535.37 681

Opto-electronic Devices and Networks—Loebner. (See 931.)

621.372:512.8/.9 682

Tensor Analysis and Linear Network Theory—R. Braae. (*Trans. S. Afr. IEE*, vol. 46, Part 8, pp. 233–239; August, 1955.) Discussion on 2553 of 1955.

621.372:512.9 683

Examination of the "Negative Frequency" Concept—A. P. Bolle and J. L. Bordewijk. (*J. Brit. IRE*, vol. 15, pp. 582–587; November, 1955. Reprinted from *PTT Bedrijf*, vol. 6; October, 1954.) Circuit-operation analysis is simplified by considering the angular velocity of the vectors representing signals as a combination of a positive and a negative velocity. Applications include the study of time-variable networks, intermodulation noise, Fourier series, and the Fourier integral.

621.372.4.011.1 684

Derivation of Two-Pole Function with Prescribed Complex Values within a Partial Range of Real Frequencies—W. Krägeloh. (*Arch. elekt. Übertragung*, vol. 9, pp. 375–380, 419–431; August, and pp. 479–483; October, 1955.) Two solutions are presented in detail for the problem of deriving a two-pole (positive) function which is a satisfactory approximation to a complex function prescribed graphically.

621.372.413 685

Cavity Resonators with Non-orthogonal Boundaries—E. Ledinegg and P. Urban. (*Acta phys. austriaca*, vol. 9, pp. 335–350; August, 1955.) Analysis based on the particular-integral method described by Hahn (1336 of 1941) is presented for resonators of arbitrary cylindrical form.

621.372.413 686

On Representation of Electromagnetic Fields in Cavities in Terms of Natural Modes of Oscillation—S. A. Schelkunoff. (*J. appl. Phys.*, vol. 26, pp. 1231–1234; October, 1955.) Statements made by Teichmann and Wigner (2268 of 1953) are discussed. The assumption that when a cavity is short-circuited one mode is automatically suppressed is shown to be incorrect.

621.372.45 687

Circuits with Nonlinear Resistance—A. Liebetegger. (*Wireless Engr.*, vol. 33, pp. 24–29; January, 1956.) Analysis is presented for circuits comprising a diode in series with linear impedances which may consist of R, L, C or a combination of these.

621.372.5 688

Transformation Rules for Noisy Quadripoles—W. Dahlke. (*Arch. elekt. Übertragung*, vol. 9, pp. 391–401; September, 1955.) Experimental determination of the noise parameters [2568 of 1955 (Rothe and Dahlke)] becomes difficult at hf because of lead reactances; formulas for the transformations thus introduced are derived and tabulated.

621.372.5 689

Transient Response Calculation—O. P. D. Cutteridge. (*Wireless Engr.*, vol. 33, pp. 29–30; January, 1956.) Advantages of the method using standard gain/frequency curves over the linear-segment method described by Sarma (2559 of 1955) are indicated.

621.372.5:621-526 690

The Response Functions and Vector Loci of First and Second Order Systems—D. Morris. (*Electronic Engng.*, vol. 27, pp. 402–404, 442–444, 499–501; September, and pp. 546–548; December, 1955.) Techniques for applying Nyquist's stability criterion to determine the behavior of automatic control systems operating according to equations of the first degree in frequency are summarized; constructions are given for determining the vector loci of the functions relating performance and frequency. Second-order low-Q systems having a circular vector response locus, giving symmetrical resonance curves, are discussed and the theory of the bandwidth method for determining Q is deduced. Passive and active second-order systems are considered which have finite response at zero frequency and an antiphase or a quadrature zero-limit at infinite frequency, with "cordiform" vector loci. All second-order response loci may be expressed in terms of two basic loci.

621.372.5:621.396.822 691

The Noise Factor of Four-Terminal Networks—A. G. T. Becking, H. Groendijk, and K. S. Knol. (*Philips Res. Rep.*, vol. 10, pp. 349–357; October, 1955.) A simple formula is derived on the basis that the noise properties of the network can be characterized by two noise sources introduced at the input; these sources and their correlation are determined by four quantities which can be evaluated from measurements.

621.372.5.024:621.387 692

A D.C.-Coupled Circuit using Voltage-Stabilizing Valves—G. W. G. Court. (*Electronic Engng.*, vol. 27, pp. 549–550; December, 1955.) Voltage-stabilizing tubes connected in series are used to transfer dc voltage variations from an initial reference level of +150 v to a level of −450 v, providing the reflector potential of a klystron oscillator.

621.372.543.2 693

Staggered Triple Crystal Filter—D. E. Hildreth. (*Electronics*, vol. 28, pp. 166–167; December, 1955.) Narrow-band filters with very sharp cutoff characteristics are more simply with piezoelectric crystals than by the usual lattice construction over the frequency range 400 kc–5 mc. A typical filter has a bandwidth of 400 cps at 3 db attenuation and 800 cps at 40 db attenuation, centered on 400 kc.

621.372.56.027.7 694

Compensated Precision High-Voltage Attenuator—J. G. Cottingham. (*Rev. sci. Instrum.*, vol. 26, pp. 876–877; September, 1955.) The frequency response of a voltage divider consisting of twelve 2.5-MΩ units enclosed by metal corona shields is made linear up to 200 kc by connecting capacitances between the shields and the high-voltage source, thus compensating for the capacitance to ground.

621.372.57:621.316.721 695

Some High-Impedance Current-Generating Circuits—J. H. McGuire. (*Electronic Engng.*, vol. 27, pp. 529–531; December, 1955.) A basic circuit is described and practical designs are developed for the efficient generation of current waveforms which are independent of the load impedance.

621.373:[621.314.632+621.316.825 696

Electrical Oscillations in Thermistors and Germanium Point-Contact Rectifiers—R. E. Burgess. (*J. Electronics*, vol. 1, pp. 297–302; November, 1955.) The conditions for small-amplitude oscillations in nonlinear temperature-dependent circuit elements are deduced analytically and discussed in relation to thermistors and Ge point-contact rectifiers, whose impedance comprises inductive reactance in series with a resistance which can be made negative, so that these devices can sustain oscillations when shunted with a suitable capacitance. The upper limiting frequency and the dependence of frequency on the shunt capacitance are deduced in terms of the fundamental parameters. Departures from the postulated model which can arise with the point-contact rectifier are discussed.

621.373.421 697

Constant-Frequency Oscillators—A. S. Gladwin. (*Wireless Engr.*, vol. 33, pp. 13–19; January, 1956.) "The conditions for the frequency of a regenerative oscillator to be independent of changes in the input and output resistances of the maintaining amplifier are derived. It is shown that the input, output, and transfer impedances of the feedback network must be resistive at the oscillation frequency. This can be achieved by inserting reactances of definite specified values in the input and output leads. Expressions for the value of these reactances are derived for the usual type of LC oscillator with mutual-inductance coupling. The values depend greatly on whether the losses in the coils can be represented mainly by a series or by a parallel resistance. Expressions for the stability of an imperfectly-stabilized oscillator are also derived. The results are compared with those obtained by Llewellyn. It is shown that the values of stabilizing reactances specified by Llewellyn are not in general satisfactory, though in some instances they may produce a marked improvement in stability. Full experimental confirmation is given."

621.373.43 698

A Sufficiently Fast and Economical Sweep Circuit—D. Brini, L. Peli, O. Rimondi, and P. Veronesi. (*Nuovo Cim.*, vol. 2, pp. 644–646; September 1, 1955. In English.) A circuit giving sweep times from 40 mμs to 3μs is based on a design described by Moody et al., (2730 of 1952) using secondary-emission tubes.

621.375.2.029.3:621.396.712.2 699

Peak-Limiting Amplifiers—R. Yadav. (*J. Instn. Telecommun. Engrs., India*, vol. 1, pp. 147–154; September, 1955.) A description is given of a circuit designed by All India Radio; the steps taken to reduce the attack time to <16 μs are indicated. The control range above threshold is 15 db and the compression ratio is 10:1.

621.375.221.029.63 700

Wide-Band Amplifier for U.H.F. Receivers—R. B. McWhirt. (*Electronics*, vol. 28, pp. 158–160; December, 1955.) A detailed design is given for an amplifier using a Type-416B planar triode, providing 5–10 db gain, with a bandwidth >200 mc centered on 1.1 mc.

621.375.23.029.3 701

Tetrodes with Screen Feedback—*Wireless World*, vol. 62, pp. 24–26; January, 1956.) The operation of the "ultralinear" af amplifier [1512 of 1952 (Hafler and Keroes)] is discussed; the advantage is that triode performance as regards low inherent distortion is combined with power efficiency approaching that of a pentode.

621.375.4:621.314.7 702

Some Properties and Circuit Applications of Super-Alpha Composite Transistors—Pearlman. (See 927.)

621.375.4:621.314.7 703

Bases for the Calculation of Transistor Amplifiers with Series and Parallel Feedback—W. Benz. (*Telefunken Ztg.*, vol. 28, pp. 95–107; June, 1955. English summary, pp. 133–134.) The various possible arrangements are classified and discussed in terms of network theory. Relevant definitions, inter-relations, equivalent circuits, and formulas are set out in ten tables.

621.375.4:621.314.7 704

The Equations and the Equivalent Circuit of the Transistor—Skalicky. (See 928.)

621.375.4:621.314.7 705

Germanium Transistor Amplifiers Stable to 95°C—W. Greatbatch and W. Hirtreiter. (PROC. IRE, vol. 43, p. 1974; December, 1955.) Though transistors are available whose current gain does not change greatly on heating up to 100°C, the operating point of the corresponding amplifier circuit may shift appreciably at far lower temperatures. In a particular circuit using two transistors the operating point can be stabilized by including a parallel-*RLC* network in a feedback loop. Si units can be stabilized at temperatures above 100°C.

GENERAL PHYSICS

53.088.3 706

Correcting for Running Means by Successive Substitutions—R. N. Bracewell. (*Aust. J. Phys.*, vol. 8, pp. 329–334; September, 1955.) The proposed procedure leads to a divergent result but is nevertheless useful in many practical cases.

530.145.6:621.385.833 707

Wave-Mechanics Theory of Electron-Optical Image Formation: Part 2—W. Glaser and G. Braun. (*Acta phys. austriaca*, vol. 9, pp. 267–296; August, 1955.) Part 1: 2253 of 1955.

534.372 708

Connection between the Logarithmic Decrement of Attenuation and the Fractional Attenuation of the Amplitude of Potential Energy (Cyclical Viscosity)—D. I. Shil'krut. (*C.R. Acad. Sci. U.R.S.S.*, vol. 104, pp. 237–238; September 11, 1955. In Russian.) Earlier experimental evidence on torsional and other oscillations suggests that the logarithmic decrement, δ, is time-dependent. It may be expressed by the formula $2\delta = \Delta a/a - \Delta c/c$, where a is the total energy of oscillation, Δa the energy loss per cycle, c the potential energy for unit amplitude, and Δc the change in c per cycle; c is a function of time.

535.376 709

Reinforcement of the Electroluminescence of Certain Crystals by Rotation in the Field—G. Destriau. (*C.R. Acad. Sci., Paris*, vol. 241, pp. 869–870; October 3, 1955.) When electroluminescent crystals are embedded in a dielectric and not in contact with the field electrodes, luminescence is normally observed only in alternating fields, though it can be produced in direct fields by rotating the crystals. Experimental evidence is now reported that the luminescence in an alternating field is enhanced by rotating the crystals. Long-term falling off of electroluminescence may be partly due to surface effects at the crystal/dielectric boundary, these effects being eliminated when the crystals are continually displaced.

535.41:535.215 710

Photoelectric Mixing of Incoherent Light—A. T. Forrester, R. A. Gudmundsen and P. O. Johnson. (*Phys. Rev.*, vol. 99, pp. 1691–1700; September, 15, 1955.) Experiments are described in which beats were obtained between incoherent light sources by mixing Zeeman components of a visible spectral line at a photoemissive surface. A 3-cm-λ cavity-resonator arrangement was used for detection. The results are consistent with the view that the photoelectric emission probability is proportional to the square of the resultant electric-field amplitude, implying interference between light from independent sources. The results also indicate that any time delay between photon absorption and electron emission must be significantly less than 10^{-10} seconds.

537.311.62 711

Effect of Collisions between Electrons on Electrical Conductivity and Skin Effect in Metals—V. L. Ginzburg and V. P. Silin. (*Zh. eksp. teor. Fiz.*, vol. 29, pp. 64–74; July, 1955.) Theoretical considerations indicate that the results of the theory of the anomalous skin effect are not greatly affected by taking electron collisions into account. The work on the anomalous skin effect and the reflectivity of metals by Benthem and Kronig (3523 of 1954) is criticized.

537.322.2 712

Measurements of the 3rd-Order Thermoelectric Single-Material Effect (1st Benedicks Effect)—G. Kocher. (*Ann. Phys., Lpz.*, vol. 16, pp. 210–226; September 20, 1955.) The coefficient c_3 in the formula $u - c_3 \cdot \Delta t_3$, where u is the emf and Δt is the temperature difference, was determined experimentally using small rectangular loops of Pt, Au, and Ag, heated and cooled at two diametrically opposite points. For a loop of dimensions 38×58 mm made of material with cross-section 4 mm², the value of c_3 was about -2.7×10^{-14}v/deg³ for Pt, -4.6×10^{-15} for Au, and -2.6×10^{-15} for Ag.

537.322.3 713

Remarks on the 1st Benedicks Effect—W. Meissner. (*Ann. Phys., Lpz.*, vol. 16, pp. 227–228; September 20, 1955.) Comments on paper by Kocher (712 above). The coefficient c_3 may be a function of Δx, the distance between the two points with temperature difference Δt; this is of importance when Δx is smaller than the mean free path of the conduction electrons.

537.5 714

Traveling Density Waves in Positive Columns—S. Watanabe and N. L. Oleson. (*Phys. Rev.*, vol. 99, pp. 1701–1740; September 15, 1955.) "The continuity equations for positive ions and for electrons coupled to each other by the Coulomb force and the effect of ionizing collisions have a solution representing traveling density waves whose frequencies are widely different from the usual plasma oscillations."

537.52:538.56 715

Amplification of 1c-m Waves in the Helium Negative Glow—Z. Geller and W. Low. (*Nature, Lond.*, vol. 176, pp. 1021–1022; November 26, 1955.) A detailed examination was made of the attenuation of 1.25-cm-λ waves in an abnormal discharge in He at a pressure of 1–3 mm Hg, by directing a beam from a klystron at right angles to the discharge and detecting the radiation transmitted. A graph shows attenuation as a function of distance from the discharge-tube cathode for given discharge conditions. Negative attenuation values are observed at distances around 13 mm; the electron density in this region is estimated to be about 7×10^{12}/cm³. The amplification may be due to plasma oscillations; a secondary dip in the attenuation curve may be due to a harmonic.

537.525:621.3.029.53/.55 716

Influence of a Transverse Magnetic Field on the Breakdown Voltage of a Gas Discharge at High Frequency—L. Ferretti and P. Veronesi. (*Nuovo Cim.*, vol. 2, pp. 639–643; September 1, 1955.) Measurements were made on tubes with cylindrical electrodes containing air at pressures of 0.1, 0.5, and 1 mm Hg, at frequencies from 10 to 50 mc, with magnetic fields of strength 0–650 G. Results are presented graphically and discussed in relation to those of Herlin and Brown (3390 of 1948).

537.533:537.311.33 717

Field emission from Semiconductors—R. Stratton. (*Proc. phys. Soc.*, vol. 68, pp. 746–757; October 1, 1955.) Theory is presented derived from that developed by Fowler and Nordheim for field emission from metals (*Proc. roy. Soc. A*, vol. 119, pp. 173–181; May 1, 1928). If there are an appreciable number of surface states, the emission current is very low until the field is strong enough to break down the internal barriers. When this occurs, or if there are no surface states, the field penetrates into the semiconductor, lowering the conduction-band edge. This causes degeneracy near the surface and greatly enhanced emission. Emission associated with surface states is temperature dependent while emission associated with field penetration is not.

537.533:621.385.83 718

Relation between an Arbitrary Field $\vec{B}(r, \phi, z)$ and the Field of Revolution $B^R(r, z)$ such that $B_z^R(0,z)\vec{B}_z(0,z)$. Applications—P. Gautier. (*C.R. Acad. Sci., Paris*, vol. 241, pp. 930–932; October 10, 1955.) Electron-optical analysis using cylindrical coordinates leads to a theorem on the measurement of an arbitrary magnetic field by means of search coils with rotational symmetry.

537.533:[546.74+546.77 719

Inelastic Scattering of Electrons by Ni and Mo Targets—A. R. Shul'man and I. I. Farbshtein. (*C.R. Acad. Sci. U.R.S.S.*, vol. 104, pp. 56–59; September 1, 1955. In Russian.) An experimental investigation is briefly reported. Energy-loss peaks for electrons inelastically scattered by Mo were observed at 5.5, 11.6, and 17 ±0.8 v and these values were independent of the energy of the incident electrons in the range from 40 to 120 ev. In scattering by Ni the peak occurred at 12.5 ±0.5 v. The results for the Mo target agree to within 11 per cent with the values calculated for molybdenum oxide, and evidence that such a film was probably present in spite of precautions was also obtained from secondary electron emission measurements.

537.533.8 720

Theory of Secondary Electron Emission from Solids—H. Fröhlich. (*Proc. phys. Soc.*, vol. 68, pp. 657–660; September 1, 1955.) In the derivation of a formula permitting the calculation of the yield of secondary electrons from the yield of photoelectrons, the primary particle is replaced by a moving point charge. The electric displacement is developed in a Fourier series in time and to each frequency a photoelectron yield is ascribed.

537.56:538.56 721

Eigen Oscillations of Compressible, Ionized Fluids—R. E. Loughhead. (*Aust. J. Phys.*, vol. 8, pp. 416–418; September, 2955.) The frequencies of eigen-oscillations are derived from the hydromagnetic equations for a compressible fluid of infinite electrical conductivity for uniform-magnetic-field regions bounded by cylindrical and plane-parallel surfaces.

537.56:538.6 722

Hydromagnetic Stability of a Current Layer—R. E. Loughhead. (*Aust. J. Phys.*, vol. 8, pp. 319–328; September, 1955.) "The hydromagnetic stability of a uniform current flowing along a magnetic field and confined within a pair of parallel planes is discussed by the method of normal modes. The condition for marginal stability is derived and discussed with reference to two special cases. It is also shown that the velocity of Alfvén waves along a magnetic field in a region bounded by parallel planes is reduced due to the inertia of the surrounding medium."

537.56:538.6 723

Solution of Problems involving the Hydromagnetic Flow of Compressible Ionized Fluids—R. E. Loughhead. (*Phys. Rev.*, vol. 99, pp. 1678–1681; September 15, 1955.) "The characteristic forms of the hydromagnetic equations for a compressible fluid are examined from the viewpoint of obtaining by the method of finite differences numerical solutions for continuous initial value problems involving unidimensional motion of the fluid."

537.562:[537.29+538.69 724
Velocity-Distribution Function of Electrons in Alternating Electric and Constant Magnetic Fields—A. V. Gurevich. (*C. R. Acad. Sci. U.R.S.S.*, vol. 104, pp. 201–204; September 11, 1955. In Russian.) A calculation similar to that of Jancel and Kahan (1995 of 1953) on the em effects in the ionosphere is presented and some errors in that paper are indicated. Using the calculated velocity distribution, the mean electron energy and the plasma conductivity are calculated.

538:536.48 725
Conference on Low-Temperature Magnetism: Kharkov, 1st–3rd July 1954—(See 822.)

538.11 726
The Spin-Wave Theory of Antiferromagnetism—W. Marshall. (*Proc. roy. Soc. A*, vol. 232, pp. 69–77; October 11, 1955.) Following a discussion of antiferromagnetism (*ibid*, pp. 48–68) with particular reference to the calculations of Kubo (*Rev. mod. Phys.*, vol. 25, pp. 344–351; January, 1953), various criticisms of the spin-wave theory are considered; conclusions drawn from this theory, particularly as to the conditions required for antiferromagnetism, are not to be relied upon.

538.114 727
Quantum Theory of Ferromagnetism—S. V. Vonsovski, K. B. Vlasov, and E. A. Turov. (*Zh. eksp. teor. Fiz.*, vol. 29, pp. 37–50; July, 1955.)

538.24 728
Dependence of Magnetic Viscosity on Dimensions of Specimens—A. N. Remizov. (*C.R. Acad. Sci. U.R.S.S.*, vol. 104, pp. 389–390; September 21, 1955. In Russian.) Results are presented graphically of an experimental investigation on cylindrical armco specimens of lengths up to 50 cm and diameters up to 10 mm.

538.566+534.2 729
On the Focusing Effect of Reflection and Refraction in a Velocity Gradient—W. J. Noble. (*J. acoust. Soc. Amer.*, vol. 27, pp. 888–891; September, 1955.) Analysis based on geometrical optics is presented for wave propagation in a stratified medium. Divergence factors are tabulated for four different system configurations.

538.566+534.2 730
Dispersion and Simple Harmonic Point Sources in Wave Ducts—I. Tolstoy. (*J. acoust. Soc. Amer.*, vol. 27, pp. 897–907; September, 1955.) Analysis is presented for wave propagation in stratified media. An exact solution is obtained for the general case of waves from a point source in a duct by introducing an integral solution due to Brekhovskikh (1394 of 1950).

538.566:537.56 731
Theory of Propagation of Plane Waves in Homogeneous Plasma—G. Winkler. (*Ann. Phys., Lpz.*, vol. 16, pp. 414–428; September 20, 1955.) The theory combines the macroscopic-phenomenological and microscopic-statistical approaches. Assumptions made include: (*a*) that charge-carrier velocity is very much smaller than the velocity of light; (*b*) that the plasma is homogeneous and stationary; (*c*) that the amplitudes of disturbances due to the wave propagation are small; (*d*) that collisions between charge carriers and neutral particles are inelastic, and (*e*) that the electric and magnetic moments of neutral molecules and of ions and the magnetic moments of free electrons may be neglected.

538.566:538.6:537.56 732
Investigation of the Faraday Effect in a Plasma at 10 kMc/s—M. Bonnet, M. Matricon, and É. Roubine. (*Ann. Télécommun.*, vol. 10, pp. 150–158; July/August, 1955.) Experimental and theoretical investigations are reported. An arrangement comprising a circular waveguide containing a pulsed neon tube was used to demonstrate rotation of the plane of polarization; the magnetic field obtainable in the experimental arrangement was not strong enough to produce gyromagnetic resonance at the frequency used. Analysis based on bicomplex numbers is presented for propagation in an ionized medium in a magnetic field.

538.566.2/.3:537.56:538.6 733
The Four Possible Waves in a Magneto-ionic Medium—J. H. Piddington. (*Phil. Mag.*, vol. 46, pp. 1037–1050; October, 1955.) The theory presented shows that a magneto-ionic medium composed mainly of electrons and heavy ions is quadruply refracting, transmitting four, and only four, different waves. The hydromagnetic ordinary and extraordinary waves are identical with the radio *O* and *E* waves, the latter being the limiting case of the former for frequencies at which the motion of the heavy ions is negligible. The other two waves comprise a "magnetic sound" wave (a type of hydromagnetic wave) and a "magnetic plasma" (*P*) wave. The *P* wave is discussed using transport equations and an explanation is given of nonthermal solar radio noise. The effect of electron pressure on the *O* and *E* waves is determined quantitatively. A short account of the work is given in *Nature, Lond.*, vol. 176, pp. 875–876; November 5, 1955. See also 91 of January and back references.

538.569.4:535.34:546.21 734
Microwave Absorption in Compressed Oxygen—A. A. Maryott and G. Birnbaum. (*Phys. Rev.*, vol. 99, p. 1886; September 15, 1955.)

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 735
Galactic Survey at 400 M/cs between Declinations −17° and −49°—R. X. McGee, O. B. Slee, and G. J. Stanley. (*Aust. J. Phys.*, vol. 8, pp. 347–367; September, 1955.) The survey was carried out using an antenna with beam width of 2 degrees between half-power points. Results are tabulated and contour diagrams of equivalent sky temperature are given. The method of calibrating the equipment in terms of temperature is described.

523.16:538.561 736
The Observation and Interpretation of Radio Emission from some Bright Galaxies—B. Y. Mills. (*Aust. J. Phys.*, vol. 8, pp. 368–389; September, 1955.) Observations are reported of eight southern galaxies in the magnitude range 7 to 11, the Magellanic Clouds, and the Milky Way in the region of the galactic center. A radiotelescope operating at 3.5-m λ was used. The results obtained tend to support the ideas of Shklovski (*Astr. J., Moscow*, vol. 29, p. 418; 1953, and vol. 30, p. 15; 1954) according to which the galactic emission comprises two subsystems, one showing a discoidal distribution highly concentrated towards the galactic plane, the other a very dispersed and approximately spherical distribution centered on the galactic center. Contrary to Shklovski's interpretation, the contribution of thermal emission from ionized hydrogen to the discoidal distribution is likely to be small.

523.2:538.12 737
The Magnetic Field in the Vicinity of the Solar System—W. L. Kraushaar. (*Progr. theor. Phys.*, vol. 14, p. 77; July, 1955.) A brief discussion of the significance for particle trajectories of the assumption of the existence of an approximately uniform magnetic field in the vicinity of the solar system. See also *ibid.* pp. 78–79.

523.5:621.396.11 738
Diurnal Variations in the Number of Shower Meteors detected by the Forward-Scattering of Radio Waves: Part 1—Theory—C. O. Hines. (*Canad. J. Phys.*, vol. 33, pp. 493–503; September, 1955.) System factors affecting the observations are combined to provide a weighting factor varying in correspondence with the diurnal variation of the shower-radiant position, so that the effect of this variation can be distinguished from that of random changes in the incidence rate of the meteors.

523.53:621.396.96 739
The Radar Determination of Meteor Showers in the Southern Hemisphere—C. D. Ellyett and K. W. Roth. (*Aust. J. Phys.*, vol. 8, pp. 390–410; September, 1955.) A report is presented of observations made during 1953, at a frequency of 69 mc using radar pulses of peak power 75 kw, duration 3.5 μs, and recurrence rate 145-per second.

523.72:538.56.029.6 740
Three-Millimeter Wave Radiation from the Sun—W. Gordy, S. J. Ditto, J. H. Wyman, and R. S. Anderson. (*Phys. Rev.*, vol. 99, p. 1905; September 15, 1955.) Radiation of wavelength 3.2 mm has been received with a signal /noise ratio of about 25. Typical records obtained with movable and fixed antennas are reproduced.

523.72:621.396.822 741
A Method of discriminating Disturbances in V.H.F. Solar Noise Observation—K. Kawakami, T. Takahashi, and M. Onoue. (*J. Radio Res. Labs, Japan*, vol. 2, pp. 111–123; April, 1955.) Continuous records are made of the total noise signal received at 200 mc and of the signal after passage through a waveform discriminator which separates out the atmospheric and man-made noise. Solar noise, being random in character, appears only on the first record.

523.78:551.510.535 742
Solar Eclipses and the Ionosphere—W. J. G. Beynon. (*Nature, Lond.*, vol. 176, pp. 947–948; November 19, 1955.) Report of a symposium held in London in August, 1955.

550.372 743
The Electrical Conductivity of the Earth's Mantle—S. K. Runcorn. (*Trans. Amer. geophys. Union*, vol. 36, pp. 191–198; April, 1955.)

550.385:551.593 744
Changes in Brightness, Polarization, and Colour of the Zenith Day Sky accompanying Geomagnetic Activity—D. R. Barber. (*J. atmos. terr. Phys.*, vol. 7, pp. 170–172; September, 1955.) Observed effects are attributed to variations in the size of the particles causing the magnetic disturbance.

551.510.5 745
The Molecular Diffusive Rate of Change of Composition in the Atmosphere—S. Chapman. (*J. Met.*, vol. 12, pp. 111–116; April, 1955.) Calculations are made for an atmosphere uniformly mixed, for heights up to 80 km.

551.510.5:621.396.96 746
Resolution in Height of a Radar Pulse—Miles. (See 764.)

551.510.53 747
The Aeronomic Problem of Nitrogen Oxides—M. Nicolet. (*J. atmos. terr. Phys.*, vol. 7, 7, pp. 152–169; September, 1955.) The concentration of nitric oxide in the upper atmosphere is calculated for various heights; its vertical distribution follows that of the atmosphere. Maximum production occurs in the *E* layer; diffusion and atmospheric mixing can provide an explanation of the ionization of the *D* layer.

551.510.534 748
New Experimental and Theoretical Inves-

tigations on the Atmospheric Ozone Layer—H. K. Paetzold. (*J. atmos. terr. Phys.*, vol. 7, pp. 128–140; September, 1955.) For altitudes from 24 km to 60 km the ozone distribution determined experimentally is in satisfactory agreement with theory; below 24 km the predicted values of ozone content are higher than the measured values. Further world-wide measurements are required and an ozone radiosonde apparatus is being developed for the purpose.

551.510.535 749

Chapman Theory of Photo-ionization for a Nonisothermal Atmosphere—F. Mariani. (*Ann. Geofis.*, vol. 8, pp. 59–73; January, 1955.) The atmospheric temperature is assumed to vary linearly up to a certain height, beyond which it is constant. Analysis indicates that the assumption of a plane earth makes no appreciable difference to the value found for the ionization density, at least for zenith angles <60 degrees. The ionization distribution and the height of maximum ionization are appreciably different from the values given by Chapman's original theory.

551.510.535 750

Critical Considerations on the Ionosphere F Layer—P. Dominici and F. Mariani. (*Ann. Geofis.*, vol. 8, pp. 103–120; January, 1955.) From general considerations of the layer-formation processes, observed anomalies are found to be associated with the F_2 layer. They cannot be explained by thermal expansion alone, but can be explained by the dynamo theory. Superposition of the F_1 and true F_2 layers can account for some anomalies, but calculations depend on the correct estimation of the virtual height of the F_2 layer. The importance of the temperature/height distribution is indicated in relation to the hypothesis that the whole of the F layer is formed by the same ionization process. To reduce experimental $h'(f)$ curves to $N(z)$ curves it is necessary to use numerical methods not based on *a priori* models.

551.510.535 751

On the Influence of Electron-Ion Diffusion on the Electron Density and Height of the Nocturnal F_2 Layer—T. Yonezawa. (*J. Radio Res. Labs, Japan*, vol. 2, pp. 125–136; April, 1955.) Analysis indicates that electron-ion diffusion will not change the distribution in a Chapman layer, but will cause a fall in the height of maximum density. Since the height of the F_2 layer does not change measurably after sunset, this suggests a lower limit to the molecular density in the F_2 region, considered as a Chapman layer of 5×10^9 cm^{-3}.

551.510.535 752

Distortion of the E Layer of the Ionosphere by Electrical Currents flowing in it—E. Appleton, A. J. Lyon, and A. G. Turnbull. (*Nature, Lond.*, vol. 176, pp. 897–899; November 12, 1955.) The divergence of actual ionospheric conditions from conditions corresponding to an ideal Chapman layer is investigated in the light of the asymmetrical diurnal variation of E-layer properties and its dependence on latitude. The evidence indicates that the apparent values of recombination factor are affected by a factor depending on electron transport, and that the S_q current system exerts a marked influence on the maximum value N_m of ionization in the E layer and on the height at which the maximum occurs. Abnormal diurnal asymmetry and the latitude dependence of N_m (*Proc. Fourth Meeting Mixed Commission on the Ionosphere*, p. 14; 1955) together confirm the presence of ionospheric currents of the type predicted by Balfour Stewart in 1882.

551.510.535:523.78 753

Electron Production and Recombination in the E-Layer—C. M. Minnis. (*J. atmos. terr. Phys.*, vol. 7, p. 172; September, 1955.) An error in a previous paper is pointed out; corrected values reinforce the conclusion reached.

551.510.535:621.3.087.4 754

Measurement of Ionospheric Heights by the Method of Delayed Coincidence—H. Rakshit and S. D. Chatterjee. (*Z. Phys.*, vol. 141, pp. 540–549; September 20, 1955. In English.) See 713 of 1955.

551.510.535:621.396.11 755

The Physics of the Ionosphere—Publishers: The Physical Society, London, 406 pp.; 1955. This report contains the text of the following papers presented at the conference noted previously (1340 of 1955):

Survey: The Lowest Ionosphere—A. H. Waynick (pp. 1–13).

Ionospheric Absorption at Low and Medium Frequencies—G. McK. Allcock (pp. 14–19).

Ionospheric Layer Formation under Quasi-stationary Conditions—E. Appleton and A. J. Lyon (pp. 20–39).

A Study of Vertical Incidence Ionospheric Absorption at 2 Mc/s—W. J. G. Beynon and K. Davies (pp. 40–52).

Short Wave Echoes from the Lower Ionosphere—W. Dieminger (pp. 53–57).

Solar X-ray Emission and the Height of the D Layer during Radio Fade-Out—H. Friedman and T. A. Chubb (pp. 58–62).

D-Region Echoes with a Radio Wave of Frequency 1.4 Mc/s—S. Gnanalingam and K. Weekes (pp. 63–70).

Self-Gyrointeraction—S. N. Mitra (pp. 71–73).

The Observed Polarization of High-Frequency Sky-Wave Signals at Vertical Incidence—M. G. Morgan and W. C. Johnson (pp. 74–77).

M-Type Ionospheric Reflections at 150 kc/s—R. L. Schrag (pp. 78–87).

Survey: A Survey of Existing Knowledge of Irregularities and Horizontal Movements in the Ionosphere—J. A. Ratcliffe (pp. 88–98).

Scattering at Oblique Incidence from Ionospheric Irregularities—Summary by J. A. Ratcliffe of informal talk by D. K. Bailey (pp. 99–100).

Variations of the F_2-Layer Critical Frequency during Undisturbed Ionospheric Conditions—W. Becker (pp. 101–103, in German with English abstract).

Interpretation of Drift of Field-Strength Nonuniformities as Ionospheric Winds—H. Berg (pp. 104–112, in German with English abstract).

Some Observations of Ionospheric Movements—K. Bibl, E. Harnischmacher, and K. Rawer (pp. 113–118).

An Empirical Study of Random Functions which arise in the Interpretation of Ionospheric Movements—B. H. Briggs and E. S. Page (pp. 119–122).

The Variability of Time Shifts in Measurements of Ionospheric Movements—B. H. Briggs and M. Spencer (pp. 123–135).

A Note on the Motion of a Cylindrical Irregularity in an Ionized Medium—P. C. Clemmow, M. A. Johnson, and K. Weekes (pp. 136–139).

Statistical Analysis of Satr Scintillations as a means for Investigating Atmospheric Disturbances—R. Fürth (pp. 140–149).

Irregularities in the E Region caused by Atmospheric Electricity—G. A. Isted (pp. 150–162).

Interpretation of Observed F_2 "Winds" as Ionization Drifts associated with the Magnetic Variations—D. F. Martyn (pp. 163–165).

Investigation of F-Region Drift Movements by Observations of Radio Star Fading—A. Maxwell (pp. 166–171).

Radio Star Scintillations due to Ionospheric Focusing—J. L. Pawsey (pp. 172–173).

Diffuse Ionospheric Reflection of Medium Waves at Night—G. J. Phillips (pp. 174–180).

Travelling Disturbances in the Ionosphere—R. E. Price (pp. 181–190).

Measurement of Ionospheric Wind from the Drift Velocity of a Condition (e.g. Field Strength of Echo) along the Surface of the Earth—P. St. Pütter (pp. 191–201, in German with English abstract).

A Method for studying Large-Scale Irregularities in Region E of the Ionosphere—R. L. Schrag (pp. 202–211).

Survey: A Survey of Present Knowledge of the F_2 Region—D. F. Martyn (pp. 212–218).

Storm Phenomena and the Solar-Cycle Variations of the F_2-Layer Ionization at Noon—E. Appleton and W. R. Piggott (pp. 219–228).

Electrodynamics of the Outer Atmosphere—J. W. Dungey (pp. 229–236).

The Ionospheric F_2 Region—R. J. Havens, H. Friedman, and E. O. Hurlburt (pp. 237–244).

Geomagnetic Distortion in the F_2 Layer—K. I. Maeda (pp. 245–253).

Theory of Height and Ionization Density Changes at the Maximum of a Chapman-like Region, taking account of Ion Production, Decay, Diffusion and Tidal Drift—D. F. Martyn (pp. 254–259).

Geomagnetic Anomalies of the F_2 Region and their Interpretation—D. F. Martyn (pp. 260–264).

Ionospheric F_2 Variations associated with Geomagnetic Disturbances at the Equatorial Zone—S. Matsushita (pp. 265–269).

Information about the F_2 Layer taken from Ionization Maps—K. Rawer (pp. 270–275).

Survey: The Mathematics of Wave Propagation through the Ionosphere—K. G. Budden (pp. 276–287).

Propagation of a Plane Wave in a Transparent Anisotropic Ionized Medium and Propagation in a Heterogeneous Ionized Medium—E. Argence (pp. 288–298, in French).

Electron Density and Ray Tracing in the Ionosphere Layers—P. A. Bricout (pp. 299–307).

The Transient Response of the Ionosphere at Low Frequencies—S. A. Bowhill (pp. 308–319).

The Nonexistence of a "Fourth Reflection Condition" for Radio Waves in the Ionosphere—K. G. Budden (pp. 320–331).

A Method for Determining the Variation of Electron Density with Height ($N(z)$ curves) from Curves of Equivalent Height versus Frequency (h', f) curves)—K. G. Budden (pp. 332–339).

The Dependence of the Refractive Index in Magnetoionic Theory on the Direction of the Wave Normal—P. C. Clemmow and R. F. Mullaly (pp. 340–350).

Methods of Solving the Coupled Equations of Plane Wave Propagation in the Ionosphere—J. J. Gibbons (pp. 351–354).

Ray Theory and a New Method for Ray Tracing—J. Haselgrove (pp. 355–364).

Statistical Mechanics of Lorentz-Type Electron Plasmas and Applications to the Ionosphere—R. Jancel and T. Kahan (pp. 365–373, in French).

Analysis of the Coupling of the Ordinary and Extraordinary Electromagnetic Waves in a Lorentz-Type Plasma and Application to the Ionosphere—R. Jancel and T. Kahan (pp. 374–383, in French).

Graphical Constructions for Ray-Tracing in the Ionosphere—R. F. Mullaly (pp. 384–393).

Studies of the Refractive Index in the Ionosphere: the Effect of the Collision Frequency and of Ions—W. Pfister (pp. 394–401).

Tables of Group Refractive Index for the Ordinary Ray in the Ionosphere—D. H. Shinn (pp. 402–406).

551.510.535:621.396.812.3 756

Three Components of the Field Strength of the Wave Reflected from the Surface of the Ionosphere their Level and Time Variation—

Uyeda, Ogata, Uchikura, Arima, and Obayashi. (See 900.)

551.515.4:621.396.96 757
Radar Observations of a Thunderstorm—Mitra. (See 765.)

LOCATION AND AIDS TO NAVIGATION

621.396.93 758
Adcock Direction Finder—W. C. Bain. (*Wireless Engr.*, vol. 33, pp. 20–24; January, 1956.) "It was found that the pull of the triatics (supporting the central sense antenna) on the main antennas of an Adcock hf direction finder caused the main antennas to bend inwards, so that a large polarization error was produced on frequencies near 20 mc. Theoretical calculations are made for the case of antennas with an over-all slope of 1 degree, with a distant loop transmitter at 6.3 degrees elevation and with a loop tilt of 6 degrees from the horizontal: the total polarization error found is 6 degrees for a perfectly-reflecting ground and 19 degrees for ground of conductivity 0.02 mho/m. In practical direction finding the errors are not likely to be as high as this on the average; but the figure of 3 degrees is obtained for the root-mean-square error from a transmitter on 20 mc at 20 degrees elevation despite the small amount of antenna bending assumed."

621.396.932/.933 759
Night Effect in the Consol Navigation System—R. Kümmich. (*Fernmeldetech. Z.*, vol. 8, pp. 494–500; September, 1955.) It is shown that the combination of the ground wave with the sky wave causes deviation of all the beams except the main one; in some cases the rhythm of the signal is also disturbed.

621.396.932/.933 760
Accuracy of the "Sonne" [consol] Navigation System—K. Rawer. (*Fernmeldetech. Z.*, vol. 8, p. 510; September, 1955.) Brief note summarizing results of an unpublished 1944 report on an extensive series of observations. Night-time deviations and scatter of observations were consistent with present-day knowledge of ionospheric propagation.

621.396.933 761
Long-Range Radio Navigation in the Long-Wave Band—E. Roessler. (*Fernmeldetech. Z.*, vol. 8, pp. 485–489; September, 1955.) The Navaglobe-Navarho system [432 of 1955 (Clark, et al.)] is discussed and compared with other systems in use, from the point of view of the number of ground stations required, the bandwidth requirements and the sensitiveness to noise.

621.396.933.1:621.397.5 762
Telegee—D. A. Levell. (*Wireless World*, vol. 62, pp. 41–42; January, 1956.) The possibility of using existing synchronized television transmitters for determining the position of aircraft is discussed. An aircraft would need to receive three television stations; the times of arrival of corresponding frame synchronizing pulses could be measured and position determined as in the Gee and Decca systems.

621.396.96 763
Experimental Study of Rain Echoes using 3-cm-λ Pulsed Radar with Circular Polarization—M. Bouix, M. Clément, and C. Frémiot. (*Ann. Telecommun.*, vol. 10, pp. 159–168; July/August, 1955.) Observations were made using radar equipment with two different types of antenna permitting rapid transition from linear to circular polarization; constructional details are given. Results confirm that rain clutter is reduced by using circular polarization at this wavelength.

621.396.96:551.510.5 764
Resolution in Height of a Radar Pulse—V. G. Miles. (*J. Met.*, vol. 12, pp. 107–110; April, 1955.) "An analysis is made of the ability of a radar pulse to resolve, either completely or partially, the echoes from two closely-spaced horizontal reflecting layers situated close to the radar. The general expressions obtained are illustrated graphically." See also 3306 of 1953.

621.396.96:551.515.4 765
Radar Observations of a Thunderstorm—H. Mitra. (*Indian J. Met. Geophys.*, vol. 6, pp. 119–136; April, 1955.) A detailed illustrated account. The radar echoes as displayed on a 3-cm-λ Type-AN/APQ-13 set are explained by reference to meteorological data.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5:535.33 766
The Emission of Light from Gas Discharges in High Vacuum—K. H. Reiss. (*Z. angew. Phys.*, vol. 7, pp. 433–437; September, 1955.) A method of analyzing gases at low pressure is based on their emission spectra. A specially constructed ionization manometer with cold electrodes is used as light source. Apparatus including filter, photomultiplier, and recorder is described. The technique is useful for testing vacuum apparatus for leaks.

533.5:621.3.032.73 767
Outgassing of Glass—B. J. Todd. (*J. appl. Phys.*, vol. 26, pp. 1238–1243; October, 1955.) The temperature dependence of the water-diffusion rate in vacuum has been determined for several glasses. From the results it is possible to calculate the amount of water that will be evolved from glass for any time-temperature conditions following any bake-out in which the surface gases have been removed.

535.215:[537.311.33+537.226 768
Performance of Photoconductors—A. Rose. (Proc. IRE, vol. 43, pp. 1850–1869; December, 1955.) A survey paper. A phenomenological analysis is made of processes whereby the conductance of a semiconductor or insulator material is increased by exposure to radiation (including particle radiation but excluding thermal radiation) the theory is based on consideration of bound energy states. The lifetime of free carriers is an important characteristic parameter. The mechanism of conduction at barriers is discussed and noise phenomena are examined. 54 references.

535.215:537.311.33 769
Photoeffects in Intermetallic Compounds—H. P. R. Frederikse and R. F. Blunt. (Proc. IRE, vol. 43, pp. 1828–1835; December, 1955.) "The intermetallic semiconductors are classified with respect to their crystal structure and to the place of the component elements in the periodic system. A survey is given of the properties of compounds with the zincblende and fluorite lattice. Photoeffects of individual members of these two groups are discussed. Such phenomena include photoconduction, photovoltage, and the photoelectric-magnetic effect. 40 references."

535.215:537.311.33 770
Photoconductivity of the Sulfide, Selenide, and Telluride of Zinc or Cadmium—R. H. Bube. (Proc. IRE, vol. 43, pp. 1836–1850; December, 1955.) Recent research on evaporated, powder, and sintered layers and on single-crystal specimens is reviewed. The aspects discussed include spectral response, impurity sensitization, electrode contact problems, speed of response, dependence of photocurrent on illumination intensity and on temperature, infrared quenching and thermal stimulation, as well as related luminescence, photovoltaic, and photoemissive effects. 118 references.

535.215:537.311.33:546.26-1 771
Photoelectric Emission from Polycrystalline Graphite—E. Taft and L. Apker. (*Phys. Rev.*, vol. 99, pp. 1831–1832; September 15, 1955.) Results of measurements of the energy distribution of the photoelectrons are reported. The photoelectric yield from graphite is about a tenth that from common metals.

535.215:537.311.33:[546.28+546.289 772
Photoconduction in Germanium and Silicon—M. L. Schultz and G. A. Morton. (Proc. IRE, vol. 43, pp. 1819–1828; December, 1955.) A general account based on energy-band theory is presented explaining the photoconducting properties of Ge and Si both with and without added impurities. 58 references.

535.215:537.311.33:621.383.4 773
Lead Salt Photoconductors—T. S. Moss. (Proc. IRE, vol. 43, pp. 1869–1881; December, 1955.) A review paper discussing the mechanism of photoconductivity in PbS, PbTe, and PbSe and its application in high-sensitivity infrared-radiation detectors. The manufacture of photocells using these materials is outlined. 130 references.

535.37:535.34:546.482.21 774
The Absorption Spectrum of Excited Crystals of Cadmium Sulphide—A. Halperin and G. F. J. Garlick. (*Proc. phys. Soc.*, vol. 68, pp. 758–765; October 1, 1955.) Single crystals of CdS excited by light of wavelength 0.546 μ exhibited absorption bands extending from the excitation wavelength to 1.4 μ with a maximum intensity at 0.78 μ. Variations caused by temperature changes are discussed on the basis of electron and hole trapping and recombination.

535.37:546.472.21 775
Investigation of the Accumulation Property and Origin of Levels of Localization of Electrons in Zinc-Sulphide Phosphors—V. N. Zhukova. (*C.R. Acad. Sci. U.R.S.S.*, vol. 103, pp. 1001–1004; August 21, 1955. In Russian.)

535.37:546.472.21 776
Luminescent Centers in ZnS:Cu:Cl Phosphors—R. Bowers and N. T. Melamed. (*Phys. Rev.*, vol. 99, pp. 1781–1787; September 15, 1955.)

535.37:546.472.21:535.215:537.311.33 777
Photoluminescent Modulation in Nonuniformly Excited ZnS Phosphors—R. E. Halsted. (*Phys. Rev.*, vol. 99, p. 1897; September 15, 1955.) Experiments are briefly described in which phosphors were excited by radiation of wavelengths shorter than that corresponding to the fundamental absorption edge, while subjected to an alternating field of frequency ranging from 20 cps to 20 kc. The results demonstrate the *n*-type photoconductivity of the phosphors and clarify the role of conduction electrons in the luminescence process.

535.37.07:548.0 778
The Physical Chemistry of Crystal Phosphors—F. A. Kröger. (Proc. IRE, vol. 43, pp. 1941–1944; December, 1955.) A brief survey of inorganic crystal phosphors, with particular attention to the incorporation of atoms with a valency different from that of the main material, and to the stabilization of atoms in a particular valency.

535.376 779
Cathodoluminescence—G. F. J. Garlick. (Proc. IRE, vol. 43, pp. 1907–1911; December, 1955.) A brief review of recent advances in knowledge of the fundamental mechanisms involved. Observed brightness/voltage characteristics are related to the microcrystalline form of the phosphor; simpler behavior is exhibited by single crystals. Factors affecting screen efficiency are discussed, as are also current saturation and "electron burn."

535.376 780
Electroluminescence and Related Topics—G. Destriau and H. F. Ivey. (Proc. IRE, vol. 43, pp. 1911–1940; December, 1955.) A comprehensive review with 162 references.

535.376:546.472.21 781
Electroluminescence in Single Crystals of Zinc Sulphide—G. F. Alfrey and J. B. Taylor. (*Proc. phys. Soc.*, vol. 68, pp. 775–784; October 1, 1955.) Experimental work is described and the bearing of the results on previously proposed theories of the mechanism of electroluminescence is discussed. Predicted temperature and frequency variations of electroluminescenc in good agreement with observed values are obtained on the basis of a field-emission theory assuming a particular formation of the surface barrier.

535.376:546.472.21:537.226.2 782
The Electrical Properties of Electroluminescent Phosphors—A. N. Ince and C. W. Oatley. (*Phil. Mag.*, vol. 46, pp. 1081–1103; October, 1955.) The frequency variations of ϵ_1 and ϵ_2, the components of the complex permittivity $\epsilon_1 - j\epsilon_2$, were experimentally determined for activated ZnS both in the dark and under ultraviolet illumination. Results suggest that two different mechanisms are involved. One, which is present in the dark, produces dispersion at low frequencies; it is characteristic of electroluminescent phosphors. The other, which is stimulated by ultraviolet light, produces dispersion similar to that found in ordinary nonelectroluminescent phosphors. The effects are explained on the assumption that electronic barriers are formed in the specimen and some of the ions in the barrier region are mobile.

535.376:621.385.832:535.623 783
Phosphors for Tricolour Television Tubes—A. Bril and H. A. Klasens. (*Philips Res. Rep.*, vol. 10, pp. 305–318; October, 1955.) Measurements of the efficiency and spectral distribution of various phosphors indicate that the properties of the red phosphor are most important; compromise is necessary between color rendering and efficiency. (Zn, Cd)S-Ag and (Zn, Cd)Se-Cu phosphors can be produced with higher efficiencies for near-white colors than ZnP-Mn. The properties of a number of phosphors are shown in tables and graphs.

537.226/.228 784
Nonlinear Dielectric Materials—E. T. Jaynes. (Proc. IRE, vol. 43, pp. 1733–1737; December, 1955.) A brief physical discussion of materials in which dielectric properties depend on applied field strength. Since ferroelectric materials have high dielectric constants they may be expected to exhibit a high degree of nonlinearity.

537.226/.227 785
Some Aspects of Ferroelectricity—G. Shirane, F. Jona, and R. Pepinsky. (Proc. IRE, vol. 43, pp. 1738–1793; December, 1955.) A survey with over 160 references. The behavior of ferroelectric and antiferroelectric materials is related to their domain and crystal structure. Numerous photographs and diagrams illustrate the phenomena described.

537.226/.227:546.817.824 786
X-Ray Study of Phase Transition of Ferroelectric $PbTiO_3$ at Low Temperature—J. Kobayashi and R. Ueda. (*Phys. Rev.*, vol. 99, pp. 1900–1901; September 15, 1955.)

537.226.2 787
The Permittivity of Two-Phase Mixtures—C. A. R. Pearce. (*Brit. J. appl. Phys.*, vol. 6, pp. 358–361; October, 1955.) Experimental results obtained by various workers are compared with predictions from published formulas relating the properties of a mixture with those of the phases composing it. An empirical expression is presented which gives adequate agreement with all relevant available results.

537.226.2:546.431.824–31 788
Permittivity of Barium Titanate—I. P. Kozlobaev. (*C.R. Acad. Sci. U.R.S.S.*, vol. 104, pp. 387–388; In Russian.) Experimental results, presented graphically, show that ϵ decreases as the hydrostatic pressure increases from atmospheric to 4000 kg/cm². A mechanical hysteresis effect was also observed; in a specimen with a Curie temperature Θ of about 127°C, the final value of ϵ at atmospheric pressure was greater than the initial value at high pressure, while in specimens with Θ about 10–20°C above room temperature this difference was smaller, and in a specimen with $\Theta = 1°C$ the final value was smaller than the initial one. The relaxation of internal stresses, as indicated by changes in ϵ, is practically complete in about 20 minutes.

537.228.1 789
Anisotropy of Polarized Polycrystalline Barium Titanate—D. S. Moseley. (*J. acoust. Soc. Amer.*, vol. 27, pp. 947–950; September, 1955.) Observations were made of the longitudinal vibrations of $BaTiO_3$ specimens subjected to a range of polarization conditions; pairs of bars $1 \times \frac{3}{8} \times \frac{3}{8}$ in. were used, one of each pair having electrodes applied to the ends and the other having electrodes on opposite long sides. The effect of different treatments on the ratio between the elastic compliances at constant electric field and at constant electric displacement is indicated and discussed.

537.311.31 790
Effect of Point Imperfections on the Electrical Properties of Copper: Part 1—Conductivity—F. J. Blatt. (*Phys. Rev.*, vol. 99, pp. 1708–1716; September 15, 1955.) Detailed calculations are presented for the scattering of conduction electrons by lattice imperfections in Cu.

537.311.33 791
Unipolar Conduction at the Semiconductor/Metal Boundary—E. Wallauschek. (*Acta phys. austriaca*, vol. 9, pp. 351–366; August, 1955.) Measurements on silver/silver-iodide cells are reported; the current/time characteristics for the forward and reverse directions are different. A unipolar conduction mechanism responsible for this asymmetry is discussed.

537.311.33 792
Carrier-Concentration Disturbances in Semiconductors—R. E. Burgess. (*Proc. phys. Soc.*, vol. 68, p. 793; October, 1 1955.) Critical comment on a paper by Low (3296 of 1955).

537.311.33 793
Fluctuations of the Numbers of Electrons and Holes in a Semiconductor—R. E. Burgess. (*Proc. phys. Soc.*, vol. 68, pp. 661–671; September 1, 1955.) Fluctuations are calculated by either (*a*) a thermodynamical method, in terms of the electronic free energy of the system whose minimal properties are used to determine the equilibrium distribution and fluctuations of electrons between the conduction and valence bands and the impurity levels, or (*b*) a statistical method based on the probabilities of the elementary processes of electron transitions between energy levels. The first method is particularly useful in complex cases. Both intrinsic and extrinsic semiconductors are discussed, as well as the case of a nearly intrinsic semiconductor.

537.311.33 794
Theory of the Infrared Absorption by Carriers in Semiconductors—R. B. Dingle. (*Phys. Rev.*, vol. 99, pp. 1901–1902; September 15, 1955.) Critical discussion of theory presented by Kahn (2649 of 1955).

537.311.33 795
Surface Recombination and the "Light-Disk" Method of measuring Bulk-Lifetime in Semiconductors—J. Butterworth. (*J. Electronics*, vol. 1, pp. 293–296; November, 1955.) The method considered is broadly the same as that discussed *e.g.* by van Roosbroeck (2644 of 1955), but the linear dimensions of the illuminated area at which minority carriers are injected are assumed to be comparable with the diffusion length. Two expressions useful for analyzing experimental data are derived, valid for complementary ranges of the system parameters.

537.311.33 796
A Comparison of some Methods for the Determination of Trace Impurities in Semiconductors—A. A. Smales. (*J. Electronics*, vol. 1, pp. 327–332; November, 1955.) A discussion of analytical methods; coulometry, x-ray spectroscopy, microbiological, and magnetic-resonance techniques are excluded.

537.311.33 797
History of Semiconductor Research—G. L. Pearson and W. H. Brattain. (Proc. IRE, vol. 43, pp. 1794–1806; December, 1955.) An account going back to Faraday's observation in 1833 of the negative temperature coefficient of resistance of silver sulphide; most of the emphasis is on the work leading up to the development of the transistor. 118 references.

537.311.33:535.37 798
Radiative Transitions in Semiconductors—R. Braunstein. (*Phys. Rev.*, vol. 99, pp. 1892–1893; September 15, 1955.) Observations are reported of radiation produced by carrier injection into thin plates of GaSb, GaAs, InP, and Ge-Si alloys, at room temperature and at 77°K; an outline is given of the technique used. The spectral distributions exhibit peaks at energy values close to previously estimated values of the energy gaps.

537.311.33:536.2 799
Thermal Conductivity of Semiconductors—J. M. Thuillier. (*C.R. Acad. Sci., Paris*, vol. 241, pp. 1121–1122; October 24, 1955.) An exact calculation indicates that the ratio between the electrical contribution to the thermal conductivity and the electrical conductivity may be much greater than the value predicted by the Lorentz theory for an electron gas.

537.311.33:537.533.8 800
New Ni-Be Alloy with [high] Secondary Emissivity—J. Millet. (*Le Vide*, vol. 10, pp. 96–102; July/September, 1955.) A description of the preparation of Ni-Be and Ni-Mg alloys suitable for use as nonlinear resistors. The structure and physical properties of the alloys are discussed with the aid of microphotographs. The electrical properties are discussed in a paper by Bobenrieth, et al. (776 of 1955). See also 434 of 1954 (Teszner, et al.).

537.311.33:538.632 801
Saturation Hall Constant of Semiconductors—J. A. Swanson. (*Phys. Rev.*, vol. 99, pp. 1799–1807; September 15, 1955.) Analysis is presented for the motion of electrons in semiconductor crystals in the presence of crossed electric and magnetic fields. For increasing values of the magnetic field the value of the Hall constant tends asymptotically to $1/(p-n)ec$ where p and n are the hole and electron concentrations and e and c have their usual significance; this value is unaffected by scattering. The saturation magnetoresistance is also discussed.

537.311.33:546.23 802
Fixation of Iodine in Selenium—R. A. Hyman and D. H. Tomlin. (*Nature, Lond.*, vol. 176, p. 977; November 19, 1955.) Brief report of experimental results indicating that halogen atoms can be bound in the Se structure, probably at the ends of chains of Se atoms. This conclusion does not agree with that of Nijland (744 of 1955).

537.311.33:546.26–1 803
Recent Investigations on Graphite—R. Arnold. (*Z. angew. Phys.*, vol. 7, pp. 453–460; September, 1955.) A review with 58 references.

537.311.33:546.27 804
Study of the Frequency Dispersion of the Conductivity of Boron and of $B_{12}Al$—J. Lagrenaudie. (*J. Phys. Radium*, vol. 16, pp. 731–732; August/September, 1955.) Report of measurements at frequencies in the range 100 cps-1 mc and temperatures down to 110°K, using specimens solidified after fusion. The dispersion of conductivity is attributed to nonuniform distribution of impurities. Estimates are made of activation energies of impurities.

537.311.33:[546.28+546.289 805
Effect of Water Vapor on Grown Germanium and Silicon *np* Junction Units—J. T. Law and P. S. Meigs. (*J. appl. Phys.*, vol. 26, pp. 1265–1273; October, 1955.) Photoresponse and reverse current were measured over a wide range of water-vapor pressures. At low relative humidities the surface recombination rate for Ge decreases relative to the value in vacuum, but Si is not affected. Humidities above 30 per cent produce channels on both Ge and Si units. Channel length varied with voltage as observed by McWhorter and Kingston (3587 of 1954). Steady illumination decreased the channel length on both Ge and Si units, but affected the excess reverse current only for Si; hence some mechanism other than channel conduction, probably an ionic mechanism, is operative at Ge surfaces. The existence of the two mechanisms is consistent with Christensen's results (3585 of 1954). See also 3586 of 1954 (Law).

537.311.33:[546.28+546.289]:548.0 806
The Electronic Energy Band Structure of Silicon and Germanium—F. Herman. (PROC. IRE, vol. 43, pp. 1703–1732; December, 1955.) This introduction to the physics of crystals covers electronic quantum states, the hole concept, the effective mass tensor, motion of electrons and holes, spin-orbit interaction, the energy-band structure of Si and Ge and their alloys, theory of lattice vibrations, the phonon concept, collisions between electrons or holes and phonons, electrical conductivity, and optical phenomena. Over 70 references.

537.311.33:546.289 807
Nature of the Water-Vapor-Induced Excess Current on Grown Germanium *p-n* Junctions—F. N. Clarke. (*Phys. Rev.*, vol. 99, pp. 1899–1900; September 15, 1955.)

537.311.33:546.289 808
Theory of Acceptor Levels in Germanium—W. Kohn and D. Schechter. (*Phys. Rev.*, vol. 99, pp. 1903–1904; September 15, 1955.) Calculations based on the coupled effective-mass equations applicable for a degenerate band structure give a value for the ionization energy in good agreement with observations.

537.311.33:546.289 809
Diffusion of Antimony, Arsenic and Indium into Solid Germanium—W. Bösenberg. (*Z. Naturf.*, vol. 10a, pp. 285–291; April, 1955.) Steep impurity-concentration profiles develop during the preparation of Ge single crystals by pulling, if impurities are incorporated in the melt. The flattening of the slope due to diffusion is determined by impedance measurements; impurity concentrations between 1 in 10^5 and 1 in 10^7 atomic parts are used. Values of the diffusion coefficients are hence derived. The same values are obtained when thin films of the impurity metals are evaporated on to the Ge and allowed to diffuse into the interior.

537.311.33:546.289:535.215 810
Analysis of the Decay of Photoconductance in Germanium—B. H. Schultz. (*Philips Res. Rep.*, vol. 10, pp. 337–348; October, 1955.) "A method of analyzing surface and volume effects in the recombination of injected charge carriers is described. The influence of the capacity of a surface double layer is discussed and some of the results of measurements are given."

537.311.33:546.289:548.0 811
Structural Changes in Evaporated Ge Films in an Electron Microscope—E. W. Fischer and H. Richter. (*Ann. Phys., Lpz.*, vol. 16, pp. 193–208; September 20, 1955.) A photographically illustrated report of changes induced by the impact of the electron beam.

537.311.33:546.3-1-28-289 812
Electrical Properties of Germanium-Silicon Alloys—A. Levitas. (*Phys. Rev.*, vol. 99, pp. 1810–1814; September 15, 1955.) Various specimens were investigated; resistivity measurements were made over the temperature range 300°–800°K, and Hall-effect and resistivity measurements over the temperature range 77°–300°K. Intrinsic resistivity at room temperature is plotted against percentage Si in the alloy, by extrapolation from the high-temperature measurements. The observed variation of mobility with alloy composition is indicative of alloy scattering.

537.311.33:546.682.86 813
Thermoelectric Power of Indium Antimonide—H. P. R. Frederikse and E. V. Mielczarek. (*Phys. Rev.*, vol. 99, pp. 1889–1890; September 15, 1955.) Measurements of the thermoelectric power of a *p*-type and an *n*-type specimen over the temperature range 60°–400°K are reported; wide differences are found between the two characteristics.

537.311.33:546.682.86 814
Infrared Absorption in Indium Antimonide—W. G. Spitzer and H. Y. Fan. (*Phys. Rev.*, vol. 99, pp. 1893–1894; September 15, 1955.) Measurement results are presented as transmisson/wavelength and reflectivity/wavelength curves.

537.311.33:546.682.86 815
Optical Absorption Limit and Effective Mass of Electrons in Indium Antimonide (InSb)—P. Aigrain and J. des Cloizeaux. (*C.R. Acad. Sci., Paris*, vol. 241, pp. 859–862; October 3, 1955.) Calculations are made which provide an explanation of the discrepancy between the values of the effective electron mass found by Burstein (2134 of 1954) and by Dresselhaus, et al. (3011 of 1955). The analysis is based on the potential distribution due to impurity centers, and its application is not restricted to InSb.

537.311.33:546.817.221 816
Measurements on *p-n* Junctions in Lead Sulphide—T. S. Moss. (*Proc. phys. Soc.*, vol. 68, pp. 697–700; October 1, 1955.) "From measurements of the characteristics of several natural *p-n* junctions the intrinsic resistivity of lead sulphide is calculated to be 3.1Ω cm. Hence at room temperature the intrinsic carrier concentration is 2.7×10^{15} electrons or holes per cm^3 and the width of the forbidden zone 0.40 ev."

537.311.33:621.396.822 817
Contribution on the 1/*f* Law of Noise in Semiconductors—H. Schönfeld. (*Z. Naturf.*, vol. 10a, pp. 291–300; April, 1955.) Explanations of the 1/*f* spectral distribution of current noise in semiconductors given by van der Ziel (3035 of 1950) and by Surdin (1325 of 1952) are based on assumptions not previously justified. It is shown that discrete phenomena distributed according to a $t^{-1/2}$ time law also give rise to a 1/*f* spectral distribution. This is discussed in relation to Montgomery's hypothesis (122 of 1953) that the discrete phenomena correspond to the local injection of minority carriers.

537.311.33:621.396.822 818
Distribution Function of Semiconductor Noise—D. A. Bell. (*Proc. phys. Soc.*, vol. 68, pp. 690–691; September 1, 1955.) An experiment is described confirming that the Gaussian distribution of instantaneous values applying to "white noise" applies also to current noise with inverse-frequency spectrum.

537.311.33:669.046.54/.55 819
Zone-Melting Processes under Influence of the Atmosphere—J. van den Boomgaard. (*Philips Res. Rep.*, vol. 10, pp. 319–336; October, 1955.) Theory presented previously [*e.g.*, 740 of 1955 (Reiss)] is extended to deal with the case of elements containing volatile solutes; it is possible to introduce a volatile impurity element homogeneously into an ingot by means of zone melting under a constant vapor pressure of that element.

537.311.33:669.046.54.55:512.831 820
On Zone Refining—J. L. Birman. (*J. appl. Phys.*, vol. 26, pp. 1195–1197; October, 1955.) "A matrix method is used to solve the difference equations which describe the zone refining of a bar. The matrix method is designed for direct numerical calculation of the solute distribution, after any number of passes, with any intitial solute distribution. The method is illustrated with a simple example."

537.311.33.01+535.37.01 821
Statistical-Kinetic Theory of Luminescence and Electrical Conductivity of Impurity Semiconductors—I. Broser and R. Broser-Warminsky. (*Ann. Phys., Lpz.*, vol. 16, pp. 361–407; September 20, 1955.) By using statistical rather than kinetic considerations the mathematical analysis is simplified, and restrictive assumptions regarding discrete energy levels can be discarded.

538:536.48 822
Conference on Low-Temperature Magnetism: Khar'kov, 1st–3rd July 1954—(*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 19, pp. 387–488; July/August, 1955. In Russian.) The issue contains a summary of the 20 papers read at the conference and texts of nine papers including:

Galvanomagnetic Phenomena and Properties of Conduction Electrons in Metals—E. S. Borovik (pp. 429–443).

Investigation of Photomagnetoelectric Effect in Cuprous Oxide at Low Temperatures—A. P. Komar, N. M. Reinov, and S. S. Shalyt (pp. 444–446).

Electrical Conductivity of Ferromagnetic Metals at Low Temperatures—E. A. Turov (pp. 474–480).

538.221 823
The Direct Separation of the Reversible and Irreversible Components of the Magnetothermal Effect—L. F. Bates and N. P. R. Sherry. (*Proc. phys. Soc.*, vol. 68, pp. 642–648; September 1, 1955.) The two components are separated by effecting small "backward" increments in magnetization at given points in the hysteresis cycle. Results are reported for cobalt.

538.221 824
Investigation by Powder-Pattern Method of the Magnetic Structure of Silicon-Iron Crystals—Ya. S. Shur and V. R. Abel's. (*C.R. Acad. Sci. U.R.S.S.*, vol. 104, pp. 209–210; September 11, 1955. In Russian.) From the photographs shown, general relations are deduced of the dependence of magnetic structure on the crystallographic orientation of the specimen surface.

538.221 825
Shape and Crystal Anisotropy of Alnico 5—E. A. Nesbitt and H. J. Williams. (*J. appl. Phys.*, vol. 26, pp. 1217–1221; October, 1955.) An investigation made using the torque-measurement technique previously described [172 of 1955 (Nesbitt, et al.)] indicates that the high coercive force of single crystals of alnico 5 depends on the shape anisotropy of the fine precipitated plates, the crystal anisotropy being negligible. The plates are composed of rods of diameter about 75 Å and length 400 Å, but exhibit single-domain properties.

538.221 826
Permanent-Magnet Properties of Elongated Single-Domain Iron Particles—L. I. Mendelsohn, F. E. Luborsky, and T. O. Paine. (*J. appl. Phys.*, vol. 26, pp. 1274–1280; October, 1955.) An account is given of measurements on magnets made by aligning and compacting single-domain particles of diameter 150 Å with a median length/diameter ratio of 3 and an intrinsic coercive force before packing of 1,600 oersted. The influence of packing fraction and degree of alignment were investigated. BH_{max} values $>3\times10^6$ G. oersted were obtained. The results are compared with theoretical predictions and with the corresponding properties of existing permanent-magnet materials.

538.221 827
The Effect of Particle Interaction on the Coercive Force of Ferromagnetic Micropowders—E. P. Wohlfarth. (*Proc. roy. Soc. A*, vol. 232, pp. 208–227; October 25, 1955.) Analysis is presented for an assembly of prolate-spheroidal single-domain particles, with either parallel or nonparallel arrangement. The principal effects discussed are the dependence of the shape-anisotropy coefficient on the interaction and a mutual magnetization change due to an interaction-dependent rotation of the magnetization vectors. A general expression is derived for the coercive force in terms of packing factor and saturation magnetization; this is contrasted with formulas obtained previously based on an "effective field" treatment.

538.221 828
Production of Anisotropy in a Permanent Magnet by Pressure—J. E. Gould and M. McCaig. (*Nature, Lond.*, vol. 176, p. 977; November 19, 1955.) The "squareness" of the *B/H* curve of a rod of 35 per cent Co steel was improved by shrinking a stainless-steel block around it so as to exert lateral pressure. The method is similar to that applied to ferroxcube by Wijn, et al. (*Philips tech. Rev.*, vol. 16, p. 49; 1954). See also 3608 of 1954 (Wijn).

538.221:538.569.4.029.6 829
Ferromagnetic Resonance in Nickel and in some of its Alloys—K. J. Standley and K. H. Reich. (*Proc. phys. Soc.*, vol. 68, pp. 713–722; October 1, 1955.) Report of determinations of *g* values and line widths for polycrystalline specimens of Ni and of Ni-Cu, Ni-Al, Ni-Sb, and Ni-Mn alloys with high Ni content, from resonance measurements at 1.22-cm λ, mainly over the temperature range from 20°C to 200°C.

538.221:[621.318.124+621.318.134 830
Some Properties of Ferrites in Connection with Their Chemistry—E. W. Gorter. (PROC. IRE, vol. 43, pp. 1945–1973; December, 1955.) Introductory and survey paper, including 63 references. Aspects discussed include ferromagnetic resonance and dimensional resonance and relaxation phenomena and their relation to permeability.

538.221:[621.318.124+621.318.134 831
Anisotropy and Magnetostriction of some Ferrites—R. M. Bozorth, E. F. Tilden, and A. J. Williams. (*Phys. Rev.*, vol. 99, pp. 1788–1798; September 15, 1955.) Measurements are reported on Co, CoZn, and other ferrites, some single-crystal and others polycrystalline. Crystal anisotropy was investigated by the torque technique; values up to 4×10^6 ergs/cm³ were found. Magnetostriction was determined by strain-gauge technique; values up to 800×10^{-6} were found. Magnetic annealing is effective at temperatures as low as 150°C, and causes the hysteresis loop to become square. In polycrystalline Co ferrites magnetic annealing is most effective for compositions between $CoFe_2O_4$ and Fe_3O_4. The constants for the various specimens are tabulated and discussed.

538.221:621.318.124 832
Studies on the Oxide Magnets: Part 1—Effects of Bi_2O_3 on Barium Ferrites. Part 2—Effects of Bi_2O_3 on Strontium and Lead Ferrites—T. Okamura, H. Kojima, and S. Watanabe. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, vol. 7, pp. 411–424; August, 1955.) The saturation and remanent magnetization of a BaO-Fe_2O_3 system were increased by 20 to 30 per cent, and the coercive force by 30 to 50 per cent, by an addition of 1.5 per cent Bi_2O_2; the mechanical properties improved simultaneously. The results in SrO-Fe_2O_3, SrO-CaO-Fe_2O_3, and PbO-Fe_2O_3 systems are more complex.

538.221:[621.318.124+621.318.134]:535.33/.34-1 833
Infrared Spectra of Ferrites—R. D. Waldron. (*Phys. Rev.*, vol. 99, pp. 1727–1735; September 15, 1955.)

538.221:621.318.134 834
First-Order Magnetic Anisotropy Constants of Ferrites—J. B. Birks. (*Phys. Rev.*, vol. 99, p. 1821; September 15, 1955.) Potential sources of error in the method of calculation presented by Weisz (1413 of 1955) are indicated.

538.221:621.318.134 835
Frequency Dependence of Magnetocrystalline Anisotropy—R. M. Bozorth, B. B. Cetlin, J. K. Galt, F. R. Merritt, and W. A. Yager. (*Phys. Rev.*, vol. 99, p. 1898; September 15, 1955.) Differences are observed between the values of crystal anisotropy of Ni and Mn ferrites measured by static and microwave methods. Conclusions are drawn regarding the structure of the materials.

539.23:537.311.31:546.59 836
Study of Transparent, Highly Conducting Gold Films—E. J. Gillham, J. S. Preston, and B. E. Williams. (*Phil. Mag.*, vol. 46, pp. 1051–1068; October, 1955.) Thin gold films deposited on a substratum of bismuth oxide or, better, between two films of bismuth oxide, are found to have exceptionally high electrical conductivity and optical transmission. Typical values, for films of thickness 100Å, are 3.5Ω/square for the resistance and 75 per cent for the transmission.

546.681 837
Study of the Electrical Anisotropy of Gallium in the Neighbourhood of the Fusion Point—I. Epelboin and M. Erny. (*C.R. Acad. Sci., Paris*, vol. 241, pp. 1118–1121; October 24, 1955.)

548.0:[537.311.33+549.514.51 838
Densities and Imperfections of Single Crystals—A. Smakula, J. Kalnajs, and V. Sils. (*Phys. Rev.*, vol. 99, pp. 1747–1750; September 15, 1955.) The densities of Si, Al, CaF_2, CsI, Ge, TlCl, TlBr, and quartz have been computed from lattice constants and molecular weights and compared with values determined by hydrostatic weighing by a method described by Smakula and Sils (*ibid.*, pp. 1744–1746). Discrepancies between the two sets of values are discussed in relation to the reliability of the basic data, and conclusions are drawn regarding the crystal imperfections.

548.0:537.311.33:[546.28+546.289 839
Energy Levels of a Crystal modified by Alloying or by Pressure—R. H. Parmenter. (*Phys. Rev.*, vol. 99, pp. 1759–1766; September 15, 1955.) A perturbation method of calculating the energy levels is discussed; application to crystals with diamond-type lattice is illustrated. See also *ibid.*, pp. 1767–1776.

548.0:537.311.33:621.386 840
A Scanning X-Ray Camera for the Detection of Crystal Imperfections—P. J. Holmes. (*J. Electronics*, vol. 1, pp. 324–326; November, 1955.)

549.514.51 841
Liquid Inclusions in Artificial Quartz—V. P. Butuzov and N. Yu. Ikornikova. (*C.R. Acad. Sci. U.R.S.S.*, vol. 104, pp. 76–77; September 1, 1955. In Russian.) An experimental investigation of methods for eliminating bubbles is briefly reported.

549.514.51 842
Experimental Evidence for Dislocations in Crystalline Quartz—H. E. Bömmel, W. P. Mason, and A. W. Warner, Jr. (*Phys. Rev.*, vol. 99, pp. 1894–1896; September 15, 1955.)

621.315.61:537.533.9 843
Effects Produced by Metal Electrodes in Studies of Electron-Bombardment Conductivity—J. C. Firmin and C. W. Oatley. (*Proc. phys. Soc.*, vol. 68, pp. 620–624; September 1, 1955.) The effect of the metallic film used as an electrode in the determination of electron-bombardment-induced conductivity of crystal insulators has been investigated and its bearing on the results of earlier workers is discussed. In thick films x-rays may be generated, which penetrate the specimen and produce hole/electron pairs in the body of the material.

621.315.612 844
Phase Equilibrium Relations in the Systems Titania-Niobia and Zirconia-Niobia—R. S. Roth and L. W. Coughanour. (*J. Res. nat. Bur. Stand.*, vol. 55, pp. 209–213; October, 1955.) Fundamental studies on ceramic dielectrics are reported.

621.315.613.1:621.396.822 845
Pre-breakdown Current and Noise in Insulators—D. A. Powers and T. Suita. (*J. appl. Phys.*, vol. 26, pp. 1244–1247; October, 1955.) The frequency spectrum of noise associated with prebreakdown currents has been investigated for mica over the af range. The experimental results indicate that prebreakdown avalanches are absent or rare; the noise is apparently due to fluctuations in field-emission current.

621.315.615 846
Conduction and Breakdown in Hexane—W. B. Green. (*J. appl. Phys.*, vol. 26, pp. 1257–1264; October, 1955.) Report of an experimental investigation using field strengths up to 250 kv/cm.

MATHEMATICS

512.831:621.3.012.1 847
Matrix Representation of Complex Vectors—S. Mayr. (*Elektrotech. u. Maschinenb.*, vol. 72, pp. 460–464; October 1, 1955.) Technique useful for dealing with *e.g.* parallel-connected nonlinear resistors and nonlinear quadripoles is indicated.

MEASUREMENTS AND TEST GEAR

531.788:537.52:538.6 848
Methods for the Measurement of Low Gas Pressures by means of a Self-Maintained Gas Discharge in a Transverse Magnetic Field—R. Haefer. (*Acta phys. austriaca*, vol. 9, p. 200–215; August, 1955.) The gas pressure can be determined from measurements of (*a*) the value of magnetic induction at which ignition occurs for a given voltage, (*b*) the ignition voltage for a given magnetic induction, (*c*) the discharge current for given values of anode voltage, circuit resistance, and magnetic induction, or (*d*) the running voltage of the discharge for given values of the same parameters. Gases considered are argon, air and hydrogen. Use of the short cylindrical electrode arrangement previously described (2914 of 1954 and back references) has proved satisfactory. The pressure range from 10^{-1} to 10^{-8} Torr can be covered using one or more of these methods.

621.317.3:537.533 849
Automatic Instrument for Electron Scattering Measurements—L. Marton, J. A. Simp-

son, and T. F. McCraw. (*Rev. sci. Instrum.*, vol. 26, pp. 855–858; September, 1955.) Measurements are made of the distribution in energy and angle of electrons scattered from solids irradiated by primary electrons having energies of 10–50 kev. Energy peaks less than 20 ev apart at 20 kev and angular details of less than 10^{-3} radian can be resolved.

621.317.3:537.533.8 850
Measurement of Electron Energies by Deflection in a Uniform Electric Field—G. A. Harrower. (*Rev. sci. Instrum.*, vol. 26, pp. 850–854; September, 1955.) The parallel-plate analyzer described is designed for the study of secondary electrons having energies in the range 100–900 v, and will separate electrons differing in energy by 2 per cent.

621.317.3:537.533.8.08 851
Apparatus for the Measurement of Secondary-Emission Coefficients—A. Bobenrieth. (*Le Vide*, vol. 10, pp. 103–104; July/September, 1955.) Primary electrons are provided by a directly heated tantalum cathode, and the secondary electrons are collected on a screen enclosing the specimen under test. The apparatus is enclosed in a continuously evacuated glass tube from which the specimen may easily be withdrawn.

621.317.444 852
An Electronic Magnetometer—B. G. Cragg. (*J. sci. Instrum.* vol. 32, pp. 385–386; October, 1955.) An electron beam deflected by the magnetic field to be measured is automatically restored by an es potential which is proportional to the field for fields alternating at frequencies up to 10 kc. With the instrument described, the weakest field measurable is 10^{-4} oersted.

621.317.7 853
New Methods for the Construction of Shockproof Indicating and Recording Measurement Instruments—T. Staub. (*Bull. schweiz. elektrotech. Ver.*, vol. 46, pp. 837–840, 857–860; September 3, 1955.) The most important methods for testing shock-proof instruments are described. Tensioned-strip and pivoted mechanisms are compared as regards their shock resistance, as are also 90-degree and 250-degree deflection instruments.

621.317.72:621.396.812.3 854
Special Instruments for Observation and Analysis of V.H.F. Fading—K. Hirao and H. Maruyama. (*J. Radio Res. Labs, Japan*, vol. 2, pp. 207–216; April, 1955.) Description of the design of equipment suitable for the autocorrelation analysis of observations of fading of vhf signals due to tropospheric refractive-index variations, with a discussion of the necessary duration of observation periods and intervals between readings.

621.317.723 855
Vibrating-Capacitor Electrometer—M. Brière and J. Weill. (*J. Phys. Radium*, vol. 16, pp. 695–703; August/September, 1955.) Detailed description of an instrument capable of detecting dc down to 10^{-16} A or less and having very low zero drift.

621.317.725:621.375.2 856
Wide-Range Logarithmic Voltmeter—F. V. Hunt and J. F. Hersh. (*Rev. sci. Instrum.*, vol. 26, pp. 829–835; September, 1955.) A triode tube may be made to give a logarithmic output/input characteristic by including a diode rectifier in its grid circuit, increasing input then driving the tube towards cutoff. If the ac input signal is divided between the biasing diode and the triode to make the response curve symmetrical several stages may be cascaded without discontinuity. In the instrument described four such stages are used; on reaching cutoff each successive tube is clamped by a diode fed from the preceding stage, and the range of the voltmeter is approximately equal to the voltage gain of the cascade amplifier. The response is independent of frequency from 50 cps to 100 kc and is logarithmic over the range 1 mv to 10 v.

621.317.725.029.4:621.317.733 857
Expanded-Scale Voltmeter for A.C. Measurements—H. Galman. (*Electronics*, vol. 28, pp. 164–165; December, 1955.) Use of a hot wire in a bridge circuit gives an rms indication of the voltage to be measured. This is passed through a negative-feedback amplifier to a milliammeter. By use of a tapped input transformer a range of 100–500 v is obtained with a reading accuracy within 0.25 per cent. The instrument is independent of frequency from 50 cps to 2 kc.

621.317.742 858
Relationship between Eccentricities and Voltage Correction Factors for Concentric Slotted Lines—Y. Kita. (*J. Radio Res. Labs, Japan*, vol. 2, pp. 193–199; April, 1955.) The voltage correction factor for coaxial-line standing-wave detectors differs from unity by $2/(1-r^2)$ times the eccentricity, where r is the ratio of the radii of the inner and outer conductors.

621.317.742 859
Slotted-Section Standing-Wave Meter—E. M. Wareham. (*J. Brit. IRE*, vol. 15, pp. 539–564; November, 1955.) A detailed account is given of the theory and techniques of measurement, and some results are analyzed; the design of a particular instrument is discussed.

621.317.755 860
An Automatic Cathode-Ray-Oscilloscope Beam Brightening Device for Transient Recordings—J. Wood. (*J. sci. Instrum.* vol. 32, pp. 401–402; October, 1955.) By slight circuit alterations to a commercially available cro a brightening pulse of 70v is automatically applied to the grid of the tube. The time taken for the beam to reach full brilliance, after the triggering of the timebase by a transient, is brought down to about 1 μs by setting the ordinary brightness control so that the spot is just visible in the absence of a transient.

621.317.755 861
Large-Screen Curve Tracer—R. Duchamp. (*Télévision*, pp. 249–253; October, 1955.) Details are given of a cro using a standard 43-cm television cr tube with magnetic deflection. The frequency range covered is 1–250 mc, with a frequency sweep variable between 1 and 40 mc. The vertical scale is linear or logarithmic, and the horizontal scale is in mc.

621.317.755 862
Sawtooth Pulser gives Voltage/Current Curves—N. Sclar and R. L. McFolin. (*Electronics*, vol. 28, pp. 156–157; December, 1955.) A thyratron pulse generator with external trigger enables the voltage/current characteristics of circuit elements to be displayed on a cro. The load circuit includes a series resistance small compared with that of the specimen; the voltage across this gives a measure of the current passed. Heating of the specimen is minimized by the use of a short test pulse and a low repetition rate.

621.373.421.12 863
Audio-Standard Generator—P. Koustas. (*Electronics*, vol. 28, pp. 161–163; December, 1955.) "Simultaneous output voltages are available at frequencies of 500 cps, 1, 1.5, 2, 3, 5, and 10 kc from unit employing tuning-fork standard. Technique involves full-wave rectifiers as doublers to achieve desired frequencies."

621.374 864
Precision Digital Delay Generator—W. Perzley. (*Electronics*, vol. 28, pp. 148–151; December, 1955.) A crystal-controlled 1-mc pulse generator is followed by two cascaded divide-by-ten circuits, providing three ranges of pulse frequency. The output from these is passed through a ten-stage binary counter and pulses marking any required interval are selected by flip-flop controlled diode gates. The pulse timing is independent of flip-flop transition time.

621.385.001.4 865
The Rating of Thermionic Valves for use under Abnormal Ambient Conditions—Mills and Wright. (See 937.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.767:621.396.96 866
How Accurate are Radar Speed Meters?—J. Q. Brantley, Jr. (*Electronics*, vol. 28, pp. 132–134; December, 1955.) Tests on a type of motor-vehicle speed meter depending on the measurement of the Doppler shift in reflected signal frequency are described. Most of the possible errors are shown to result in speed indications which are too low.

539.32:534.2–8 867
Measurement of the Elastic Constants of Solids, using Ultrasonics—G. Mayer and J. Gigon. (*J. Phys. Radium*, vol. 16, pp. 704–706; August/September, 1955.) A cylinder of the material under test is suspended in an ultrasonic field of continuously variable frequency, produced by an ionophone; the frequency is varied until the specimen resonates, this condition being detected, *e.g.*, by arranging a graphite or Ge point in light contact with the specimen and completing the circuit.

621–52 868
Automation—a Survey—R. J. Bibbero. (*Elect. Engng., N.Y.*, vol. 74, pp. 775–780; September, 1955.) Historical and current aspects of development are discussed generally.

621.317.083.7(47):016 869
List of Russian and Translated Literature on Telemetering for 1950–1954—V. P. Demeshin, I. A. Kostetskaya, A. I. Novikov, N. V. Pozin, and V. A. Kashirin. (*Avtomatika i Telemekhanika*, vol. 16, pp. 409–410; July/August, 1955.)

621.317.39 870
Ice Detector for Lighter-than-Air Craft—E. G. Thurston. (*Electronics*, vol. 28, pp. 142–144; December, 1955.) A piezoelectric Meacham-bridge oscillator system is used to determine the amount of ice or other deposit on the surface of balloons, etc., by cementing the piezoelectric element to the inside surface of the fabric so that the ice formation adds to its effective mass.

621.383.2:535.376 871
Realizable Light Gain in Photoelectronic Image Intensifiers—L. Mandel. (*J. sci. Instrum.*, vol. 32, pp. 405–406; October, 1955.) By combining the most efficient photoelectric and luminescent surfaces an effective light gain of the order of 100 should be realizable. For photographic purposes two intensifiers in cascade, with optical focusing of the beam incident on each photocathode and on the photographic plate, should give a gain of about 115, with possibilities of further improvement.

621.384.6 872
Radiation Damping in Particle Accelerators with Circular Focusing Guiding Field—W. Humbach. (*Z. Naturf.*, vol. 10a, pp. 347–348; April, 1955.)

621.384.611/.612 873
The Extraction of the Beam from the Liverpool Synchrocyclotron—(*Proc. roy. Soc. A*, vol. 232, pp. 236–251; October 25, 1955.)

Part 1—Theoretical. K. J. Le Couteur (pp. 236–241). The beam is extracted by a magnetic deflector designed to produce radially unstable orbits while maintaining vertical stability; the theoretical specification of the apparatus is given.

Part 2—Experimental Work. A. V. Crewe and J. W. G. Gregory (pp. 242–251). In the final result, 3 per cent of the circulating beam was extracted from the cyclotron tank. The beam energy is 383 mev, with no measurable energy spread. By means of an auxiliary magnet the beam has been focused to give a spot in which the proton flux density is at least 1,000 times greater than that obtained from any similar machine.

621.384.612 874
Three-Hundred-MeV Nonferromagnetic Electron Synchrotron—W. B. Jones, H. R. Kratz, J. L. Lawson, D. H. Miller, R. D. Miller, G. L. Ragan, J. Rouvina, and H. G. Voorhies. (*Rev. sci. Instrum.*, vol. 26, pp. 809–826; September, 1955.) Description of an instrument in which electrons are injected at an energy of 100 kev, accelerated to 4 mev by betatron action and then to to 300 mev by synchrotron action. The magnetic guiding fields for the betatron and synchrotron operation are produced by suitable coils.

621.384.612 875
The Accuracy of Alignment required for a Strong-Focusing Synchrotron—W. Humbach. (*Z. angew. Phys.*, vol. 7, pp. 423–427; September, 1955.) Analysis for an alternating-gradient synchrotron shows that a statistical distribution of alignment errors causes the particles to diffuse from the prescribed orbit; the tolerances on alignment are governed by the permissible loss of particles due to this diffusion.

621.384.613 876
A 20-MeV Betatron for X-Ray Therapy—D. Major, F. R. Perry, and K. Phillips. (*Proc. IEE*, Part A, vol. 102, pp. 845–856; December, 1955.) Full description of equipment installed at Manchester.

621.384.613:620.179.1 877
A 15-MeV Betatron for the Nondestructive Testing of Materials—R. Schittenhelm. (*Arch. tech. Messen*, Nos. 236, 237 and 239, pp. 205–208; September, 225–228; October, and 275–276; December, 1955.) Design details are given; emphasis is on low magnet weight and high electron-beam intensity.

621.384.622 878
Linear Acceleration of Charged Particles to High Energies—C. W. Miller. (*Engineering, Lond.*, vol. 180, pp. 340–343; September 9, and 374–377; September 16, 1955. *Metrop Vick. Gaz.*, vol. 26, pp. 373–387; December, 1955.) An account is given of the development of linear accelerators, some post-war models are described and possible applications are reviewed. 45 references.

621.384.622.2 879
45-MeV Medical Linear Electron Accelerator—C. L. Hsieh. (*Elect. Engng., N.Y.*, vol. 74, pp. 790–795; September, 1955.) Detailed description of a machine developed for a Chicago hospital.

621.385.833 880
The Field Distribution in Asymmetrical Magnetic Electron Lenses—G. Liebmann. (*Proc. phys. Soc.*, vol. 68, pp. 679–681; September 1, 1955.) The problem was investigated by the resistance-network analog method (1954 of 1950). A series of typical characteristic curves is given.

621.385.833 881
The Magnetic Pinhole Electron Lens—G. Liebmann. (*Proc. phys. Soc.*, vol. 68, pp. 682–685; September 1, 1955.) An asymmetrical lens is considered with bore radii R_0 and R_1, where R_1 is very small compared with R_0. The focal length is not greatly altered, but the spherical abberation may be considerably increased by the presence of the pinhole.

621.385.833 882
A Unified Representation of Magnetic Electron Lens Properties—G. Liebmann. (*Proc. phys. Soc.*, vol. 68, pp. 737–745; October 1, 1955.) It is shown that many magnetic electron lenses, both symmetrical, can be represented by a single focusing curve.

621.385.833 883
Experimental Investigation of Chromatic Aberration in the Electron Microscope—S. Katagiri. (*Rev. sci. Instrum.*, vol. 26, pp. 870–873; September, 1955.) Experimental results indicate that chromatic field aberration can be completely compensated by a proper combination of two magnetic lenses. Micrographs with a resolution of about 5 mμ can be obtained with accelerating-voltage fluctuations as great as 1 per cent.

621.385.833 884
On Electron Mirror Microscopy—L. Mayer. (*J. appl. Phys.*, vol. 26, pp. 1228–1230; October, 1955.) A brief indication is given of the potentialities of electron-microscope systems in which the field in front of the specimen reflects the exploring beam. A resolving power comparable to that of the optical microscope has been obtained.

621.387.424 885
Behaviour of Externally Graphited Glass-Walled Geiger-Müller Counters exposed to Intense γ Radiation—D. Blanc. (*J. Phys. Radium*, vol. 16. pp. 681–687; August/September, 1955.)

621.387.464 886
The Physics of the Scintillation Counter—G. F. J. Garlick. (*J. sci. Instrum.* vol. 32, pp. 369–371; October, 1955.) Factors limiting the performance of scintillation counters used as energy spectrometers are discussed. Spread in output-pulse magnitude when detecting monoenergetic particles results from statistical fluctuations in the photomultiplier cathode; in practice other variables can also contribute. The problem of nonlinear variation of phosphor response is treated.

621.387.464 887
Problems of Large-Area Scintillation Counters—F. X. Roser. (*Acta phys. austriaca*, vol. 9, pp. 297–323; August, 1955.) The various, sometimes conflicting, requirements for optimum resolving power are discussed. A method involving masking of the central region of the photomultiplier cathode has given good results. 102 references.

621.387.464 888
Portable Scintillation Counter for Applications in Geology—H. Schneider. (*Z. angew. Phys.*, vol. 7, pp. 413–416; September, 1955.) The probe unit of the instrument described comprises a NaI/Tl crystal of diameter 30 mm and thickness 15–20 mm in combination with a German Type-FS9-A photomultiplier; the amplifier, batteries etc. are in a separate case. Sensitivity is ten times that of a G-M counter with the same active area.

771.36:[621.382.2+621.317.755 889
A [camera-] Shutter Tester using a Photoelectric Integrator—R. J. Hercock and D. M. Neale. (*J. Brit. IRE*, vol. 15, pp. 565–575; November, 1955.) An instrument comprising a photoelectric integrator together with a single-sweep cro is described.

534–8:62 890
Ultrasonic Engineering [Book Review]—A. E. Crawford. Publishers: Butterworths Scientific Publications, London, 344 pp. (*J. Electronics*, vol. 1, pp. 355–356; November, 1955.) "... should prove of considerable help to the acoustics engineers interested in the effects and applications of ultrasonic waves."

PROPAGATION OF WAVES

621.396.11 891
Ground-Wave Propagation over a Non-homogeneous Earth—K. Ventkitaraman. (*J. Instn. Telecommun. Engrs., India*, vol. 1, pp. 155–171; September, 1955.) The relative merits of known methods of calculating the ground-wave field strength are assessed by reference to measurement results. The superiority of Millington's method is confirmed.

621.396.11:551.510.535 892
The Physics of the Ionosphere—(See 755.)

621.396.11:551.510.535 893
The Influence on the Recordings of Ionosphere E-Layer Soundings of the Separation Level of Regions of Quasi-longitudinal (Q.L.) and Quasi-transverse (Q.T.) Propagation—D. Lepechinsky. (*C.R. Acad. Sci., Paris*, vol. 241, pp. 897–900; October 3, 1955.) Continuation of previous discussion [1767 of 1955 (Lepechinsky and Durand)]. Conditions are considered for the four cases when the *E*-layer maximum-ionization level is (*a*) above, (*b*) coincident with, (*c*) a little below, and (*d*) considerably below the separation level. An indication is given of the nature of the *h'f* traces to be expected and of the significance of the various conditions for the evaluation of the maximum *E*-layer ionization from the f_0 value. Recordings made at Poitiers are reproduced showing absence of trace corresponding to case *b* and the inflection point corresponding to case *c*.

621.396.11:551.510.535 894
Characteristics of F_2-Layer Multiple Reflections (10–16 Times)—Y. Echizenya, S. Katano, and Y. Ogata. (*J. Radio Res. Labs., Japan*, vol. 2, pp. 137–141; April, 1955.) Routine vertical-incidence ionospheric soundings during 1953 and 1954 show that high-order multiple reflections occur only at night and are confined to the ordinary ray. The frequency of occurrence of the phenomenon is high in spring and autumn and low in summer and winter and shows marked correlation with the noon value of f_0F_2.

621.396.11:551.510.535 895
Magneto-Ionic Triple Splitting over Delhi—S. N. Mitra. (*J. Instn. Telecommun. Engrs., India*, vol. 1, pp. 124–129; September, 1955.) Records of triple splitting obtained on six occasions between 1951 and 1955 are discussed with particular reference to the low geomagnetic latitude of Delhi. The splitting may be caused by longitudinal propagation of the ordinary ray, associated with an increase in the collision frequency in the ionosphere.

621.396.11:551.510.535 896
Observations of Short-Wave Back-Scatter with Commercial Telegraphy Signals—B. Beckmann and K. Vogt. (*Fernmeldetech. Z.*, vol. 8, pp. 473–481; September, 1955.) Observations were made using the "ring" method, in which the back-scattered signal is transposed to IF and applied with 90-degree phase difference to the vertical and horizontal plates of a cro so that signal amplitudes are displayed as rings. Results show that the azimuthal distribution of back-scatter is governed by ionospheric conditions and by the radiation patterns of the transmitting antenna; when a directional antenna is used the main direction from which energy is received is that theoretically expected. The back-scatter mainly uses the great-circle path. The time variation of the back-scatter intensity agrees with that for ordinary propagation in the same wave-band. Determinations of the dead zone can be made from the variation of back-scatter transmission time with frequency.

621.396.11.029.62 897
On the Relationship between the Hourly Variation of Field Strength and the Structure of the Lower Atmosphere—K. Tao. (*J. Radio Res. Labs, Japan*, vol. 2, pp. 181–191; April, 1955.) Field-strength observations made on 60-mc signals transmitted over a distance of 125 km are correlated with meteorological data.

621.396.11.029.63/64 898
Experimental Studies on Diffracted Waves from a Mountain at 3000 Mc/s—T. Kono, K. Nishikori, M. Fukushima, M. Ikeda, and N. Yoshida. (*J. Radio Res. Labs., Japan*, vol. 2, pp. 163–180; April, 1955.) Continuous photographic records were made of the reception of pulsed transmissions over distances of 234 km and 166 km across a mountain 1000 m high. The measurements were made in September and December, 1954, respectively. For the first path, with the receiving station situated 90 km beyond the mountain, the observed field strength agreed approximately with that calculated by Fresnel's knife-edge-diffraction theory; for the second path, with the receiver 35 km beyond the mountain, the field strength was about 10 db below the calculated value.

621.396.11.029.64:621.396.96:551.578.4 899
The Back-Scattering Coefficient of a Spherical Homogeneous Mixture of Ice and Air at Wavelengths between 1 and 10 Centimeters—J. C. Johnson. (*J. Met.*, vol. 12, pp. 188–189; April, 1955.)

621.396.812.3:551.510.535 900
Three Components of the Field Strength of the Wave Reflected from the Surface of the Ionosphere, their Level and Time Variation—H. Uyeda, Y. Ogata, K. Uchikura, Y. Arima, and H. Obayashi. (*J. Radio Res. Labs., Japan*, vol. 2, pp. 143–161; April, 1955.) Tentative theory is developed to account for observed variations in received field strength of 4-mc signals. Formulas are derived for various configurations of the ionospheric reflecting surfaces, considered (*a*) at rest and (*b*) moving with constant horizontal velocity.

'621.396.812.3:621.317.72 901
Special Instruments for Observation and Analysis of V.H.F. Fading—Hirao and Maruyama. (See 854.)

RECEPTION

621.374 902
Frequency of Coincidence of Two Sets of Recurrent Pulses—H. Rakshit and S. C. Mukherjee. (*J. Instn. Telecommun. Engrs., India*, vol. 1, pp. 130–135; September, 1955.) Conditions are analyzed for narrow and for broad pulses.

621.396.62:621.372.543.2:621.396.822 903
Bandwidths of Various Filter Circuits and C.C.I.R. Recommendations on Effective Noise Bandwidths—E. Henze. (*Fernmeldetech. Z.*, vol. 8, pp. 512–515; September, 1955.) According to the C.C.I.R. recommendation, effective noise bandwidth of receivers is the width of a rectangle having the same area as the actual selectivity curve. The ratio of the effective noise bandwidth to the actual bandwidth is calculated for several practical selective circuits.

621.396.621:621.396.822 904
The Effect of a Random Noise Background upon the Detection of a Sinusoidal Signal—H. S. Heaps. (*Canad. J. Phys.*, vol. 33, pp. 509–520; September, 1955.) Analysis is presented for the case when the signal envelope exhibits negligible fluctuations. The effect of large signal fluctuations is demonstrated by comparing these results with those obtained previously (2099 of 1955).

621.396.82:551.594.6:621.317.3 905
Atmospheric Noise Interference to Broadcasting in the 5-Mc/s Band at Poona—K. R. Phadke. (*J. Instn. Telecommun. Engrs., India*, vol. 1, pp. 136–146; September, 1955.) Measurements of atmospheric noise interference were made during the hours 1800–2300 I.S.T., using Aiya's method (257 of 1955). The results are used to discuss noise values for satisfactory broadcasting service, in comparison with Radio Research Board and Central Radio Propagation Laboratory estimates. Noise levels estimated from lightning-discharge data [3263 of 1955 (Aiya)] are in close agreement with the measured values.

621.396.828 906
Reducing Radio Interference—(*Elect. Times*, vol. 128, p. 352; September 8, 1955.) Note on a new Code of Practice CP 1006:1955, published by the British Standards Institution, on the general aspects of radio-interference suppression in the medium and long-wave bands as well as the 41–61-mc television band. Requirements and tests for suppressor components and complete filter units are covered by the revised British Standard 613:1955.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.1 907
Theory of Pre-correction of Transmission Errors—B. Mandelbrot. (*Ann. Télécommun.*, vol. 10, pp. 122–134; June, 1955.) Feinstein's theorem (858 of 1955) is discussed and practical methods of pre-correction are derived.

621.395.44:621.315.212 908
Coaxial-Cable Carrier-Current Systems—J. Bauer and J. Valloton. (*Tech. Mitt. schweiz. Telegr.-Teleph Verw.*, vol. 33, pp. 337–368; September 1, 1955. In German and French.) Detailed description of the Swiss system providing 960 telephony channels in the frequency band 60 kc–4.028 mc.

621.396.1 909
Future of European Broadcasting—G. H. Russell. (*Wireless World*, vol. 62, pp. 18–22; January, 1956.) Failures of the Copenhagen plan (832 of 1949) are discussed and a suggested new plan is outlined providing 15 long-wave channels between 150 and 285 kc and 121 medium-wave channels in four bands between 525 kc and 1.602 mc. Total power would be allocated to each country on an area basis.

621.396.41:551.510.52 910
Designing Over-Horizon Communications Links—D. Davidson and A. J. Poté. (*Electronics*, December, 1955.) Existing knowledge on long-range tropospheric propagation is summarized, with references to original papers, and the design considerations necessary to achieve a system having a given reliability are discussed. See also 263 of January (Mellen et al.).

SUBSIDIARY APPARATUS

621-526:621.372.5 911
The Response Functions and Vector Loci of First and Second Order Systems—Morris. (See 690.)

621.311.6:621.314.67 912
Reliable Power-Pack Design—A. F. Standing. (*Electronic Engng.*, vol. 27, pp. 532–535; December, 1955.) "A method is described which reduces the design of capacitor-input power packs to simple calculations aided by graphs."

621.316.722:[621.375.23+621.373.421 913
A Constant-Voltage Amplifier and Oscillator—G. N. Patchett. (*Electronic Engng.*, vol. 27, pp. 536–539; December, 1955.) The equipment is based on the use of a thermistor bridge to regulate feedback and was designed for a frequency of 1 kc. The amplifier output voltage is constant to within 0.1 per cent for an input-voltage range of 0–4 v, with load varying from zero to 10 w or supply voltage varying from 130 to 250 v. Used as an oscillator, the equipment gives an output-frequency change of 1 cps for a supply-voltage change from 180 to 250 v when operating on full load, and <3 cps for maximum load variation at a given supply voltage.

621.318.435:621.316.722.076.25:621.372.54 914
Stabilization of Random Voltage Fluctuations by Ferroresonant [saturated-choke] Stabilizer—A. N. Malakhov. (*Bull. Acad. Sci. U.R.S.S., tech. Sci.*, no. 7, pp. 3–8; July, 1955. In Russian.) An investigation is reported of the transfer of voltage fluctuations by a stabilizer comprising an air-cored choke (L), in series with the input, and a saturated choke (D) in parallel with the load. The calculated and experimentally determined transfer-coefficient frequency characteristics presented in Figs. 3 and 4, respectively, show (*a*) that there is a maximum at a frequency $\Omega_0<\omega$, where ω is the supply frequency, and (*b*) that the attenuation is high only for fluctuations with angular frequencies $\Omega \ll \Omega_0$.

621.319.339 915
High-Voltage Generator of van de Graaff Type with Liquid Charge Carrier—K. Janner, S. Magun, and E. Schopper. (*Z. angew. Phys.*, vol. 7, pp. 446–450; September, 1955.)

TELEVISION AND PHOTOTELEGRAPHY

621.397.2:621.375.2 916
The "Chasseral"-Type Video Distribution Amplifier—H. A. Laett. (*Tech. Mitt. schweiz. Telegr.-Teleph Verw.*, vol. 33, pp. 369–374; September 1, 1955. In German and French.) The amplifier is designed to suit both 75-Ω and 150-Ω coaxial cable; the bandwidth is 10 mc and the over-all gain is unity with a reserve of 10 db. Low-frequency equalization is effected by a series combination of integrating and differentiating *RC* networks. There are two separate feedback loops, one with a level frequency response and the other with a rising frequency response. Four parallel output stages are provided, using pentodes with negative current feedback.

621.397.5:535.623 917
Colour Television—B.B.C. Experiments—(*J. Brit. IRE*, vol. 15, pp. 576–581; November, 1955.) Details are given of the equipment and modified NTSC signal used.

621.397.5:778.5 918
The Suppressed-Frame System of Telerecording—C. B. B. Wood, E. R. Rout, A. V. Lord, and R. F. Vigurs. (*B.B.C. Engng. Div. Monographs*, no. 1, pp. 1–14.) Description of apparatus comprising a high-grade television monitor [1753 of 1953 (Lord and Wood)] together with a commercially available 35-mm motion-picture camera.

621.397.5:778.5 919
16-mm Telerecording for Sequential Television Systems—V. B. Hulme. (*Electronic Engng.*, vol. 27, pp. 516–522; December, 1955.) "A system of telerecording using continuous film motion through a shutterless gate is described. The camera motor speed is controlled by means of negative feedback components of phase, velocity, and acceleration. A dc generator and pulse generator driven by the motor together with a phase-discriminator provide these components in the form of voltages." The picture is presented line by line on the screen of the cr tube, the frame scan being eliminated.

621.397.5(083.74) 920
Television Waveform—(*Wireless World*, vol. 62, p. 26; January, 1956.) The black level in the British 405-line waveform has been lifted by 5 per cent of peak white amplitude and the suppression period preceding the synchronizing

signal has been lengthened by 0.5 μs compared with the previous standards (2912 of 1952).

621.397.5(494):621.396.11 921

Television in Switzerland—(*Wireless World*, vol. 62, pp. 33–35; January, 1956.) Service to thickly populated areas is provided by three stations working on frequencies in band I and a fourth using band III. The stations are sited at altitudes of 2,000–5,000 feet and the mean height of towns served is about half that of the corresponding transmitter. The transmitters are linked by a radio-relay system working on 2 kmc. A map shows the effective coverage of one of the transmitters. See also 277 of 1955 (Gerber).

TRANSMISSION

621.396.61:621.372.54 922

High-Frequency Filters and Tuning of Amplitude-Modulated Transmitters—R. Guertler. (*Telefunken Ztg.*, vol. 28, pp. 116–123; June, 1955. English summary, pp. 134–135.) Transmitter output stages are discussed comprising coupled tuned circuits of which the primary one, in the output-tube anode circuit, is parallel resonant while the secondary is series resonant. A frequency response symmetrical with respect to carrier frequency, and hence absence of unwanted phase modulation, is attained by adjusting the secondary rather than the primary circuit to make the anode impedance purely resistive. Locus diagrams indicate the circuit operation.

TUBES AND THERMIONICS

621.314.63:546.28 923

Zener-Voltage Breakdown Uses in Silicon Diodes—C. N. Wulfsberg. (*Electronics*, vol. 28, pp. 182, 192; December, 1955.) For back voltages less than the Zener voltage the back resistance of a Si junction diode may exceed $10^{10}\Omega$. The Zener voltage may be made to have any value from 3 v to several hundred volts. A wide variety of applications is discussed.

621.314.7 924

Junction Transistor Electronics—J. L. Moll. (Proc. IRE, vol. 43, pp. 1807–1819; December, 1955.) A review paper. The mechanism of current transport at semiconductor junctions is described and parameters of junction transistors are hence derived. Amplifying, oscillating, and switching performance is discussed. Replacement rates as low as one per 20,000 transistor-hours are reported.

621.314.7 925

Temperature Variation of the "Punch-Through" Voltage of a Transistor—O. Garreta. (*C.R. Acad. Sci., Paris*, vol. 241, pp. 857–859; October 3, 1955.) The variation of the "punch-through" voltage [3329 of 1953 (Dacey)] over the temperature range 14°–300°K was studied for *p-n-p* transistors. The potential distribution through the transistor exhibits its maximum value at a point within the base distant x_m from the emitter junction, where $x_m = T^{1/2}\phi(z_m)$ and z_m is a dimensionless parameter. A formula is derived indicating that the value of the punch-through voltage at any temperature is equal to the value at zero temperature less a term proportional to x_m. This formula is verified by experimental results on three Ge transistors. By extrapolating to zero temperature the thickness of the base can be determined exactly.

621.314.7:546.28 926

Recent Developments in Silicon Fusion Transistors—R. A. Gudmundsen, W. P. Waters, A. L. Wannlund, and W. V. Wright (Trans. IRE, vol. ED-2, pp. 74–81; January, 1955.)

621.314.7:621.375.4 927

Some Properties and Circuit Applications of Super-Alpha Composite Transistors—A. R. Pearlman. (Trans. IRE, vol. ED-2, pp. 25–43; January, 1955.) Circuits are discussed in which two junction transistors are interconnected in such a way that the gain of the combination is increased; α values >0.99 can be achieved. The properties of such arrangements are in many ways equivalent to those of a vacuum triode. Voltage-follower and voltage-amplifier circuits using composite transistors are described. Characteristics of two- and three-stage composite transistors are shown graphically.

621.314.7:621.375.4 928

The Equations and the Equivalent Circuit of the Transistor—M. Skalicky. (*Elektrotech. u. Maschinenb*, vol. 72, pp. 422–423; September 1, 1955.) Analysis leads to the development of a passive-quadripole representation of the transistor.

621.383.27:621.396.822 929

Hysteresis Effect in Multiplier Phototube Noise—C. A. Ziegler and H. H. Seliger. (*J. appl. Phys.*, vol. 26, pp. 1225–1227; October, 1955.) Measurements of the variation of noise with time and with operating parameters are reported for two types of photomultiplier. Possible explanations of the observed effects are discussed.

621.383.27.032.21 930

Dark Current of Secondary-Electron Multipliers with (Ag) – Cs_2O, Cs – Ag Photocathodes—F. Eckart. (*Ann. Phys., Lpz.*, vol. 16, pp. 322–330; September 20, 1955.) The temperature dependence of the thermionic emission of the cathode was investigated over the temperature range of about 20°C–40°C. At 19.7°C the dark current at the anode was 3.2×10^{-8}A with an amplification of 1.4×10^{6}; this corresponds to a thermionic emission of 10^{-16} A/cm². The work function, calculated from the specific-emission/reciprocal-temperature curve, is 0.95 ev, that calculated from the cut-off frequency of the external photoeffect is 1.06 ev. The construction and manufacture of the photomultiplier is described in detail.

621.383.4:535.37:621.318.57 931

Opto-electronic Devices and Networks—E. E. Loebner. (Proc. IRE, vol. 43, pp. 1897–1906; December, 1955.) "The transducing properties of electroluminescent and photo-responsive cells are described. The light amplifying and spectrum-converting characteristics of a circuit consisting of an electric power supply and a series combination of the two types of cells whose impedances have been matched are discussed. The construction and operation of an optoelectronic bistable device—the "optron"—employing positive radiation feed-back, is reported. The optron is both a storage cell and a switch with dual (optical and electrical) signal input and output. Numerous logic networks, composed of electric series and parallel combinations of electroluminescent and photoconductive cells with selected optical couplings have been designed, constructed, and operated."

621.383.4:535.371.07 932

Theory and Experiments on a Basic Element of a Storage Light Amplifier—J. E. Rosenthal. (Proc. IRE, vol. 43, pp. 1882–1888; December, 1955.) Theory is developed to explain the operation of solid-state image intensifiers of the type described, *e.g.* by Orthuber and Ullery (3061 of 1954), in which a layer of a photoconductor such as CdS is used to control the voltage across an electroluminescent element. Optimum values of operating parameters deduced are in good agreement with experimental results.

621.383.4:535.371.07 933

An Electroluminescent Light-Amplifying Picture Panel—B. Kazan and F. H. Nicoll. (Proc. IRE, vol. 43, pp. 1888–1897; December, 1955.) The construction of large-area solid-state image intensifiers of the general type described by Orthuber and Ullery (3061 of 1954) is made feasible by providing the photoconductive CdS in powder form [612 of February (Nicoll and Kazan)]; the powder may be supported on a mesh or laid in grooves. Using an excitation frequency of *e.g.* 400 cps, good resolution of half-tone pictures has been obtained with panels 12-inch square. Response time is 0.1 second or more; other photoconductive materials may be found to give quicker response.

621.383.4:537.311.33:535.215 934

Lead Salt Photoconductors—Moss. (See 773.)

621.383.5:[621.314.63+621.314.7 935

Photodiodes and Phototransistors considered as Infra-red-Radiation Detectors—G. A. Boutry and F. Desvignes. (*Nuovo Cim.*, Supplement to vol. 2, pp. 541–563; 1955. In French.) A Comprehensive survey with 59 references.

621.385. 936

Some Factors affecting Transmitting Valve Life—T. N. Basnett. (*J. Brit. IRE*, vol. 15, pp. 588–592; November, 1955.) Reprint. See 3717 of 1954.

621.385.001.4 937

The Rating of Thermionic Valves for use under Abnormal Ambient Conditions—B. D. Mills and W. W. Wright. (*J. Electronics*, vol. 1, pp. 276–292; November, 1955.) Manufacturers published ratings may need revision if tubes are to be operated in either very high or very low ambient temperatures. An interim report is presented of a program of life tests designed to determine suitable maximum temperatures for various types of tube. The procedure adopted is to set the ambient conditions as desired and to measure the "hot-spot" bulb temperatures for a range of input anode powers. Results are presented graphically, the variables being the total input power, the "hot-spot" temperature, the ambient temperature, and the equivalent altitude. Simple tests can give enough information to evolve a whole family of rating curves.

621.385.029.6 938

Theory of the Pre-oscillating Magnetron: Part 2—Perturbations of a Double-Stream Steady State—G. D. Sims and D. Gabor. (*J. Electronics*, vol. 1, pp. 231–262; November, 1955.) An attempt is made to determine theoretically the dispersion relation for oscillations in a cutoff magnetron on the assumption of a double stream in the steady state, and of a particular distribution of electron velocities. A perturbation method is used; the electron density of the perturbed cloud is deduced for the case of a finite cathode temperature. The implications of the results in relation to magnetron starting, back-bombardment and noise spectrum are discussed. Part 1: 3787 of 1955 (Gabor and Sims).

621.385.029.6 939

Comments on Magnetron Theory, with Particular Reference to some Recent Publications—O. Buneman. (*J. Electronics*, vol. 1, pp. 314–323; November, 1955.) Recently published papers are criticized as being unduly preoccupied with symmetrical steady states and indicating possible ignorance of wartime development work on magnetrons. The question is posed, whether the magnetron is worth the intense theoretical effort required to solve the problems involved.

621.385.029.6:537.533 940

Energy Distribution of Electrons in Beams with Strong Space Charge—W. Veith. (*Z. angew. Phys.*, vol. 7, pp. 437–443; September, 1955.) Experiments with specially-constructed traveling-wave tubes show that if the beam is retarded between the end of the helix and the collector electrode the range of electron energies is very greatly extended. The effect is attributed to a redistribution of energy occurring as the beam traverses the retarding path; the extra-high energy values increase with increasing length of the retarding path. The explanation is relevant also to phenomena observed with hollow cathodes.

621.385.029.6:537.533:621.386 941
An X-Ray Method for studying Radial Current Distributions in Electron Beams—J. S. Thorp. (*Brit. J. appl. Phys.*, vol. 6, pp. 366–368; October, 1955.) The cathode producing the beam to be studied is mounted in a tube whose anode is a thin foil forming part of the envelope. Soft x-rays, generated by the impact of electrons on the foil, are transmitted and produce an image of the transverse section of the beam on a photographic film pressed against the outer surface of the foil. The method has applications in the development of klystrons, etc.

621.385.029.6:621.372.2 942
Propagation in Linear Arrays of Parallel Wires—J. R. Pierce. (TRANS. IRE, vol. ED-2, pp. 13–24; January, 1955.) Analysis is presented for propagation in periodic structures incorporating transverse wires, of the general type used in the traveling-wave tubes discussed by Leblond and Mourier (1204 of 1955) and Karp (1212 of 1955).

621.385.029.6:621.386 943
X-Ray Production by Magnetron—A. C. Wesley. (*Wireless Engr.*, vol. 33, p. 29; January, 1956.) Incidental production of soft x-rays by a Type-4J50 magnetron has been observed. A bend should be included in waveguide systems used with the magnetron, to act as an x-ray filter.

621.385.029.64 944
The Design and Performance of a High-Power Demountable Klystron Amplifier for X-Band—J. D. Lawson, R. S. Barton, T. F. Gubbins, W. Millar, and P. S. Rogers. (*J. Electronics*, vol. 1, pp. 333–354; November, 1955.) The design of a two-cavity klystron to give cw power of 1 kw and a gain of 10 at 9.375 kmc is discussed. Constructional techniques are described in detail. Performance is in fair agreement with basic theoretical predictions, except as regards the gain.

621.385.032.216 945
Thermionic Emission from Sintered Mixtures of Powdered Tungsten with Alkaline-Earth Carbonates—G. Mesnard and R. Uzan. (*Le Vide*, vol. 10, pp. 105–118; July/September, 1955.) The chemical reactions occurring during the heat treatment of cathodes formed from mixtures of powdered refractory metals with emissive powders are discussed; emissivity/temperature curves for various mixtures are given. The cathodes considered are superior to the usual types as regards sensitiveness to the presence of gas and ionic bombardment, and they can readily be reactivated. See also 3728 of 1954 (Uzan and Mesnard), etc.

621.385.032.216 946
The Effect of Zirconium on Sintered Alkaline-Earth Oxide Cathodes—G. Mesnard and R. Uzan. (*Le Vide*, vol. 10, pp. 124–134; July/September, 1955.) The addition of Zr as a reducing agent to the mixtures used in the formation of sintered cathodes is shown to have harmful effects on the emissivity. Heat treatment in an atmosphere of hydrogen gives bad results, favoring oxidation of the metals present. Among possible alternatives, the use of Ni gives best results. See also 3728 of 1954 (Uzan and Mesnard), etc.

621.385.2:621.396.822 947
The Rise of Noise Temperature in Space-Charge-Limited Diodes—H. W. König. (*Arch. elekt. Übertragung*, vol. 9, pp. 411–418; September, 1955.) Continuation of previous discussion (2794 of 1955). Calculations made taking account of electron-velocity fluctuations at the cathode and of the correlation between velocity and current fluctuations give results in improved agreement with observations. Very small irregularities in cathode structure may cause a reduction of the space-charge factor of several thousandths below the ideal value of unity, leading to a very large rise of noise temperature with increasing anode voltage. The theory presented enables the imaginary part of the correlation factor to be determined from experiments.

621.385.2.029.6 948
Electron Trajectories in Coaxial Diodes—R. Dehn. (*Wireless Engr.*, vol. 33, pp. 10–12; January, 1956.) Calculations have been made for diodes in which the electron transit time corresponds to several cycles at 3 kmc. Depending on the rf phase at which they are emitted, electrons may proceed to the anode or may return to the cathode either within a single rf cycle or in the course of several rf cycles. Experimental results indicate that the numbers of electrons reaching the anode and returning to the cathode are about equal.

621.385.832 949
A Theory of determining the Dynamic Sensitivity of Cathode-Ray Tubes at Very High Frequencies by Means of Fourier Transforms—E. F. Bolinder. (TRANS. IRE, vol. ED-2, pp. 44–50; January, 1955.)

621.385.832.032.2 950
The Electron Gun of the Cathode-Ray Tube—Tetrode or Triode?—J. A. Darbyshire. (*Electronic Engng.*, vol. 27, pp. 523–528; December, 1955.) The triode gun gives a smaller spot at the center of the screen and is therefore preferable for some applications, but for television purposes the tetrode gun is better since it has a narrower beam width in the region of the scan coils and hence better deflection defocusing characteristics, together with lower cathode loading.

621.387 951
Initiation of Hot-Cathode Discharges—N. R. Daly and K. G. Emeleus. (*Brit. J. appl. Phys.*, vol. 6, pp. 370–372; October, 1955.) Report of an experimental study of the effect of a voltage fall along a hot cathode on the initiation of a gas discharge in a diode at low pressure. Observations were made of the appearance of the discharge before and after the formation of plasma, and of the oscillations generated by the discharge.

621.387 952
Running-Voltage/Current Characteristics of some Glow-Discharge Tubes—F. A. Benson and G. Mayo. (*Electronic Engng.*, vol. 27, pp. 540–542; December, 1955.) Conclusions arrived at in a previous paper (3422 of 1954) concerning the important influence of the gas filling on the running-voltage/ambient-temperature characteristics of glow-discharge tubes are shown to apply also to running-voltage/current characteristics. Running-voltage drift and hysteresis effects also vary with the kind of gas used.

621.387:621.318.57 953
Response Times of Gas-Filled Valves for Switching Purposes in Communication Engineering—K. Braun. (*Fernmeldetech. Z.*, vol. 8, pp. 490–493; September, 1955.) Equipment for measuring firing time is described. For a coincidence thyratron with anode voltage 60 v and grid overvoltage 1v, a value of 10μs was found; for some cold-cathode triodes with anode voltage 180–250 v and starter overvoltage 5–10 v the value was 100μs.

621.387:621.385.3 954
Amplifier Action of Gas-Filled Triodes—A. Székely. (*Acta phys. austriaca*, vol. 9, pp. 258–266; August, 1955.) Continuation of experiments reported previously (599 of 1954). If it is assumed that the internal resistance of the gas-filled triode is determined by the resistance of the positive space-charge layer around the grid, the amplifier action can be described by means of the Barkhausen formula for the vacuum triode and the dependence of the amplification on the frequency and amplitude of the grid voltage and on the intensity of the auxiliary grid-anode discharge can be explained. High gain is obtainable only for frequencies below about 150 kc. Gain is constant only for grid alternating voltages below about 0.5–1 v; the value for the Type-4690 tube used is 44.

MISCELLANEOUS

061.3:[55+621.396.11 955
The XIth General Assembly of the International Scientific Radio Union (U.R.S.I.), held at The Hague, 1954—(*Onde élect.*, vol. 35, pp. 555–629; June, 1955.) Nine papers are presented summarizing the work of the assembly in the fields of standards, tropospheric and ionospheric propagation, noise, radio astronomy, circuitry, and general electronics. For a short report see *Nature, Lond.*, vol. 176, pp. 451–452; September 3, 1955.

061.3:621.3 956
Electronics and Television Convention: Proceedings [Book Notice]—Supplement to *Ricerca sci.*, Milan, 2 vols; 1954. (*J. Telev. Soc.*, vol. 7, p. 482; July/September, 1955.) Includes all the papers given at this international convention, in the original languages, with summaries in French, German, and English. The subjects discussed included magnetic, dielectric and semiconductor materials, electron optics, radar, servomechanisms, computers, and cybernetics.

061.6:621.3 957
The "Galileo Ferraris" National Electrotechnical Institute—P. Lombardi. (*Ricerca sci.*, vol. 25, pp. 1988–2047; July, 1955.) Report of activities for the two years 1951–1953. The subjects covered include properties of ferromagnetic materials, electroacoustics, electronics, radio engineering, and television.

539.1(44) 958
Nuclear Energy and its Industrial Applications: Part 1—Atomic Piles and Radioisotopes—(*Onde élect.*, vol. 35, pp. 783–945; October, 1955.) This issue comprises a group of papers providing a survey of developments of atomic energy in France, with descriptions of equipment and methods used.

621.3.002.2 959
Automatic Circuit Production—(*Wireless World*, vol. 62, p. 23; January, 1956.) Improvements to the E.C.M.E. machine [1913 of 1947 (Sargrove)] will reduce the preparatory period before a production run from three months to a few days.

621.3.002.2 960
Punched Cards control Job-Lot Assembly System—G. W. Gamble, C. J. Goodwin, and F. S. Feldheim. (*Electronics*, vol. 28, pp. 122–128; November, 1955.) An automatic assembly system for producing complete subassemblies of electronic equipment in batches of up to 20 is described. Etched wiring and dip soldering are used.