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PROCEEDINGS OF THE IRE®

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 44 — *November, 1956* — NUMBER 11

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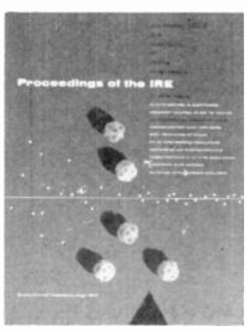
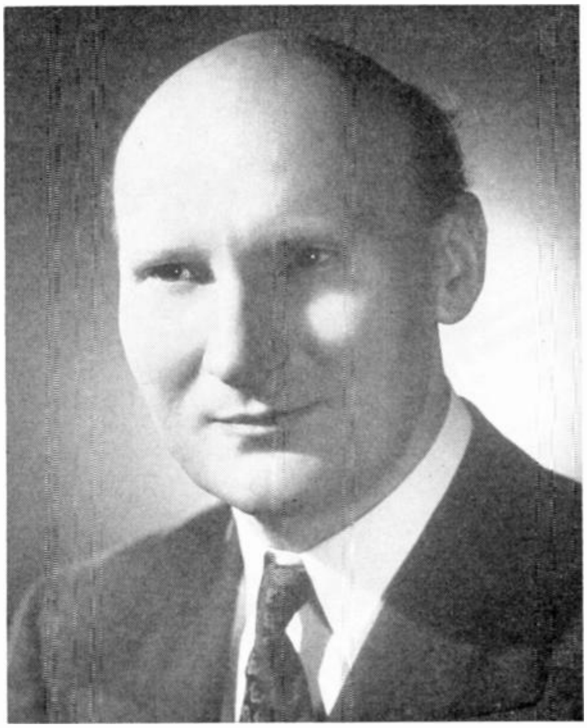
ADVERTISING SECTION



THE COVER—The layout suggests the ideas of randomness and variability—both important in the field of quality control. Twenty-sided dice, each carrying two sets of numbers from zero to nine, are used as a substitute for tables of random numbers to insure that sample units are selected at random. Variability, which is inherent in any manufacturing process, may be controlled, and often reduced, through the use of quality control techniques. An excellent survey of quality control methods starts on page 1521 of this issue.



Ernst Weber

DIRECTOR, 1955–1957

Ernst Weber was born in Vienna, Austria, on September 6, 1901, and received his education there. He was graduated with the diploma of electrical engineer from the Technical University in Vienna in 1924, and joined the Austrian Siemens-Schuckert Company as research engineer. On the basis of several papers on field theory applied to machinery, he received the degree of Sc.D. from the Technical University in Vienna in 1927.

While still working for his engineer's degree, Dr. Weber studied physics and mathematics at the University of Vienna, completing his Ph.D. in 1926 with a dissertation on the diffraction of light on submicroscopic spherical particles.

In January, 1929 he transferred to the Siemens-Schuckert Company in Berlin-Charlottenburg, where he was appointed lecturer at the Technical University. In the fall of 1930 he accepted an invitation as visiting professor to the Polytechnic Institute of Brooklyn, New York, where he has since remained. He accepted in 1931 the permanent position of research professor of electrical engineering in charge of graduate study. From 1942 to 1945 he was professor of graduate electrical engineering and head of graduate study and research in electrical engineering. From a few initial graduate courses offered only evenings, the graduate program developed into one of the largest in the country, adding to the master's program a complete and outstanding program leading to the doctor's degree in a combination of day and evening courses.

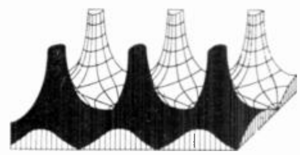
Early in the Second World War, Dr. Weber organized a group on microwave research in the electrical engineering department which expanded rapidly. In recognition of the contributions of the research group, he was awarded the Presidential Certificate of Merit in 1948. Out of this wartime research grew the Microwave Research Institute of Polytechnic Institute of Brooklyn, primarily engaged in projects under sponsorship of the military services; and the Polytechnic Research and Development Company, founded in 1944 and owned by Polytechnic Institute of Brooklyn.

Since 1945, Dr. Weber has been Head of the Department of Electrical Engineering and Director of the Microwave Research Institute of the Polytechnic Institute of Brooklyn. Since 1944 he has also been a Director, and since May, 1952 the President of the Polytechnic Research and Development Co., Inc.

He has published many scientific papers on electromagnetic fields, linear and nonlinear circuits, and microwave measurements; he contributed to several books, and published *Mapping of Fields* and *Linear Transient Analysis*. He is a Fellow of AIEE and the American Physical Society, and a member of the American Mathematical Society and others.

Dr. Weber joined the IRE as a Member in 1941 and attained Senior Member grade in 1943. He became a Fellow in 1951. He had been chairman of the Standards and Professional Groups Committees, the latter of which he is presently Eastern Vice-Chairman.

Poles and Zeros



Error. President M. S. Coover of the AIEE has written to point out that the professional society membership figures quoted in P and Z in the August issue were in error. He is entirely correct, and we hasten to correct the record. The second-largest professional society is not the IRE. It is the AIEE. We took the figures from an independent source without realizing that these figures omitted the AIEE student membership, while taking full credit for IRE students. As of April 30, 1956, the correct AIEE figures are 49,949 members, 9,076 student members and 382 student affiliates, total 59,407. As of the same date the IRE membership, including students, was 49,757.

We sincerely regret this error, particularly since it has been construed as contributing to the spirit of excessive competition which, in some quarters at least, afflicts the two societies. Such competition, while perhaps inevitable between associations covering overlapping fields, raises the stature of neither. The many thousands of engineers who are members of both societies (including 14 of the IRE Directors) can well take pride in the fact that IRE and AIEE, with a combined membership of over 100,000 engineers, are together handling the affairs of the electrical profession in this country completely and competently.

While decrying the competitive spirit to which we unwittingly added fire, we cannot shut our eyes to the inter-society problem posed by the increasing interest of "straight electrical" engineers in electronic techniques. As AIEE Past President Hooven wrote last August in *Electrical Engineering* "the electrical engineer who was formerly interested most in wires, cables, and switchgear is now showing an even greater interest in electronic tubes, semiconductors, and electronic components." This tendency may well explain the relative growth rates of the two societies (the percentage increase in total membership for the year ending April 30, 1956 was AIEE, 4.9 per cent; IRE, 16.4 per cent). More important, it indicates that the group of electronics-oriented engineers common to both organizations is increasing rapidly and that the problems of duplication and cross-purposes in committee effort, conferences, student activities and publications are growing proportionately. These, not size or rate of growth, are the problems to which the officers and directors of the two societies must address themselves—as allies in a common cause.

Correspondence. The August issue was noteworthy for more than the reason noted above. The Correspondence section (which follows immediately after the technical papers in each issue) ran to 19 pages that month and comprised no less than 23 communications, most of them received within the space of two months. This is roughly four times the usual number and we are somewhat mystified, but highly gratified, at the outburst. Whatever its cause, we would like to encourage more of the same.

The correspondence columns of a technical journal perform a unique function by virtue of being untrammeled by the rules governing formal technical papers. The Institute assumes no responsibility for the technical accuracy or for the opinions expressed in the communications printed, although the editors must of course reserve the right to reject items which are evidently off-base technically or trivial in content.

Since no responsibility is assumed, no formal review procedure is required and very rapid publication can be arranged, usually within two to three months of receipt. The Correspondence section is, therefore, the ideal place to announce technical discoveries and developments in the initial stages, prior to the preparation of a formal paper. Another appropriate candidate for these columns is the "small" item, which fills a chink in theory or technique, without in itself having sufficient importance to warrant more formal treatment. Still another is the commentary, which may observe a significance previously missed or draw a lesson from history. By no means barred is the argumentative essay, with rebuttals and surrebuttals, which brings into the open the conflicts, all too often hidden, which attend the development of our art.

The rules covering this section are simple. The items should be short, preferably under 1000 words, and illustrations used only when essential to the treatment. Authors are furnished proofs before publication and may order reprints. Anonymous communications will not be printed. Technical treatments are preferred, but letters on any subject affecting our profession will be printed if deemed by the editors to have sufficient breadth of interest. Finally, since the Institute receives many letters not intended for publication, correspondents should be explicit in extending permission to print.

No correspondence column amounts to anything if it is read only by the authors of the letters. So we invite the attention of all readers to this section. Much of value and interest will be found there, quickly assimilable and often indicative of important things to come.

—D.G.F.

Scanning the Issue

Quality Control in Electronics (Torrey, p. 1521)—This month's invited review paper discusses a subject which has become of vital interest to an industry which must produce quantities of highly complex apparatus that will operate reliably. It is the aim of quality control to produce products with characteristics that are suitable and, within limits, predictable. How this aim is achieved and what role statistics play in the process is unfolded in this instructive discussion. With the aid of some well-chosen examples of the use of the quality control operation and of statistical techniques in the electronics industry, the author brings to the reader the insight and experience of an organization which has been the fountainhead of new ideas in this important field.

Frequency Control in the 300–1200 MC Region (Fraser and Holmes, p. 1531)—The urgent need for making the most economical use of our crowded radio-frequency spectrum has made methods of obtaining very accurate control of frequency increasingly important in recent years. Techniques for controlling frequencies below 100 mc and above 1000 mc have been fairly well developed. In the intermediate range, however, present methods are either more unstable or more complex. The substantial improvement in the frequency stability of coaxial-cavity oscillators reported here will be of considerable interest to the many engineers now working in the uhf field.

IRE Standards on Solid-State Devices: Methods of Testing Transistors (p. 1542)—The transistor art, although young and still changing, has progressed so rapidly that it has become highly desirable to standardize, without further delay, the usages and procedures currently in force. This is the second Standard which the highly productive IRE technical committees have produced in this important field this year. As the title indicates, it deals with the methods of measurement of important characteristics of transistors and covers tests for dc characteristics, small signal applications, environmental effects and noise.

Common-Emitter Transistor Video Amplifiers (Brunn, p. 1561)—This paper presents simple and useful theories and procedures for designing various types of transistor video amplifiers. Formulas, accompanied by examples, are clearly presented for determining such characteristics as gain, bandwidth, optimum load resistor and maximum power gain.

Hazards Due to Total Body Irradiation by Radar (Schwan and Li, p. 1572)—The authors investigate the manner in which electromagnetic radiation is absorbed by the human body at frequencies ranging from 150 mc to 10,000 mc and the increases in body temperature that result. Their study shows that at above 3000 mc radiation produces heating in the skin where the sense receptors would be apt to give an exposed person adequate warning, but that at frequencies below 1000 mc the heating takes place in the deeper tissues below the sensory elements and is, therefore, potentially much more dangerous. Included in this report are estimates of the amounts of radiated energy that can be tolerated at various frequencies. The greatly increased powers that are now produced by modern electronic apparatus make these results of more than just academic interest. Moreover, this study serves to draw well-deserved attention to a general field of great future significance, namely, medical electronics.

An Analysis of Pulse-Synchronized Oscillators (Salmet, p. 1582)—In a number of important applications, most notably in single sideband and telegraph systems, it is necessary to be able to shift quickly from one working frequency to another, and to maintain the new frequency with extreme accuracy. This could be accomplished by providing a crystal to control each working frequency, but in many applications this would be impracticable because of the number of frequencies involved. This paper analyzes a substantially improved version of an earlier development in which the various harmonics of a single crystal-controlled pulse generator are utilized in a new type of phase comparison circuit to govern a variable frequency oscillator, with the result that it will either produce a continuous range of very accurately controlled frequencies or will synchronize exactly on any one of a large number of closely spaced frequencies.

A Sideband-Mixing Superheterodyne Receiver (Cohn and King, p. 1595)—A microwave receiver has been developed that combines the advantages of the high sensitivity of a superheterodyne receiver and the wide bandwidth of a crystal-video receiver. The scheme involves the use of two local oscillators, one microwave and the other vhf, to produce a large number of sidebands centered on the microwave oscillator frequency and separated from one another by the frequency of the vhf oscillator. Each sideband acts like a conventional local oscillator signal, and the received signal can mix with any one of the many sidebands, spread out over a wide range, to produce the desired IF signal. This multiple mixing technique thus presents a novel and useful method of magnifying the bandwidth of microwave receivers.

Frequency-Temperature-Angle Characteristics of *AT*-Type Resonators Made of Natural and Synthetic Quartz (Bechmann, p. 1600)—It has been found that while natural quartz from different sources is remarkably uniform, various types of synthetic quartz differ somewhat with respect to their frequency-temperature characteristics and optimum cutting angles. The author thoroughly investigates these differences, producing new data that will be of substantial interest and practical use to those working with piezoelectric materials and, in a broader sense, contributing in an area that is basic to progress in frequency control.

Distortion in Frequency-Modulation Systems Due to Small Sinusoidal Variations of Transmission Characteristics (Medhurst and Small, p. 1608)—A method of analysis is presented which sheds new light on the important problem of minimizing intermodulation distortion in fm multiplex systems. The analysis relates this distortion to various transmission characteristics of the system in such a way as to provide the systems designer with a clearer picture of the limits within which he may safely permit these characteristics to vary. These results will find important application in radio telephony and probably other broad-band microwave systems involving data transmission and telemetry uses.

Precision Electronic Switching with Feedback Amplifiers (Edwards, p. 1613)—An excellent report is presented, covering both original and prior work, on a class of electronic switches which has been developed in recent years to control the transmission of signals within various types of equipment. Unlike the on-off switches used in digital computers, in these switches the primary concern is not the speed of switching but rather the precise control of voltage or current level. This precision is achieved by utilizing a high gain feedback amplifier to minimize the differences and nonlinearities in the electronic elements that are used to switch the transmission paths. Although the principal application of this technique to date has been in analog multipliers, it should find increasing use in other fields as well, especially in signal comparison and communication switching schemes.

Quality Control in Electronics*

MARY N. TORREY†

Summary—**This paper reviews these two types of literature on quality control: 1) books, pamphlets and articles that describe what quality control is and what role statistics plays in quality control; and 2) some published examples of the use of the quality control process and of statistical techniques in the electronics industry. The quality control process is described as a dynamic operation concerned with all the coordinate steps in the specification, production, and inspection of goods to satisfy consumer wants. This is in accord with the writings of Dr. W. A. Shewhart, the originator of statistical quality control. Progress is being made in the use of quality control for improving the reliability of electronic components and equipments.**

Introduction

THE RELIABILITY problem that has plagued manufacturers of complex electronic equipments in recent years has stimulated their use of quality control methods. The early tendency was to treat the problem of reliability as a phenomenon peculiar to the electronics industry; there was a suggestion that *reliability* be controlled just as *quality* is controlled. Some recent articles have shown that quality control methods are being used for obtaining reliability.

The term reliability has been defined in many ways, but basically, according to one author, reliability includes *predictability* and *suitability*.[1] Predictability may be attained through statistical quality control, a method of controlling the quality of a product through the use of statistics. Suitability may be improved through the feedback of information in the over-all quality control operation.

Whereas quality control is often thought to comprise only the control of production processes, or a combination of process control and some inspection functions, the theory of quality control as developed by Shewhart of the Bell Telephone Laboratories encompasses a dynamic operation concerned with all steps in the specification, production, and inspection of goods having characteristics desired by the consumer. In the Bell System some of the activities referred to herein are considered to relate more to what is called *quality assurance*[2] than to *quality control*, as for example, the standard procedure for using customer complaint information to improve product design and quality which was used before the conception of the quality control operation with its formalized system of feedback. However, the term *quality control* will be used quite generally in this paper.

Since World War II there has been notable expansion in the use of quality control methods, both in this country and abroad. This paper describes quality control as an over-all operation and gives some recent examples of its use in the electronics industry. It also describes the role of statistics in quality control and gives examples of its use in particular steps of the quality control process.

Many references are cited, but they are by no means exhaustive. The literature on quality control methods and their use becomes more extensive every day. The problem of selection is a difficult one, and its solution depends, to some extent, on the background of the selector.

Quality Control as an Operation

Some Definitions

Since quality control is not a tangible object that can be photographed and described in detail, like a particular type of vacuum tube, it is necessary to begin by explaining what some of the terms mean. Shewhart defines *quality* and *control* as follows:

Quality: The quality of a thing is a set of characteristics of that thing.[3] It does not imply "high quality" necessarily; it is that which makes a thing what it is.

Control: A phenomenon will be said to be controlled when, through the use of past experience, we can predict within limits how the phenonemon may be expected to vary in the future.[3]

The phenomenon referred to is a perceptible aspect of a characteristic of the thing under consideration. No two things are exactly alike. In production, each unit is different from the ones produced immediately before and after it. The object of control is to secure the highest degree of uniformity that is economically attainable in the output of a process.

This type of control is attained through the use of statistical methods and is usually referred to as *statistical control*. The term statistical control is used in three senses: 1) as a concept of a statistical state which constitutes the limit to improving uniformity, 2) as an operation or technique of attaining uniformity, and 3) as a judgment of when uniformity has been attained.[4] Experience has shown that product characteristics are rarely in statistical control until some action has been taken to get them in control.

Product quality includes all characteristics of the product, mechanical as well as electrical. When the

* Original manuscript received by the IRE, March 1, 1956; revised manuscript received, June 14, 1956.

† Bell Telephone Labs., Inc., New York, N. Y.

[1] E. B. Ferrell, "Reliability and its relation to suitability and predictability," *Proc. Eastern Joint Computer Conf.*, pp. 113–116; December, 1953.

[2] E. G. D. Paterson, "An Over-all Quality Assurance Plan," *Ind. Qual. Control*, vol. XII, pp. 32–37; May, 1956.

[3] W. A. Shewhart, "Economic Control of Quality of Manufactured Product," D. Van Nostrand Co., Inc., New York, N. Y.; 1931.

[4] W. A. Shewhart, "Statistical Method from the Viewpoint of Quality Control," Graduate School, Dept. of Agriculture, Washington, D. C., edited by W. E. Deming; 1939.

characteristics are statistically uniform, the product is predictable. But a predictable product is not necessarily suitable for the customers' needs, as for example, a brand of components that can be depended on to fail in a standard test.

The over-all operation required to provide product quality that is predictable and suitable is referred to as *quality control*. The aim of quality control is to provide quality that is not only *dependable* (predictable) but *satisfactory* and *adequate* (suitable) for the customers' needs, as well as *economic* with respect to the use of raw materials and available production processes.[5,6] There are several related steps in the over-all operation.

Steps in the Quality Control Operation

Shewhart has given three basic steps in the quality control operation:[4]

I. The *specification* of what is wanted.
II. The *production* of things to satisfy the specification.
III. The *inspection* of the things produced to see if they satisfy the specification.

These three steps are not independent; rather, they are interrelated in a circular manner as shown in Fig. 1 and

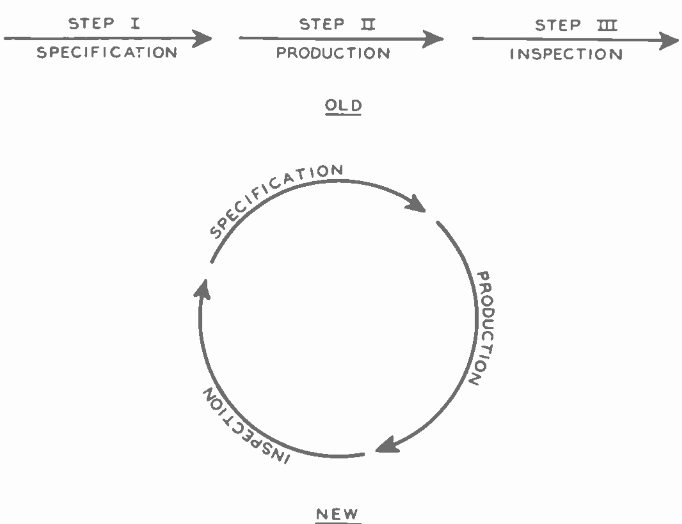


Fig. 1—Contrast of static and dynamic concepts of the relationship of the three basic steps in the quality control operation. (Reproduced from Shewart,[4] Fig. 10, permission of U. S. Dept. of Agriculture Graduate School.)

can be thought of as components of a system with feedback. They might be pictured in the form of a spiral gradually approaching a circle which would represent the idealized case where no evidence is found in step III to indicate the need for changing the specification or production process. This is a continuing and self-corrective method for making the most efficient use of raw and fabricated materials.[4]

The steps involved in the economic control of quality of manufactured product have been increased in number by broadening the three basic steps[6–9] and bringing together functions that, though they were already being carried on, had not previously been presented as part of the over-all operation. Thus Olmstead lists six steps:[8]

1) Determine the quality that is wanted through *consumer research*.
2) Perform *research and development* work to devise means for fulfilling these wants at a reasonable cost.
3) *Design and specify the product* selected and in so doing set tolerance limits.
4) *Make the product* that is specified.
5) *Inspect the product* for conformance to design and specification.
6) *Test the product in service* (*operational research*) to see that it satisfies the wants of the user in an adequate, dependable, and economic way.

Shewhart's step I has been expanded into three steps which include finding out what is wanted and how to make it. The last of the six steps listed above enlarges on the inspection function to include determination of whether or not the product satisfies the consumer.

The economic control of quality may then be considered to constitute a complete operation, including all steps necessary to make sure that consumers will get satisfactory, adequate, dependable, and economic quality.[5] It is a dynamic operation, flexible enough to take advantage of improvements in methods of manufacture or information on changes in consumer wants.

In a manufacturing plant the quality control group must be in a position to get and keep the cooperation of all three operating groups—engineering, production, and inspection.[10] Prompt interchange of information among these groups is a necessity if the quality control operation is to function properly.[11] The quality control group is a natural clearing house for such information, usually in the form of data.[12] To be most useful these data should be statistically analyzed and interpreted before being transmitted to the other groups for use as bases for action.

[5] W. A. Shewhart, "Some aspects of quality control," *Mech. Engrg.*, vol. 56, pp. 725–730; December, 1934.

[6] W. A. Shewhart, "Statistical control in the conservation and utilization of resources," *Proc. U. N. Scientific Conf. on Conservation and Utilization of Resources*, vol. 1, pp. 188–192; August 17–September 6, 1949.

[7] E. C. Harris, "Consumer Research for Quality Control," Master's thesis, Faculty of Political Science, Columbia Univ., New York, N. Y.; May, 1948.

[8] P. S. Olmstead, "How to detect the type of assignable cause," Part II, *Ind. Qual. Control*, vol. 9, pp. 22–32; January, 1953.

[9] W. E. Deming, "Statistical techniques and international trade," *J. of Market.*, vol. 17, pp. 428–433; April, 1953.

[10] E. G. Olds, "The place of SQC in an industrial organization," *Ind. Qual. Control*, vol. 9, pp. 30–34; May, 1953.

[11] H. F. Dodge, "Inspection for quality assurance," *Ind. Qual. Control*, vol. 7, pp. 6–10; July, 1950.

[12] C. E. Ellis, "How design quality control can help engineering," IRE Trans., PGEM-1, pp. 17–24; February, 1954.

Examples of the Use of Quality Control

The literature just cited tells what the quality control operation should be; the examples that follow indicate how much of the quality control operation is actually being used in a number of cases. In the second example, the name *quality control* is applied to only a small part of the over-all program, but the program is similar in many respects to the quality control operation. The importance of communication and cooperation among the design, production, and inspection groups is brought out in each example.

Example 1

This company develops and makes airborne gunfire, rocket fire, and missile weapons systems. Their ultimate consumers are members of the Air Force. In their efforts to produce reliable systems, they have developed[13] a *feedback approach to reliability* which can be related to Olmstead's six steps[8] as shown below.

1) *Determine the quality that is wanted through consumer research:* The authors do not mention how they find out what is wanted before any systems have been produced. However, the reports from the field on systems in operation give information on changes desired by the consumers.

2) *Perform research and development work to devise means for fulfilling these wants at a reasonable cost:* The authors combine research, development, design, and specification under engineering. The research and development phase is carried out, to a large extent, by the engineers in the parts application laboratory. They determine what parts, from which vendors, are likely to produce a system that will work properly under the required environmental conditions.

3) *Design and specify the product selected and in so doing set tolerance limits:* The design engineers design a system using the parts recommended and preproduction models are subjected to "rooftop" and flight tests. The system may be redesigned many times before the specification and tolerance limit stages are reached.

4) *Make the product that is specified:* The manufacturing operation comprises machining and fabricating operations and the assembly of component parts, many of which are supplied by vendors. Changes are made as need for them is indicated by information obtained in the other steps.

5) *Inspect the product for conformance to design and specification:* There are three types of inspection: a) Receiving inspection of component parts supplied by vendors. b) Inspection and tests during and at the end of assembly. c) Additional tests during installation of the system in the aircraft. The receiving inspection is a sample test to confirm the results of the vendor's inspections and is part of the over-all program for assuring that component parts conform to their specifications. After the assembly operation the product from each line for each unit is sampled and a demerit rating system is used to evaluate and control the assembly quality. Various tests are also made on the units. Thus problems are discovered at the earliest point and are more easily corrected than if inspection were made only after completion of the system. During installation the system undergoes a battery of tests to make sure that it functions properly with other systems of the aircraft.

6) *Test the product in service:* A technical liaison group works with each air frame manufacturer to observe the tests during installation and the subsequent flight tests. Field engineers are stationed wherever squadrons using the system are located to observe and report all quality problems. A maintenance depot is operated for the Air Force for repairing units which cannot be repaired in the field. Here the effect of continued service can be studied.

The fact that the "feedback approach" parallels the six steps of the quality control operation as defined by Olmstead[8] is important, but the most important aspect of the operation is the conscious, systematic plan of feedback of information. This plan consists of a closed loop of clearly defined responsibility in 1) collecting and reporting information, 2) analyzing and presenting data in such a way that the results can be readily understood, and 3) acting on the results.[13]

During the design stage, any group in the engineering organization, including the quality control group, may be called upon for information. When the system is being manufactured, the quality control group receives and analyzes all inspection data as well as information on rejection and production problems from all parts of the manufacturing operation. Corrective action requests are initiated and followed up by the corrective action unit within the quality control organization. A weekly report listing all the current problems, responsibility for action, and action being taken is circulated to the manufacturing and engineering organizations.

Weekly reports of equipment failures in the field as well as parts replacement rates are prepared by the field engineering organization. These reports are sent to the quality control organization to complete its picture of the quality of the system. They are also used by the system designers as well as the manufacturing group to steadily improve the reliability of the equipment and the ease of maintenance.[13]

Example 2

Another manufacturer of military electronic equipment has an over-all program for improving the relia-

[13] D. A. Hill and H. D. Voegtlen, "The feedback approach to reliability," *Proc. Natl. Symposium on Quality Control and Reliability in Electronics*, pp. 48–55; November, 1954.

bility of electronic systems. His program comprises five steps.[14]

1) Designing the system for reliable operation.
2) Using manufacturing and quality control techniques that are important in making the equipment reliable.
3) Packaging the equipment is such a way that it will reach the customer in a reliable condition.
4) Installing, operating, and maintaining the equipment so that optimum advantage is taken of inherent reliability.
5) Establishing a system of feedback of data from the field and taking action to improve reliability when the need is indicated by the results of analyzing such data.

A product analysis unit analyzes and coordinates the information received on various reports from the field. Daily malfunction reports from the field service representatives are used for determining which circuit components have failure rates that are significantly above expectancy.

A product analysis group[14,15] comprising heads of various sections, such as Engineering, Production, etc., meet regularly to review the report compiled by the product analysis unit on components with high failure rates. As a result of these meetings, many conditions causing or contributing to malfunctions have been eliminated by such actions as changes in design, manufacturing processes or practices, inspection methods or instructions, and vendor follow-up.[15] In this organization the Quality Control Department is concerned only with the manufacturing phase, and a product analysis group coordinates the action on field reports.

Including the product analysis group in the Quality Control Department might lead to even greater reliability improvement. This has been done by another manufacturer of complex military electronic equipment so that all failure data on complete systems can be compiled in one log and analyzed compositely.[16]

Failure data from factory final systems tests, engineering sample systems tests, air frame manufacturers, who install and test the systems, and SAC bases are analyzed to see if a trend or isolated failure is present.[16] The Quality Control Department prepares a failure report evaluating the failure and telling what corrective action has been taken or is planned. This report is sent to all interested groups who are asked to comment on the corrective action.

Example 3

A new 4000 mile broad-band transmission system is being built by the Bell System. This L3 coaxial carrier system is capable of transmitting either 1860 telephone message channels or 600 message channels and a 4.2 megacycle broadcast television channel, in each direction, on a pair of coaxials.[17] Auxiliary or line repeaters are spaced at approximately 4-mile intervals along the cable route. Equalization, power generating, and power transmission equipment are spaced at 100-to 200-mile intervals. This system requires not only a high degree of reliability of its components, but also extreme precision of certain characteristics of the components.

In order to meet the stringent system equalization and signal-to-noise objectives, all important components of the amplifier are subject to quality control procedures to assure that the average gain of groups of amplifiers will be held within narrow limits and that the gain values of individual amplifiers will form a normal distribution around the average.[18] The reason for the emphasis on the control of the amplifier components is that the quality of the amplifiers which compensate for cable and equalizer loss determines, to a large extent, the degree to which system objectives are achieved.

Besides specifying maximum and minimum engineering limits for important component characteristics, the most important characteristic, from a system standpoint, of each component is also subject to *distribution requirements*.[19] The aim of the distribution requirements is to place a continuing limitation on the pattern and the spread of measured values around their average and to impose limitations on the deviation of the average from a desired nominal value. Close cooperation between the element designer and the production engineer is essential, since the compatibility of the specification requirements and the process capability is one of the basic provisions of the general plan.[19]

Three methods are given for implementing the distribution requirements: 1) Control chart method. 2) Batch method. 3) Three-cell method. Sampling is used in the first two methods and the product is considered conforming if the sample values are controlled with respect to standards which are based on the specification limits. The third method requires 100 per cent inspection and, whereas the manufacturer may use it at any time, its use is mandatory whenever the criteria for either of the first two methods are not met. After the product has been inspected and sorted into three bins (corresponding to the three equal cells into which the

[14] G. M. Armour, "An integrated program for reliability improvement," *Proc. Natl. Symposium on Quality Control and Reliability in Electronics*, pp. 31–40; November, 1954.

[15] R. E. Landers, "Improving reliability of electronic equipment by effective analysis of field performance," 1954 IRE CONVENTION RECORD, Part 11, pp. 2–8.

[16] F. A. Davison, "The Crosley QC program for improving equipment reliability," *Electronic Applications Reliability Rev.*, RETMA, pp. 7–8; May, 1955.

[17] C. H. Elmendorf, R. D. Ehrbar, R. H. Klie, and A. J. Grossman, "The L3 coaxial system—system design," *Bell Sys. Tech. J.*, vol. 32, pp. 781–832; July, 1953.

[18] L. H. Morris, G. H. Lovell, and F. R. Dickinson, "The L3 coaxial system—amplifiers," *Bell Sys. Tech. J.*, vol. 32, pp. 879–914; July, 1953.

[19] H. F. Dodge, B. J. Kinsburg, and M. K. Kruger, "The L3 coaxial system—quality control requirements," *Bell Sys. Tech. J.*, vol. 32, pp. 943–967; July, 1953.

readings are grouped) it is packaged in groups of 5 units. As shown in Fig. 2, these packages may either contain 5 units from the center bin or 3 units from the center bin and one unit from *each* of the other bins (so that a low one is always balanced by a high one).

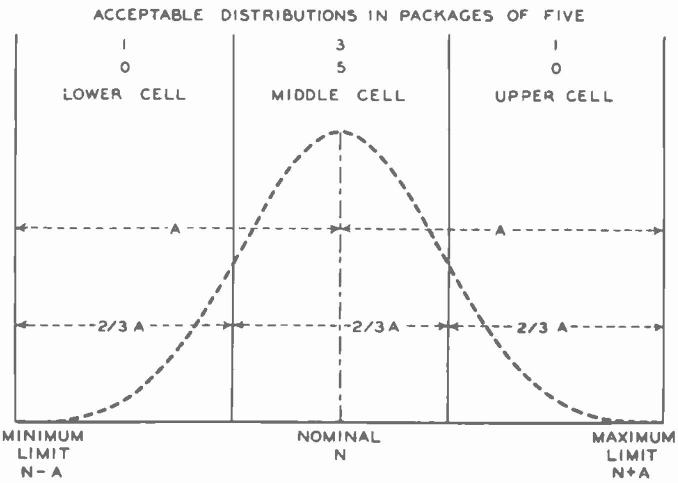


Fig. 2—Acceptable distributions of units in packages of 5, three-cell method. (Reproduced from Dodge, Kinsburg, and Kruger,[19] Fig. 6, permission *Bell Sys. Tech. J.*)

Besides subjecting the key characteristic of a component to distribution requirements, a procedure is also provided for requiring that control charts be maintained on one or more other characteristics. This gives a statistical record which indicates when remedial action should be taken.

By the end of 1953, distribution requirements were being applied to 90 components of amplifiers and transmission networks of the L3 system. Of these about one-third were being accepted by the control chart method and about two-thirds by the three-cell method.[20]

The use of distribution requirements encourages the close cooperation of the design, production, and inspection groups as well as a constant interchange of information among the groups. Extensive studies are now being made of the performance of components, particularly electron tubes, in the field. Results are being correlated with results obtained in the factory for the certification of particular tube types for system use.[21]

The three examples cited illustrate how quality control methods are being used to increase the predictability and suitability of military and communications equipment. This does not mean that its use is limited to those areas. Quality control as an over-all operation is being used more extensively in the consumer goods industries as the competition for the consumer's dollar becomes keener.[22,23]

The Role of Statistics in Quality Control

Many technical methods are needed in the over-all quality control operation; among these, statistics has an important role.[24] This section reviews books and articles on statistical methods that are applicable in quality control.

Statistics includes methods of collecting data as well as methods of analyzing, interpreting, and presenting the results in a form that assists rational decisions.[25] Statistical theory and techniques should be used in every step of the quality control operation.[4]

Statistical work is not done by statisticians alone. Ideally the statisticians work with the engineers as a team, planning how, when, and where data should be collected.[8] The supervision of data collection and analysis is in the statistician's domain but teamwork is necessary in interpreting the results.

The actual collection of data and some of the analysis are done by operators and inspectors in the factory and technicians in the laboratory or in the field.

Data Collection

Some data comprise a set of observations on all units under consideration, as in 100 per cent inspection; more often data are a set of observations on a sample of the units. (Units may be people, nails, electronic components, systems, or whatever.)

In sampling, the choice of the statistical technique to be used for analyzing the data determines how the sample units should be selected. Many techniques, such as lot sampling plans, point and interval estimates, tests of hypotheses, etc., require that sample units be selected *at random* (for example, by the use of random numbers) from the universe.[25] In continuous sampling plans, sample units are selected at random from groups of units in the order of their production.

For control charts, sample units are selected in *rational subgroups*, which are subgroups within which variations may be considered to be due to nonassignable chance causes only, but between which there may be variations due to assignable causes.[26] Other special sample designs are used in experimental work (designed experiments) and in survey sampling.[25]

[20] A. T. Chapman, "Application of quality control requirements in manufacture of components for a coaxial-carrier system," *Trans. Amer. Soc. Mech. Engrs.*, vol. 76, pp. 585–591; May, 1954.

[21] W. Van Haste and B. J. Kinsburg, "The Application of Statistical Techniques to Electron Tubes for Use in a 4000-Mile Transmission System," paper given at the meeting of the Electron Equipment Reliability Group of the AIEE; February 3, 1955.

[22] C. L. Gartner, "Quality control in television receiver manufacturing," *Ind. Qual. Control*, vol. 8, pp. 7–17; November, 1951.

[23] R. A. Posey, "Quality control in garment manufacturing," *Quality Control Convention Papers 1954*, Amer. Soc. for Quality Control, Inc., New York, N. Y., pp. 427–441; June, 1954.

[24] A. V. Feigenbaum, "Quality Control—Principles, Practices and Administration," McGraw-Hill Book Co., Inc., New York, N. Y.; 1951.

[25] W. E. Deming, "Some Theory of Sampling," John Wiley and Sons, Inc., New York, N. Y.; 1950.

[26] ASQC Standard A1-1951, "Definitions and Symbols for Control Charts," Amer. Soc. for Qual. Control, Inc., New York, N. Y.; 1953.

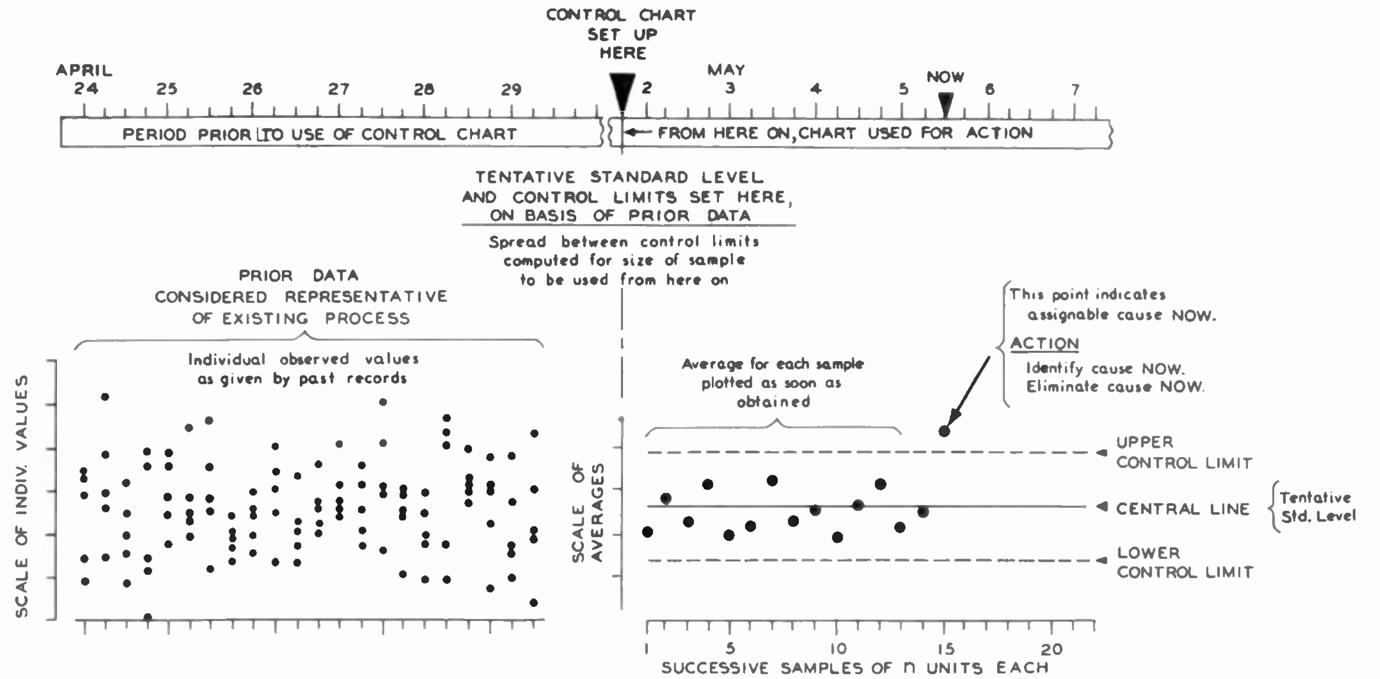


Fig. 3—Features of the control chart as used for *controlling* quality during production. This is a control chart for *averages;* each plotted point is the average of the n individual observed values for the n units in a sample. (Reproduced by permission from "American War Standard Guide for Quality Control,"[28] Fig. 2.)

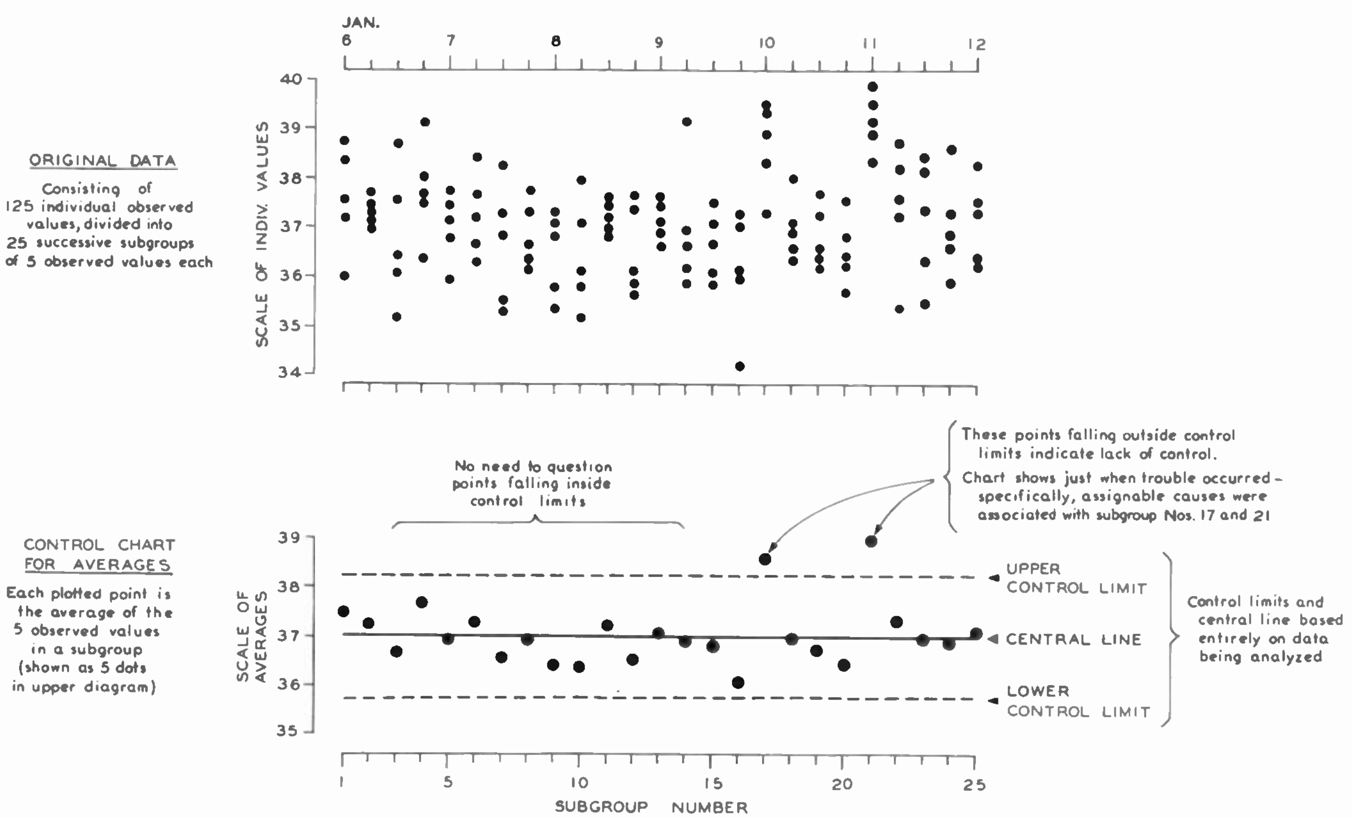


Fig. 4—Features of the control chart as used for *analyzing* a set of data to determine whether there has been lack of control. (Reproduced by permission from "American War Standard Guide for Quality Control,"[28] Fig. 1.)

After the sample units are selected the data are collected by one of two methods.[26]

Method of Variables: A characteristic is measured and a numerical magnitude recorded for each unit of the sample.

Method of Attributes: A count is made of how many units of the sample have, or do not have, some characteristic (attribute).

Whichever type of data is obtained, every effort should be made to get reliable observations. Bad data cannot be improved by any amount of analysis, but good data become more useful when they are properly analyzed, interpreted, and presented.

Statistical Techniques

Any statistical technique is appropriately a quality control statistical method if it is useful in the design and manufacture of a product the consumers want. Two statistical techniques were especially designed for quality control use. They are control charts for variables and attributes data, and statistically designed acceptance sampling plans.[27]

The control chart was invented and developed by Shewhart[3] and is an important technique for attaining statistical control. Perhaps the most important purpose of the control chart is to provide an operational procedure for controlling quality in the manufacturing plant or the laboratory.[28,29] Fig. 3, on the preceding page, illustrates this use of the control chart.

Another use of the control chart is the analysis of data for the purpose of judging whether a state of control exists or not.[28,29] Fig. 4 illustrates the use of the control chart for analyzing a set of data. In practice it is usually necessary to analyze past records before setting up the procedure for controlling future operations.

A state of statistical control is said to exist when assignable causes have been eliminated from the process (production, experimental, etc.) generating the data to the extent that practically all the points plotted on the control chart remain within the control limits.[28] Before eliminating assignable causes they must be identified. Olmstead has classified several types of trouble commonly encountered[30] and many of the statistical techniques that may be used for identifying them.[8]

When a product characteristic is in a state of statistical control, that is, when it exhibits statistical uniformity, the observed values may be considered to come from a parent statistical distribution or universe. When, in fact, there is a stable universe, statistical distribution theory may be used with confidence for predicting what values may be expected in the future.[4] Therefore, efforts toward the attainment of a state of statistical control can contribute importantly to predictability—hence to reliability.[1]

Acceptance sampling plans, based on probability theory, were also developed especially for use in quality control. A sampling plan for inspecting a lot gives the size of the first and subsequent samples, and the criteria for accepting the lot, rejecting the lot, or taking another sample.[31]

The use of sampling inspection, instead of 100 per cent inspection, by a purchaser provides the vendor with an incentive to control quality at a satisfactory level, because entire lots may be rejected and returned for correction or scrap.[11] Acceptance sampling plans are designed to provide a known degree of protection against accepting defective product. Such sampling plans may be compared by means of operating characteristic curves as shown in Fig. 5.

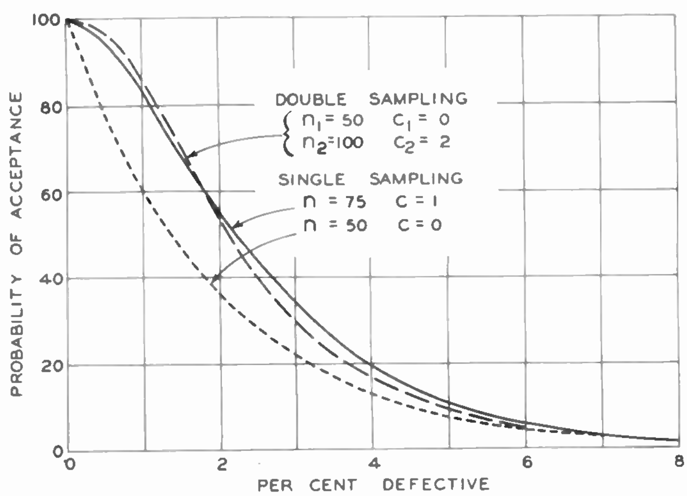


Fig. 5—Operating characteristic curves for three sampling plans. (n = sample size and c = allowable number of defectives in the sample of size n. c_2 applies to n_1+n_2.)

Several sets of sampling plans for the inspection of lots, or batches, of product by the method of attributes have been published.[27] The first set of plans to be published[32] was developed for use in the manufacturing plant. They are the only published lot inspection plans which are designed to minimize the average amount of inspection, including screening of rejected lots. Later

[27] E. L. Grant, "Statistical Quality Control," McGraw-Hill Book Co., Inc., New York, N. Y.; 1952.

[28] "American War Standard Guide for Quality Control," Amer. Standards Assn., Inc., New York, N. Y., Z1.1–1941.

[29] E. B. Ferrell, "The control chart as a tool for analyzing experimental data," Proc. IRE, vol. 39, pp. 132–137; February, 1951.

[30] P. S. Olmstead, "How to detect the type of an assignable cause," Part I, *Ind. Qual. Control*, vol. 9, pp. 32–38; November, 1952.

[31] Proposed ASQC Standard, A2, "Definitions and Symbols for Acceptance Sampling," Amer. Soc. for Qual. Control, Inc., New York, N. Y.; June, 1955.

[32] H. F. Dodge and H. G. Romig, "Sampling Inspection Tables—Single and Double Sampling," John Wiley & Sons, Inc., New York, N. Y.; 1944. (Originally published in the *Bell Sys. Tech. J.*, vol. 20, pp. 1–61; January, 1941.)

sampling tables[33–35] were designed for use by the armed services for the inspection of material submitted to them for acceptance.

Sampling plans for the inspection of lots by the method of variables have been designed to match the MIL-STD-105A attributes sampling plans.[36,37] The Armed Services specification for the inspection of reliable electron tubes, MIL-E-1B,[38,39] requires both attributes and variables inspection.

There are also acceptance sampling plans for product which is made not in discrete lots, but continuously, either on a conveyor or by some other means of continuous production.[40,41] These continuous sampling plans were originally developed for use in the manufacturing plant but they have recently been adopted for acceptance inspection by the military.[42]

There are published sampling plans available for many inspection situations, but there is room for more sampling plans which are tailor-made for particular applications.

Control charts and acceptance sampling plans have been reviewed at length because they are so important to the quality control process, especially in steps 4 and 5 given by Olmstead. In reviewing his six steps of quality control[8] it is apparent that other statistical techniques are needed. For example, in steps 1 and 6 survey sampling techniques[25] are necessary for consumer and product research. Whereas control charts are useful for analyzing data in all six steps, such techniques as frequency distributions, point or interval estimation, tests of significance, analysis of variance and regression analysis[43] are needed in steps 2 and 3, which include research, development, design, and specification of the product.

No matter which techniques have been used for collecting and analyzing the data, the results should be interpreted and presented in such a way that they do assist rational decisions and add to knowledge.[25]

Interpretation and Presentation of Data

The quantitative data, which have been analyzed, constitute only a *part* of the information used in interpretation; the judgments, or decisions, that are made depend, as well, on all available relevant information with respect to the precise conditions under which the product was manufactured, the precise conditions under which the data were obtained, etc. Such relevant information is usually qualitative and not capable of numerical expression.[44]

The presentation of data, then, should comprise two types of information.[45] 1) Essential information: functions of the observed data. 2) Relevant information: evidence that the data were obtained under controlled conditions (if possible), as well as information on the field within which the measurements are supposed to hold, and the conditions under which they were made. Graphical methods should be used as much as possible in presenting the essential information.

Examples of the Use of Statistical Methods

Examples found in published articles probably do not reflect the actual extent of the use of particular statistical techniques in the electronics industry. They do give a picture of what techniques are being applied for the first time, or in a new way.

Data Collection

Vast quantities of data are being collected on characteristics of electronic components and equipments. The amount of time and effort needed to collate and analyze these data has led to the use of punched cards and punched card equipment.

Many firms are using punched cards for records of incoming inspection of components.[14,46] Punched cards are also being used for process control inspections and tests during manufacturing.[13,14] An important use of IBM or Remington Rand cards and equipment is for handling failure data from the field, where the electronic equipments are in actual operation.[13–15,47–49] These field reports have proved invaluable for improving the reliability of equipments by determining which circuit components have high failure rates.

[33] G. R. Gause, "Quality through inspection," *Army Ordnance*, vol. 25, pp. 117–120; July–August, 1943.

[34] H. A. Freeman, M. Friedman, F. Mosteller, and W. A. Wallis, Eds., "Sampling Inspection," McGraw-Hill Book Co., Inc., New York, N. Y.; 1948.

[35] MIL-STD-105A, "Sampling Procedures and Tables for Inspection by Attributes," Supt. of Documents, Gov. Printing Office, Washington, D. C.; 1950.

[36] A. H. Bowker and H. P. Goode, "Sampling Inspection by Variables," McGraw-Hill Book Co., Inc., New York, N. Y.; 1952.

[37] ORD-M608-10 Handbook, "Sampling Inspection by Variables," Ordnance Ammunition Command, Joliet, Ill.; June, 1954.

[38] MIL-E-1B, "Military Specification, Electron Tubes," Supt. of Documents, Gov. Printing Office, Washington, D. C.; May, 1952.

[39] R. J. E. Whittier, "Inspection procedures for MIL-E-1B reliable electron tubes," IRE Trans., PGQC-3, pp. 15–27; February, 1954.

[40] H. F. Dodge, "A sampling inspection plan for continuous production," *Annals of Math. Stat.*, vol. 14, pp. 264–279; September, 1943.

[41] H. F. Dodge and M. N. Torrey, "Additional continuous sampling inspection plans," *Ind. Qual. Control*, vol. 7, pp. 7–12; March, 1951.

[42] ORD-M608-11 Handbook, "Procedures and Tables for Continuous Sampling by Attributes," Ordnance Ammunition Command, Joliet, Ill.; August, 1954.

[43] A. J. Duncan, "Quality Control and Industrial Statistics," Richard D. Irwin, Inc., Homewood, Ill.; 1953.

[44] H. F. Dodge, "Interpretation of engineering data: some observations," *Proc. ASTM*, vol. 54, pp. 603–638; 1954.

[45] ASTM Manual on Quality Control of Materials, Part 1—"Presentation of Data," Amer. Soc. Testing Mats., Philadelphia, Pa.; 1951.

[46] W. H. Bentz and R. G. Fitzgibbons, "The Bendix radio vendor quality rating system," *Proc. Natl. Symposium on Quality Control and Reliability in Electronics*, pp. 11–14; November, 1954.

[47] F. A. Hadden and L. W. Sepmeyer, "Techniques in putting failure data to work for management," *Proc. Natl. Symposium on Quality Control and Reliability in Electronics*, pp. 95–109; November, 1954.

[48] E. J. Nucci, "The navy reliability program and the designer," *Proc. Natl. Symposium on Quality Control and Reliability in Electronics*, pp. 56–70; November, 1954.

[49] H. A. Voorhees and J. E. Culbertson, "Control charts and automation applied to analysis of field failure data," *Proc. Second Natl. Symposium on Quality Control and Reliability in Electronics*, pp. 18–45; January, 1956.

The need for reliable data is even greater when the data are processed by machine than when the analysis is done "by hand." A great deal of thought has been given to the design of forms on which the data are recorded, as well as to the training of the people who actually collect the data,[15,47] so that the data recorded will be the same as those actually observed.

After the data are properly entered on the forms, the next big problem is to get the data punched on the cards correctly. Ways of minimizing mistakes are to: 1) Use mark-sense cards as the report form and a mark-sense machine to punch the cards automatically.[50] 2) Have the report form filled out in the code used for key punching the card.[50] 3) Use report forms in the punched card format that are designed so that the key punches may be made directly into the cards without obliterating the original notations.[47]

There is a tendency to collect all available failure data; it seems that sampling techniques have not yet been used for the collection of failure data as they are for the inspection of incoming and finished product.

In the collection of all types of data (failure data, inspection data, laboratory data during research and development, etc.) constant vigilance is needed for getting results that are precise and unbiased. Variables data for electonic components or equipments may be inaccurate because of test set errors. In some cases the test set errors may be of the same order of magnitude as the tolerance on the characteristic being measured.[51] Test equipment is continually being improved,[52] but a system for measuring the errors and compensating for them may be necessary in many cases.[51]

Good attributes data may often be obtained easily, but where judgment is necessary, as in the case of workmanship defects in inspection or reasons for failure in the field, different inspectors or technicians may report different data for the same trouble. The goodness of attributes data depends, to a large extent, on the design of the report form, the supervision and training of the people who take the data, and the distribution to them of a periodic report of the results.[47]

Statistical Techniques

There are many published examples of the use of statistical techniques as an aid in obtaining the desired quality of manufactured products. The examples given here are limited to those in the electronics field and are classified with respect to Olmstead's six steps of the quality control operation.[8]

Step 1: This reviewer found no published example of the use of consumer research, or the related statistical techniques of survey sampling, applied to any electronic devices. However, the importance of considering the consumer idea of quality rather than the engineer's idea of quality has been recognized.[53]

Step 2: Several examples of the use of designed experiments for research and development have been published. Analysis of variance techniques are generally used to test which design of tube is best[54] or what "treatment" combinations significantly affect the electrical properties of encased transistors.[55] However, some experimental data are analyzed more advantageously by means of control charts.[29]

Other statistical techniques which have been found useful in the development of electronic components are frequency distributions, for studying the effect of moisture treatment on the moisture seal of a certain type of capacitor, and regression analysis, for finding what caused excessive coating on the leads of a capacitor.[56]

Step 3: Although no actual example of the application of statistical methods in designing and specifying the product was found, the use of designed experiments to get information from prototypes for establishing a practical design and setting tolerance limits has been recommended.[57]

Step 4: The use of control charts for controlling production processes is widespread. A recent article describes the improvement in connector contact quality by one manufacturer through the use of control charts for variables data.[58] Another manufacturer places demerits-per-unit control charts, based on attributes data, at the end of each assembly line for each unit as a means of controlling the quality of assembly operations.[13]

Product characteristics may be controlled, but at undesirable levels. Designed experiments can be used for determining what changes in production techniques are needed to attain a desirable level. For example, information gained from a designed experiment enabled one company to increase the power output of hearing aid tubes.[59]

A series of designed experiments was used to find and eliminate the assignable causes of the uncontrolled quality of the nitrocellulose lacquer film on aluminized television tubes. The changes introduced as a result of the experiments reduced the shrinkage rate in addition

[50] J. D. Stevenson, "Electronic data processing," *Natl. Conv. Trans. 1955*, Amer. Soc. for Quality Control, Inc., New York, N. Y., pp. 141–148; May, 1955.

[51] E. J. Althaus, S. C. Morrison, and W. R. Tate, "A method of testing and evaluation of complex missile systems," 1954 IRE Convention Record, Part II, pp. 23–28.

[52] "Radio progress during 1953—quality control," Proc. IRE, vol. 42, p. 745; April, 1954.

[53] P. A. Robert, "Quality control of complex assemblies," *Quality Control Conv. Papers 1954*, Amer. Soc. for Qual. Control, Inc., New York, N. Y., pp. 155–171; June, 1954.

[54] L. Lutzker, "Statistical methods in research and development," Proc. IRE, vol. 38, pp. 1253–1257; November, 1950.

[55] M. Eder, F. Keene, and R. Warner, "Statistically designed experiment of the factorial type applied to point-contact transistors," *Proc. Natl. Symposium on Quality Control and Reliability in Electronics*, pp. 1–10; November, 1954.

[56] N. Coda, "An engineer evaluates statistical methods," *Quality Control Conv. Papers 1954*, Amer. Soc. for Qual. Control, New York, N. Y., pp. 509–511; June, 1954.

[57] H. G. Romig, "Quality control techniques for electronic components," *Ind. Qual. Control*, Vol. 10, pp. 43–47; May, 1954.

[58] J. Cannon and F. Maston, "Connector contact improvement through quality control," *Proc. Second Natl. Symposium on Quality Control and Reliability in Electronics*, pp. 8–17; January, 1956.

[59] D. Rosenberg and F. Ennerson, "Production research in the manufacture of hearing aid tubes," *Ind. Qual. Control*, vol. 8, pp. 94–97; May, 1952.

to improving the stability of the over-all process.[60]

Step 5: In the electronics field inspecting the product may include 1) incoming inspection of components purchased from vendors; 2) inspection of subassemblies during production; 3) final inspection of completed product; and 4) additional tests to see if the equipment or system is compatible with other systems after installation. Manufacturers of complex assemblies do a lot of 100 per cent inspection, but sampling inspection is often used for the first two types. Manufacturers of electronic components use sampling inspection for many of the characteristics and, of course, for any destructive tests such as life tests.

Since most of the literature refers to reliable tubes or electronic equipment for the Armed Services, the acceptance sampling plans which are mentioned are either military sampling plans (taken from MIL-STD-105A or MIL-E-1B) or plans patterned on those military plans.[46,61,62]

Life testing of electron tubes presents a special problem because it usually takes so long to get the results. Several sampling plans for life tests are in use,[63] and special statistical techniques have been devised for estimating whether the sample will pass the life test or not in a fraction of the time required for the complete life test.[64,65] Under one plan, production lots may be released early, before the life test is completed.[64]

Some manufacturers are using the incoming inspection results for rating their vendors. One rate is based on a statistical test of the significance of the difference between the sample per cent defective and the *AQL* value specified for the product.[46] The computation of the vendor rates is done automatically with punched card equipment, and these ratings serve to pick out the vendors who need corrective action.

Step 6: "Testing the product in service to see that it satisfies the wants of the user in an adequate, dependable, and economic way" is being done by many of the manufacturers of electronic equipment for the Armed Services. As mentioned above, vast quantities of failure data are being collected and collated by means of IBM and Remington Rand equipment. However, the number of published examples of the use of statistical techniques to analyze and aid in interpreting the results is small.

In one reference that has been quoted extensively[14] it is said that acceptable and unacceptable levels of failures are set for various types of components by statistical analysis. Another source[47] gives equations for limiting values of number of failures per month which may be used for judging when the number of observed failures is significantly higher (or lower) than the average number for a given type of equipment component. A recent article gives examples of the use of control charts in the analysis of failure data and describes a method of plotting the charts automatically by means of punched card equipment.[49]

These examples indicate where statistical methods are being applied in the quality control process and where their use may profitably be expanded. Throughout these examples there is interpretation of the results of the statistical analysis to show how they may be used by one or more groups for improving the predictability, or suitability, of the product. Graphical methods or visual aids are often used for presenting the results,[13,29,46,49,53,59] so that they will be readily comprehended.

Professional Society Sponsors

In this review of quality control in electronics the Proceedings of the first National Symposium on Quality Control and Reliability in Electronics has been quoted and referenced often. That Symposium, held in November, 1954, was sponsored jointly by the Professional Group on Quality Control of the Institute of Radio Engineers and the Electronic Technical Committee of the American Society for Quality Control.

The Second National Symposium on Quality Control and Reliability in Electronics, held in January, 1956, was sponsored by the same organizations, now entitled the Professional Group on Reliability and Quality Control and the Electronics Division respectively. Both the Professional Group and the Division arrange meetings on the use of quality control methods, invite people to write papers on quality control, and otherwise encourage interest in the use of statistical methods for quality control and reliability.

[60] F. Caplan, Jr., "Statistical design in electronic production-line experimentation," *Quality Control Conv. Papers 1954*, Amer. Soc. for Qual. Control, Inc., New York, N. Y., pp. 15–18; June, 1954.

[61] W. B. Hall, "Some Aspects of Quality Control in Computer Tube Applications," *Proc. Natl. Symposium on Quality Control and Reliability in Electronics*, pp. 19–22; November, 1954.

[62] R. D. Guild, "Statistical appraisal of vacuum tube reliability," *Ind. Qual. Control*, vol. 11, pp. 12–15; March, 1955.

[63] J. A. Davies, "How reliable is your life test procedure," *Quality Control Conv. Papers 1953*, Amer. Soc. for Qual. Control, Inc., New York, N. Y., pp. 255–266; May, 1953.

[64] J. A. Davies, "Life test predictions by statistical methods to expedite radio tube shipments," *Ind. Qual. Control*, vol. 4, pp. 12–17; July, 1947.

[65] W. B. Purcell, "Saving time in testing life," *Trans. AIEE*, vol. 68, part I, pp. 730–732; 1949.

Frequency Control in the 300–1200 MC Region*

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Summary—**The frequency stability of coaxial-cavity oscillators in the 300–1200-mc range can be greatly improved by the addition of a small capacitor in series with the frequency-controlling device. The series reactance thus introduced magnifies the effective capacitance external to the vacuum tube by a factor which is dependent upon the electrical length of the cavity. In theory the stabilization factor can be very high, but practical limitations due to tank-circuit losses, restrictions on reasonable values of cavity characteristic impedance, and practical minimum values of output power restrict the improvement of frequency stability over the nonstabilized oscillator to an order of ten to twenty. In the 600-mc region a preferred form of oscillator employs a tube type 6AF4 operating with an anode voltage of 50–60 volts. In the associated cavity the ratio of diameters of the outer and inner conductors is approximately 2:1 and the characteristic impedance is about 40 ohms. This oscillator, in its better range, exhibits a mean frequency stability of 0.3 cycles/mc/volt, thereby comparing favorably with overtone crystal oscillators in their upper frequency range. The oscillator produces about 80 mw of output power with a plate efficiency of slightly less than 10 per cent.**

In the frequency range of 700–1000 mc a preferred form is composed of two cavities placed end to end and employs a pencil triode, type 5876. In this oscillator the series capacitor takes the form of an iris which is interposed between the two cavities. Power output and efficiency are approximately the same as for the 600-mc 6AF4 oscillator.

The effects of temperature changes upon frequency are minimized by the utilization of materials with small coefficients of expansivity and low thermal conductivity. Commercially available Invar has attractive characteristics and an Invar-based oscillator has demonstrated the ability to maintain the frequency constant within a hundred cycles at 600 mc when the mean ambient temperature is constant. Improved Invar, such as super-Invar, may improve the temperature characteristics by as much as 3:1, and a form of ceramic, Stupalith, holds promise of even greater improvement if the problems of fabrication of the cavity can be solved.

Compensated cavities have been widely used in AFC devices, but are commonly single-frequency resonators. This paper describes and illustrates a tunable compensated coaxial cavity which covers a frequency range of ±15 mc at a center frequency of 600 mc and which exhibits a temperature sensitivity of not greater than 0.3 ppm/°C. at any point in the tuning range.

Introduction

THE INCREASING demand in recent years for greater utilization of radio facilities and communication channels has made it necessary to employ more advantageously existing frequency allocations. The consequent and necessary crowding of the rf spectrum has placed increasing emphasis upon very accurate control of frequency. Existing techniques allow satisfactory and precise direct control of frequencies below 100 mc by means of piezoelectric quartz crystals, and other techniques have achieved indirect but accurate control of microwaves by means of automatic-frequency-control devices. The techniques of control in the first-named region and in a large segment of the microwave region are well established and well-documented. A summary of data applicable to crystal-controlled oscillators is given by Buchanan[1] and extensive data on AFC circuits and microwave discriminators are due to Warner.[2]

Techniques for frequency control in the intermediate frequency range (approximately 150 to 1200 mc), by either direct or indirect means, are not as well developed. Recent investigations have sought to achieve precise frequency control by means of resonance phenomena such as molecular resonance, nuclear quadrapole resonance, or magnetic resonance. Of these, only the first is known to have been successfully applied and is thus far restricted to a few discrete frequencies above 10,000 mc. Presently used methods of frequency control in the intermediate range include indirect control by frequency multiplication (from a highly-stable low-frequency source) or direct control by means of coaxial cavities.[3] The first of these methods suffers from undesirable complexity and a (probably) poor frequency spectrum; the second possesses potentially excellent characteristics but coaxial-cavity controlled oscillators are often found to be temperature sensitive and they may in addition exhibit other instabilities.

Accurate control of frequency by means of coaxial cavities can be achieved with considerable precision. In the present paper there are described two varieties of cavity-controlled oscillators which achieve direct frequency control in the range 300–1200 mc. These oscillators exhibit improved stability in comparison to the more conventional oscillators in this range by more adequate employment of the narrow-band frequency characteristics of high-Q cavities. Inasmuch as the properties of the oscillator are to a great extent dependent upon the characteristics of the associated cavity, considerable attention has been devoted to the effect of

* Original manuscript received by the IRE, August 15, 1955; revised manuscript received June 1, 1956.

The project under which this work was conducted was supported by the Signal Corps, U. S. Army. Reproduction in whole or in part is permitted for any purpose of the U. S. Government.

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[1] J. D. Buchanan, "Handbook of Piezoelectric Crystals for Radio Equipment Designers," WADC Tech. Rep. 54–248, Wright Air Dev. Ctr., Ohio, 1953.

[2] F. L. Warner, "Review of the Methods of Stabilizing the Frequency of Klystron Oscillators by Means of Cavities," IRE Tech. Note No. 200, Telecommun. Res. Est., Gt. Malvern, Worcs., Eng. (Armed Services Document Serv. Ctr., Knott Bldg., Dayton, Ohio.)

[3] H. J. Reich, P. F. Ordung, H. L. Kraus, and J. G. Spalnik, "Microwave Theory and Techniques," McGraw-Hill Book Co., Inc., New York, N. Y.; 1947.

temperature upon sealed cavities. The paper includes a description of methods of minimizing temperature sensitivity in typical cavities.

Coaxial-Cavity Oscillator with Series Capacitor

A resonant cavity can be represented as an LCR circuit and in this form may represent the frequency-determining element of any of several basic forms of oscillators. In the simplest physical arrangement the cavity serves as a two-terminal impedance and is placed in the plate-grid circuit of a triode. The equivalent circuit of the device is then recognizable as a form of Colpitts oscillator. If a small capacitor is inserted in series with the cavity a uhf version of the Clapp[4] oscillator is produced.

Fig. 1 shows the schematic, circuit mounting, and equivalent circuit of an oscillator of this type which employs a conventional triode, type 6AF4, as the negative-resistance portion of the oscillator. The vacuum tube is mounted within the center conductor and its base projects into the space between center conductor and end-plate. Fig. 2 is a photograph of an early experimental version of the oscillator and Fig. 3 shows an external view of a 600-mc oscillator of the same type in which the cavity is constructed from Invar.

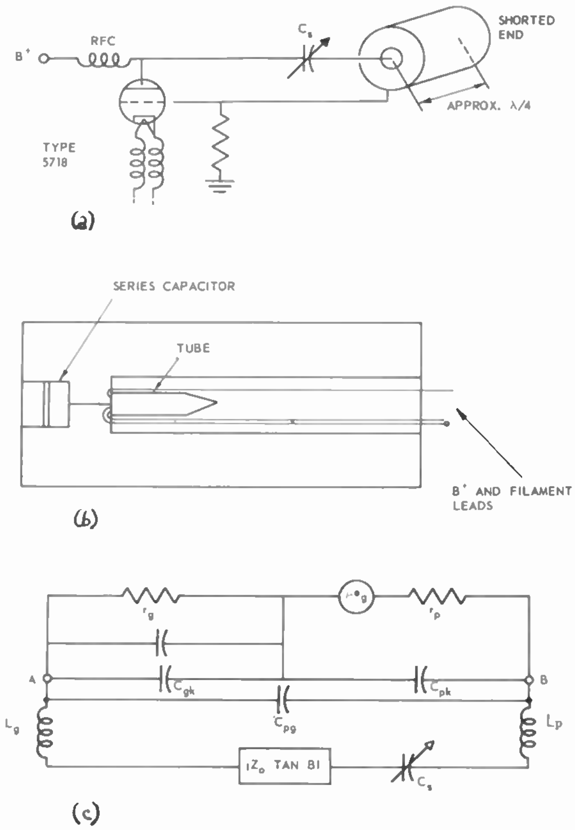


Fig. 1—Physical and electrical arrangement of series-capacitor oscillator. (a) Oscillator schematic (b) circuit mounting (c) equivalent circuit of oscillator.

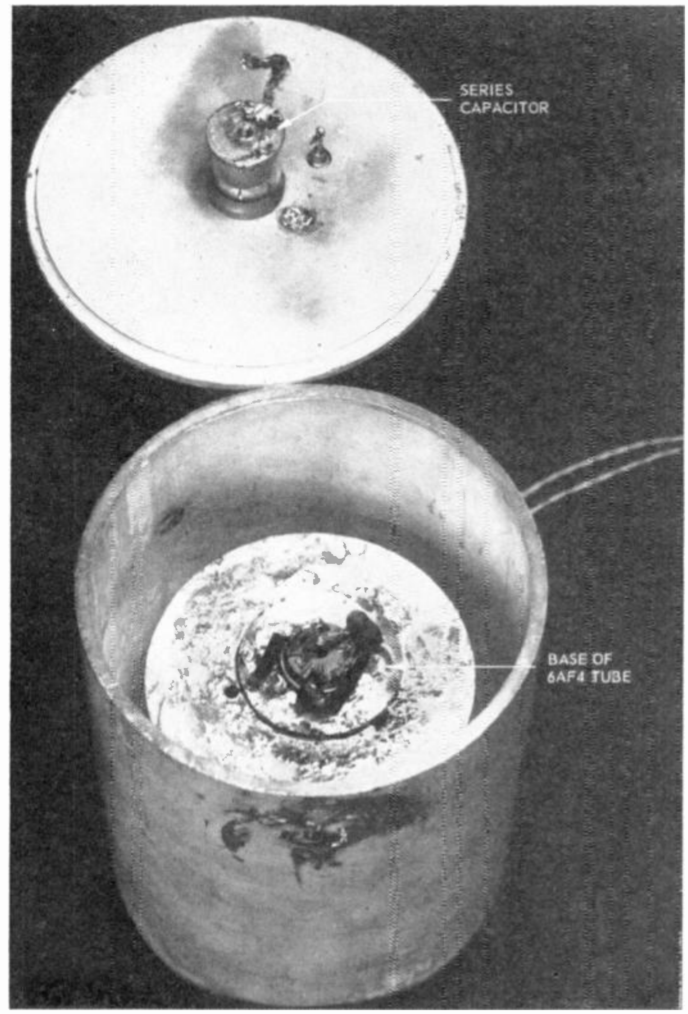


Fig. 2—Cavity-controlled oscillator, opened to show tube base and series capacitor.

Fig. 3—Assembled 600 mc oscillator.

In Fig. 1(b) the vacuum tube is mounted in the center conductor of the cavity. This arrangement has been

[4] J. K. Clapp, "An inductance-capacitance oscillator of unusual frequency stability," Proc. IRE, vol. 36, pp. 356–362; March, 1948.

used with several types of tubes including subminiature type 5718, the 6AF4, and pencil triodes. This configuration is readily adapted to hermetic sealing; the oscillators illustrated in Figs. 2 and 3 were filled with dry nitrogen and sealed under a pressure of approximately 1.1 atmospheres.

Optimum utilization of the circuit of Fig. 1 in order to provide precision frequency control demands that each of several parameters be selected with some care. The largest size cavity consistent with space requirements is usually selected in order to achieve high Q, but the cavity dimensions must be small enough that propagation of modes other than the dominant (TEM) mode is impossible. In general, no higher modes will exist if

$$\lambda > (b + a) \tag{1}$$

where b and a are the diameters of the outer and inner conductors, respectively, and λ is the free-space wavelength of the frequency of operation. In a cavity of fixed outer diameter, a diameter ratio of outer-to-inner conductors of 3.6 will result in a cavity of optimum Q.[5] It is shown later that this ratio does not necessarily promote optimum conditions in an oscillator, however it does provide a convenient guide in establishing initial parameters.

An important aspect of this form of cavity-controlled oscillator is that the oscillator will operate at a frequency which is lower than the resonant frequency of the cavity and at a point on the reactance slope which is determined by the characteristic impedance of the cavity. These statements are given greater significance by consideration of Fig. 1(c) and Fig. 4. In Fig. 1(c) the

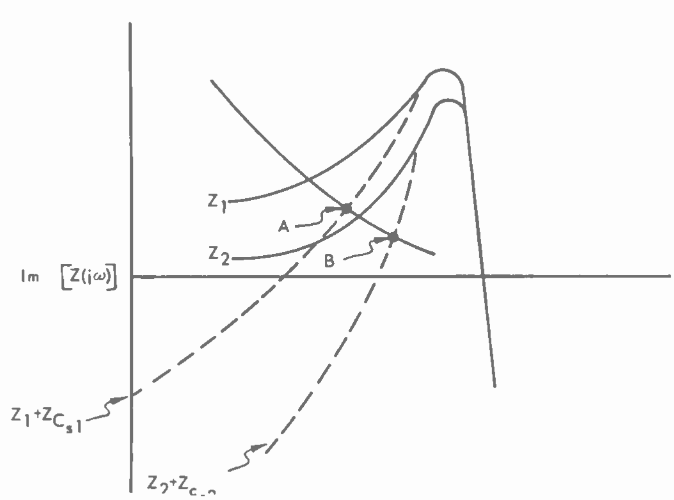


Fig. 4—Effects of varying the cavity characteristic impedance.

capacitances with double subscript represent those internal to the tube and the inductances L_p and L_g represent the effect of the leads between the tube elements and the point of contact with the external circuit. When these lead-length inductances are small, as is usually the case in vacuum tubes which are designed for operation at uhf, the net reactance presented to the external circuit is capacitive. Under these conditions oscillations can exist only if the net reactance of the external circuit is inductive. The conditions for oscillation are presented graphically in Fig. 4 which shows both the curve of the reactance due to the external circuit and also the curve representing the negative of the reactance due to the internal circuit. The latter curve, sloping downward from left to right, intersects the former to illustrate an operating point of the oscillatory circuit.

The figure shows by solid lines the reactance curves of two cavities which have the same resonant frequencies but which have different characteristic impedances. Also shown are the corresponding reactance curves when a capacitor [C_s of Fig. 1(c)] is placed in series with the cavity. Finally, the negative of the reactance seen at the terminals of the vacuum tube is shown intersecting the two dotted curves at A and B, respectively.

When the external circuit is maintained at standard conditions the possible frequency instabilities of the oscillator are usually considered to be due to a change in the reactance internal to the vacuum tube. This change is represented graphically by a vertical motion of the *internal reactance* line and a shift of the points A and B.

An optimum theoretical frequency stability should result from a cavity-reactance curve which exhibits a vertical slope at the point of intersection with the tube line. It appears that this condition could be approached by lowering the characteristic impedance of the cavity and/or increasing the series-capacitive reactance. Continuous lowering of the cavity impedance may not increase the vertical slope, however, since the Q of the cavity is decreased at the same time. There must exist an optimum characteristic impedance, not necessarily that corresponding to optimum Q, which will give optimum stability.

Efforts to determine an optimum through analytical means prove difficult because of the many parameters involved. It was found that experimental studies could be conducted rapidly under various conditions and by this means a fairly extensive compilation of data could be assembled. Some of the results obtained from experimental tests are illustrated in Figs. 5 and 6 which summarize data taken with oscillators similar in principle to that of Fig. 1 and of form similar to that shown in Fig. 3. In the referenced tests the frequency of the oscillator was changed by varying series capacitor C_s.

A measure of the frequency-stability of an oscillator is conveniently determined by changing the anode voltage in incremental steps and simultaneously noting the incremental changes in frequency. This procedure, although not providing precise results, does find common usage because it gives a convenient, if inexact, basis of comparison among oscillators of various types. Low-frequency crystal-controlled oscillators, for exam-

[5] W. A. Edson, "Vacuum-Tube Oscillators," John Wiley and Sons, Inc., New York, N. Y.; 1953.

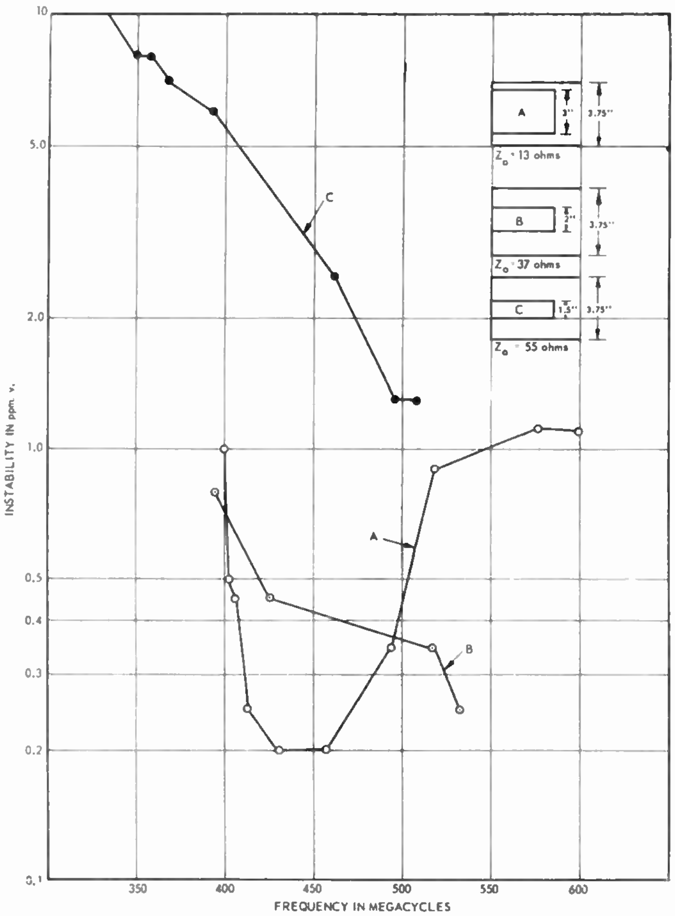


Fig. 5—Stability characteristics of oscillator controlled by cavities of different Z_0's, using tube type 6AF4.

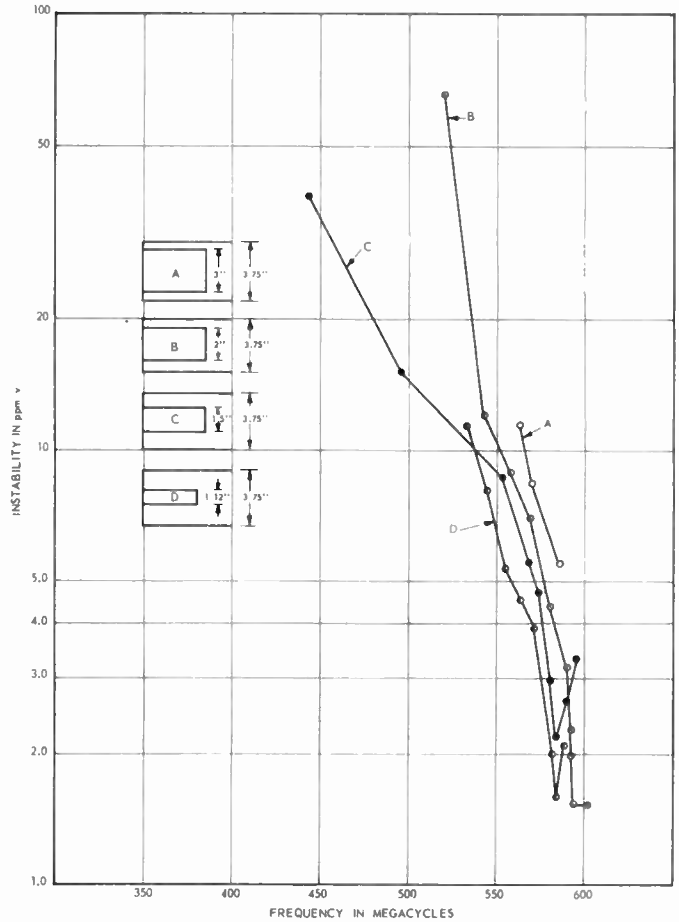


Fig. 6—Stability characteristics of oscillators controlled by cavities of various Z_0's, using tube type 5718.

ple, have been shown to exhibit negligible frequency shift as the result of small abrupt changes in anode voltage, and high-frequency (50–125 mc) overtone crystal oscillators may exhibit frequency changes in the order of 0.1 to 0.5 cycles/mc/volt, dependent upon the circuit configuration and the frequency of operation. At higher frequencies the stability tends to become progressively worse unless special techniques are employed. In the frequency range with which this paper is concerned the methods of frequency multiplication and of automatic frequency control have been employed in order to provide a desired measure of stability. However, quite precise control by direct means may be achieved in conventional oscillators at L-band frequencies, as has been shown by Stephenson[6] who describes a grid-separation type oscillator which displays frequency stabilities of 1 to 1.5 cycles/mc/volt in the frequency range under discussion.

The series-capacitor oscillator herein described is found to exhibit attractive stability characteristics when the best combinations of tube type and cavity parameters are selected, as may be observed by cross-reference between Figs. 4, 5, and 6. It will be observed from the latter two figures that in practically all cases the best stability occurred at the highest frequency, that is, when the series capacitor was at its minimum setting and the external reactance is represented by that dotted line in Fig. 4 which has maximum slope. Best stabilities were encountered when the tube type 6AF4 was employed and in cavities in which the ratio of b/a was less than 3.6. The series capacitor utilized with the 6AF4 oscillator was of the tubular hermetically-sealed variety and has a claimed zero-temperature coefficient over the normal range of ambient temperatures. It provided capacitive values of from 1–10 micromicrofarads, a range of values which permitted the tuning ranges indicated in the figures. The amplitude of oscillation varies with the magnitude of the series capacitance, being greater for large values of capacitance. For extreme minimum values of capacitance oscillations may completely cease, hence it was necessary to provide a calibrated *stop* beyond which the capacitor could not be set. For the 6AF4 the working range of values of C_s was in most cases about 1.5–6.0 mmf and the variation in amplitude was approximately 1–2. At minimum amplitude the oscillator produced about 2 volts rms into a 50-ohm coaxial cable, or about 80 milliwatts of rf power. The variation of amplitude in this oscillator was not considered to be a serious disadvantage, since the primary function is to provide a stable frequency. It

[6] J. G. Stephenson, "Designing stable triode microwave oscillators," *Electronics*, vol. 28, pp. 185–187; March, 1955.

may normally be assumed that buffer amplifiers will be employed when voltages of considerable amplitude are required. Conventional methods of amplitude control may reasonably be utilized in such buffers.

An interesting possibility of extending the range of tuning while maintaining the best possible stability exists in addition tuning *plungers* to the configuration illustrated. Although the difficulties of fabrication and control of such plungers within a hermetically-sealed cavity precluded such experimentation, it is speculated that a greater over-all optimum range of stability might be effected by the combination.

Stabilizing Effect of the Series Capacitor

The general conclusions relative to frequency stability illustrated in Fig. 4 are found to be quite well substantiated by experimental results but the number of parameters involved in a specific oscillator are large and no precise design data can be conveniently established. However, a further insight into the stabilizing effects of the series capacitor may be gained from relationships established by Helber[7] to whom the following analysis is due.

The oscillator frequency for a purely conductive load, employing a lumped circuit of inductance L and capacitance C_e will be

$$f = \frac{1}{2\pi\sqrt{LC_e}} \tag{2}$$

which may be written, employing the relationship $v = f\lambda$, as

$$\lambda = 2\pi v\sqrt{LC_e}. \tag{3}$$

Differentiating, and simplifying, one obtains

$$\frac{d\lambda}{dC_e} = \frac{1}{2C_e} \tag{4}$$

from which it is evident that to minimize the change in frequency as a function of variations in capacitance it is necessary to make C_e as large as possible.

If the external load of the oscillator is a low-loss transmission line, it is evident that the requirement for oscillation is

$$\frac{1}{\omega C} = Z_0 \tan\left(\frac{2\pi L}{\lambda}\right) \tag{5}$$

which may be written

$$\frac{\lambda}{2\pi v C} = Z_0 \tan\left(\frac{2\pi L}{\lambda}\right). \tag{6}$$

In these equations, λ = the wavelength in centimeters, L = the length of the line in centimeters, and v is the velocity of propagation along the line $\doteq 3\times10^{10}$ cm in a coaxial cable with air dielectric.

[7] C. A. Helber, "Improving stability of uhf oscillators," *Electronics*, vol. 20, pp. 103–105; May, 1947.

If (6) is differentiated, there is obtained

$$\frac{d\lambda}{dC} = \frac{\lambda}{C[1 + \theta(\tan\theta + \cot\theta)]} \tag{7}$$

where $\theta = 2\pi L/\lambda$ = the electrical length of the line in radians.

When (4) and (7) are equated, one obtains

$$C_e = \frac{C}{2}[1 + \theta(\tan\theta + \cot\theta)] \tag{8}$$

which relates the total effect of the capacitively-loaded line to the equivalent capacitance of the lumped equivalent circuit.

An intuitive method of rationalizing the effect of the series capacitance is to reason that the added series reactance forms an isolating barrier between the frequency-controlling device (in this case the transmission line or coaxial cavity) and the tube. The action of the series reactance is to reduce the effect upon the frequency-controlling device of changes within the tube. The intuitive reasoning can be reduced to an analytical basis by demonstrating that the total capacitance of the equivalent lumped circuit is increased by the addition of the series capacitance. The analysis is completed by employing the concept of energy storage in the circuit capacitances to show that

$$C_e = \frac{C_t(C_s + C_t)}{2C_s}[1 + \theta(\tan\theta + \cot\theta)] \tag{9}$$

where C_t is the original capacitance in shunt with the transmission line and the other quantities are as previously defined.

As an example, the capacitance C_t of the oscillator of Fig. 1 may normally be expected to be in the order of 3 to 5 mmf. If C_s is 2 mmf and if the electrical length of the line is 80° then it is a matter of simple computation to show that the equivalent capacitance is increased from 10 to 20 times over that which would exist in the absence of the series capacitance. The frequency stability, according to (4), is improved by one-half of this ratio.

A certain compromise between stability and efficiency is indicated. It is evident that as the series reactance is increased more circulating current, with greater tank losses, must exist in the transmission line if oscillations are to be sustained. In the oscillator which utilized the tube type 6AF4, operating in the region of better stability, plate voltages of 50–60 volts were employed and plate currents of 15–20 milliamperes were measured. If the minimum rf power of 80 milliwatts is assumed, the plate efficiency is indicated to be not greater than 10 per cent. This order of plate efficiency seems to be a necessary compromise in order to achieve the desired stability.

The analytical data presented in this section can be correlated, although in a somewhat tedious manner, with the data presented in Fig. 4. The reactance curve of a cavity can be plotted from data tabulated by measurements with an rf bridge, then the electrical length of the line corresponding to any prescribed reactance can be determined from the curve. Such a measurement procedure was followed in a few cases during the course of the experimentation and the curves of the reactance of the frequency-controlling device were plotted. However, no satisfactory measurements of the shunt capacitance of the tube (when operating with normal plate voltage and plate current) were obtained, hence the points of intersection of internal and external reactance were not considered to be sufficiently accurate to warrant correlation with calculated values. The indicated intersections did, however, correspond well in general sense to the conclusions drawn from Fig. 4.

Reduction of Temperature Sensitivity

The effects upon frequency of change of ambient temperatures have not been considered in the preceding discussion. The adverse effects of varying ambient temperatures upon a precision frequency-control device are normally minimized by enclosing the frequency-controlling element in a temperature-stabilized oven. Quartz crystals are usually enclosed in small ovens which maintain the temperature within a fraction of a degree at a prescribed level. Cavities are much larger than crystals and the ovens required to enclose them are more difficult to maintain at a fixed temperature. However, a prescribed mean temperature may be maintained without difficulty and without utilization of complex heat-controlling elements. The effects upon frequency of the relatively large variations, about a prescribed mean temperature, can be minimized by the utilization of materials in the cavity which either have little temperature sensitivity or which exhibit properties of heat transfer which minimize changes of temperature at the frequency-controlling point. In coaxial cavities it is important that an external temperature change does not quickly reach the inner conductor since the resonant frequency of the cavity is closely controlled by the physical length of its center conductor.

Various methods of minimizing the temperature sensitivity of a cavity are described herein in later paragraphs, but a series of experiments have demonstrated that the utilization of the nickel-steel, Invar, as the base material in cavity construction may adequately satisfy the prescribed requirements in cavity-controlled oscillators. Invar displays the combined properties of small expansivity (less than 1 part per million per degree centigrade) and low thermal conductivity (about one-tenth of that of brass). The effect of these properties is illustrated in Fig. 7 which is a record of the results of an abrupt change in ambient temperature upon the Invar-based, series-capacitor oscillator shown in Fig. 3. In the experiment illustrated in Fig. 7 the oscillator was suddenly subjected to a temperature change of 25°C. and was thereafter maintained in the new environment. The curve of Fig. 7 shows that the long-term effect is to reduce the operating frequency by something less than 1 ppm/°C., *i.e.*, less than 25×610 cps. An interesting and useful effect is found in the positive excursion of the frequency which occurs during the first few minutes after the new temperature is applied. This action is apparently due to a relatively rapid expansion of the outer conductor and precedes expansion of the center conductor because of the low thermal conductivity of Invar. The effect of the expansion of the outer conductor is to decrease the capacitive end-effects of the plate which encloses the end of the cavity remote from the center conductor. Since reduction of the end-effect capacity tends to raise the operating frequency and expansion of the center conductor tends to lower the frequency it is seen that a certain amount of self-compensation is present which will tend to reduce the over-all frequency change resulting from any prescribed change in temperature.

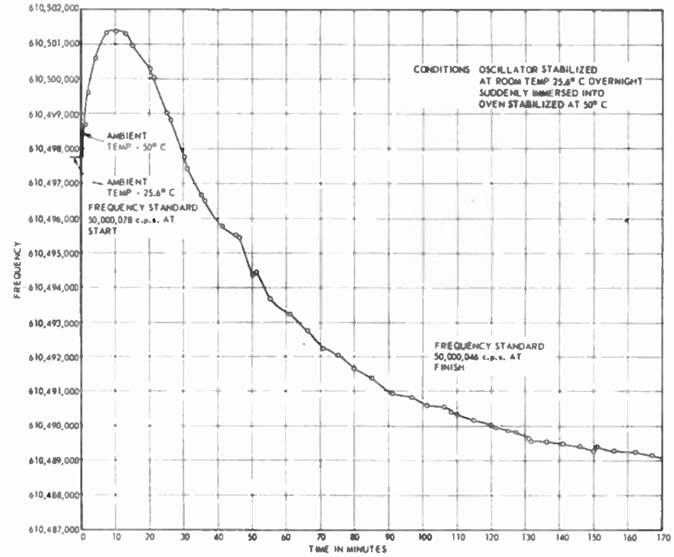


Fig. 7—Effect of abrupt change of ambient temperature on invar cavity.

The tendency for the oscillator to stabilize in frequency is illustrated by the asymptotic approach of the curve to a final value at which point a variation of not more than 100 cycles at a mean frequency of 610.4893 mc was observed. It is evident that when the oscillator is enclosed in an oven which has accurate temperature stabilization about a mean that a highly stable frequency may be maintained and that short-term temperature variations about the mean, even if of considerable magnitude, will not be reflected as significant frequency changes because of the low thermal conductivity of the cavity.

The characteristics of slow heat transfer in Invar-based cavities lead to certain disadvantages when the vacuum tube is enclosed within the center conductor. The plate dissipation of the vacuum tube, unless deliberately restricted by means of lowered anode voltage, produces an accumulation of heat which tends to reduce the life of the tube. One tube type 6AF4, when working at the anode potential of 50–60 volts quoted in the preceding example, was employed in semicontinuous operation for a period of over six months and in continuous 24-hour duty for an additional 30 days. However, when the anode voltage was raised to 75 volts there was found to be an appreciable reduction in tube life, an effect that was even more pronounced as the plate dissipation was permitted to approach the recommended maximum for that type of tube. The undesirable reduction in tube life could be partially eliminated by the introduction of an element of high thermal conductivity within the center conductor whose purpose would be to transfer excessive heat to a point exterior to the cavity. Unfortunately, such a device works in two directions because changes in external temperature are reflected back to the vacuum tube and thus tend to counteract the stabilizing effect of the Invar. The results of the experiments indicate that operation at reduced anode voltage is an acceptable, although not ideal, means of maintaining a satisfactory compromise between power, stability, and tube life.

Other Factors Influencing Operating Characteristics of the Oscillator

Other items which must be considered in discussing the over-all characteristics of an oscillator include the tendency toward frequency drift during warmup, the effect of changes of filament voltage upon frequency, the effect of mechanical vibrations, and properties of conducting surfaces on the frequency-controlling element. The Invar-based oscillator is ideally suited to continuous-duty operation, but much less so to operation in which the filament voltage is turned off and on at frequency intervals. Fig. 7, which shows the effect of sudden changes in ambient temperature gives an approximate illustration of the action when the oscillator is turned on from a cold start. A more exact illustration is given in the following tabulation. A 610-mc Invar-based oscillator which was used as a test vehicle performed as follows:

Frequency at time zero....................	610.475 mc
Frequency at time plus 5 minutes........... (A drift of −30 kc)	610.445 mc
Frequency at time plus 30 minutes.......... (A further drift of −8 kc)	610.437 mc
Frequency at time plus 60 minutes.......... (A further drift of −3 kc)	610.434 mc
Final stabilized frequency..................	610.433 mc

The relatively large drift which occurs during the first few minutes does not appear if the filament voltage is applied continuously. For this reason all other tests on this oscillator were premised on a continuous filament-voltage basis, with the result that the application of anode voltage produced a much smaller frequency drift during stabilization.

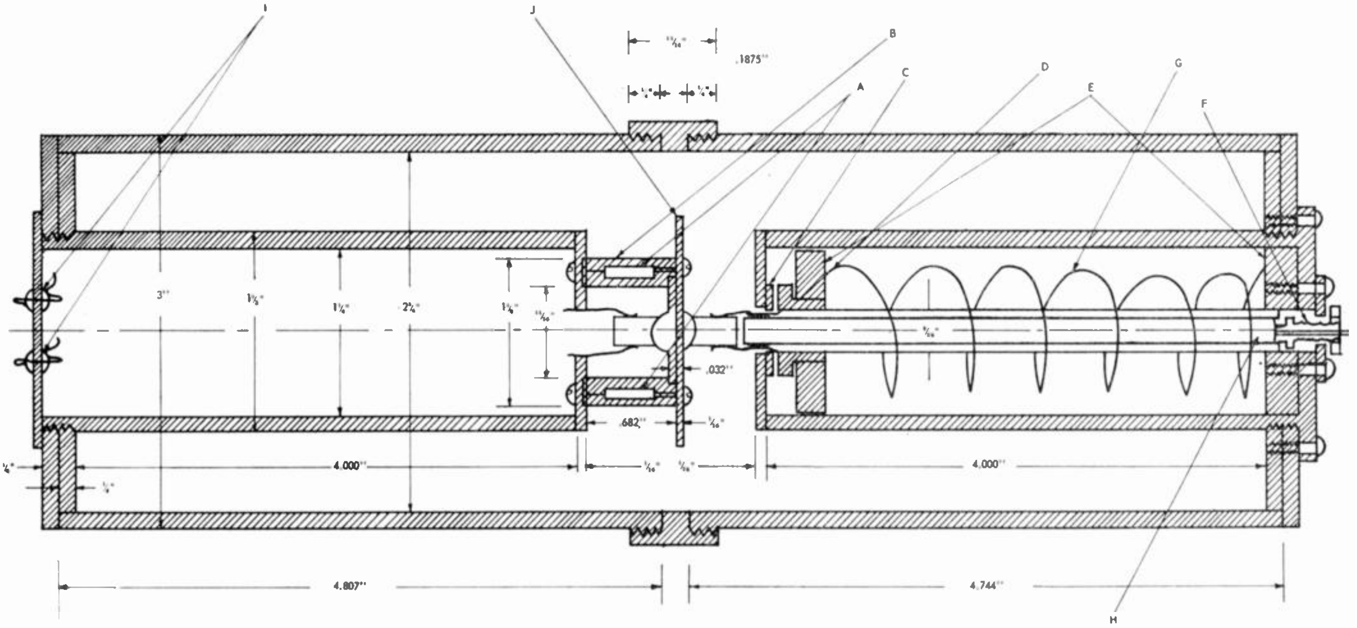
The effects of changes in filament voltage were studied by means of tests on two oscillators of identical configuration but of different materials. One oscillator employed an Invar-based cavity, the other a cavity constructed of brass. Brass has an expansivity of about 20 ppm/°C. and a thermal conductivity which is about ten times that of Invar. Both oscillators, operating in the 600 mc range, were stabilized in frequency through precise control of the ambient temperature (in a thermostatically-controlled heat chamber) and of anode and filament voltages. The filament voltage of each was then raised 0.2 volt rms from 6.3 to 6.5 volts rms. The brass-based oscillator immediately began a negative frequency drift of over 2 parts per million in each 10 minute period of time, a rate of drift which remained substantially constant during the full 60-minute period of the test. The Invar-based oscillator, on the other hand, drifted slightly more than 2 parts per million during the first 20 minutes and exhibited essentially no drift thereafter. In summary, the effect of the filament voltage change in the Invar-based oscillator was to produce an over-all frequency shift of about 1400 cycles from an original nominal frequency of 600 mc, but a very much larger change in the oscillator with brass cavity.

The effects of mechanical vibration upon the oscillator were observed, but not recorded. In an early test the oscillator was placed upon a suspended floor in a wooden cabinet; to this floor a blower motor was attached in order to provide a considerable mechanical vibration. The oscillator output, heterodyned against a stable frequency standard, was reduced to a low frequency and displayed upon an oscilloscope. Observations during the presence and absence of vibrations indicated that a frequency modulation of several kilocycles, centered at a mean frequency of 600 megacycles, was produced. In a subsequent test the oscillator was bedded in a ½ inch foam rubber matting and the observations were repeated. It was found that the frequency modulation had been almost completely eliminated. The results indicate that shock-mounting techniques commonly employed in operating equipment should eliminate objectionable frequency modulation due to mechanical vibrations.

The conductivity of the surfaces internal to the cavity is an important factor in any cavity-controlled oscillator and is of particular importance in the series-capacitor version in which the circulating tank current is of relatively large magnitude. If the material of which the cavity is constructed is quartz or a ceramic, then it must be coated with a conducting material. When Invar is employed, the natural conductivity of the basic material is not adequate to provide the desired Q in the resonant circuit and the material should be plated with a noble metal. Silver quite adequately fulfills the requirements for all of the base materials mentioned; methods of assuring satisfactory plating are described in a later paragraph and in the appendix.

Re-entrant Type Oscillator Employing Pencil Triodes

The series-capacitor oscillator with tube type 6AF4 has been found to be limited to frequencies below 700



A. Grid resistors mounted in dielectric
B. Dielectric (scoth plasticast) centered on inner conductor
C. Type NPO dielectric washer (CAP=75mmfd)
D. Brass sleeve soldered to back of C and $\frac{5}{8}''$ tubing
E. Bakelite insulating supports for $\frac{5}{8}''$ tubing
F. Feed-through capacitor (CAP-2000 $\mu\mu$fd)
G. Spring to provide pressure on C
H. Bushing soldered inside $\frac{5}{16}''$ tubing to hold F
I. Filament feed through capacitors
J. Iris

Fig. 8—Cross section of assembled reentrant cavity oscillator.

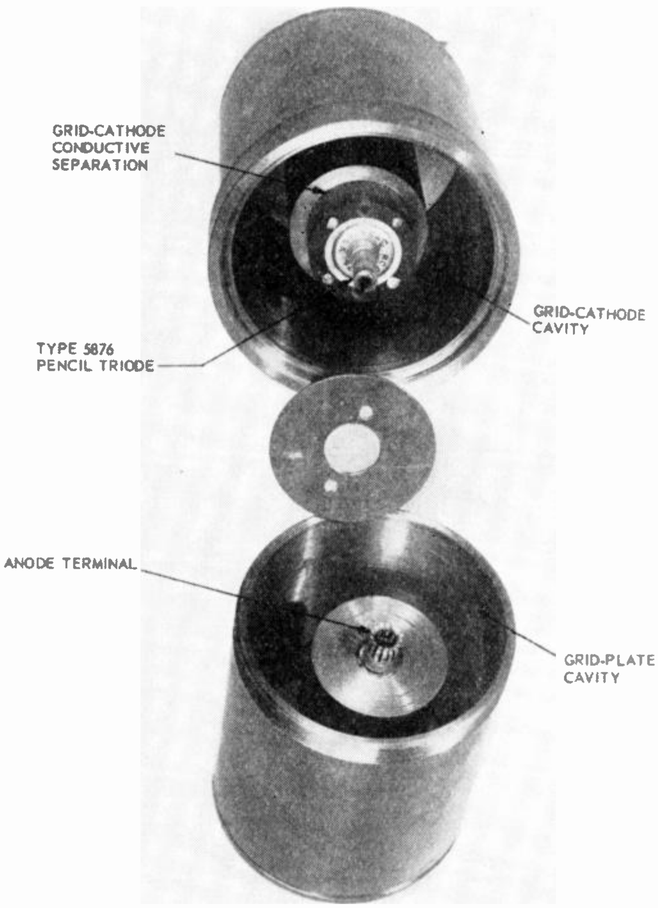


Fig. 9—Semiassembled view of reentrant cavity oscillator.

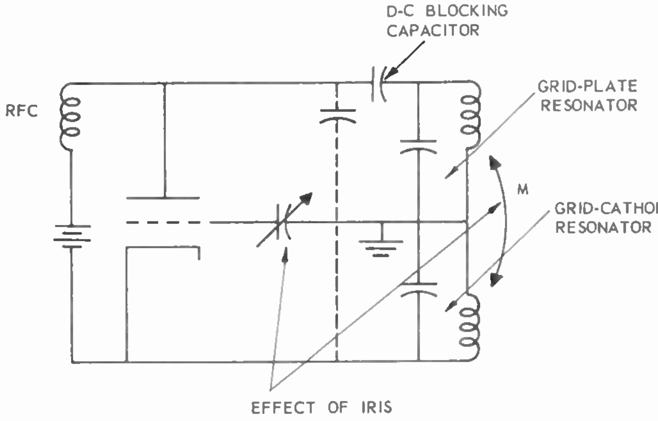


Fig. 10—Lumped equivalent of capacitively-coupled reentrant cavity.

mc. However, the same basic method of improving frequency stability through the addition of series-capacitive reactance can be utilized at higher frequencies. One configuration illustrating this method employs two cavities placed end to end and employs a pencil triode, type 5876. A cross section of an assembled oscillator of this type is shown in Fig. 8 and a semiassembled view is shown in Fig. 9.

This oscillator has a combined feedback path which involves electrostatic coupling between cavities and also energy transfer through a series-capacitor. A lumped equivalent circuit is shown in Fig. 10. The series capacitor, whose electrical action is essentially the same as that in the oscillator previously described, takes the form of a thin, circular metal plate which makes electrical contact to the grid of the pencil triode and provides capacitive coupling to the high-impedance point in the outer conductor of the grid-plate cavity. This circular plate is designated as the *iris* because it partially closes the opening between the two cavities. The iris is marked as item J in Fig. 8 and may be seen in Fig. 9 where it has been removed from the oscillator and lies between the cavities.

Fig. 10 shows that the iris has control over two quantities, the direct feedback through coupling between cavities (M) and the feedback through the capacitive coupling. It might be anticipated from the discussions about the series-capacitor oscillator that the oscillator presently described should demonstrate best stability when the series capacity is a minimum, that is, when the iris is the smallest consistent with oscillations. Experimental tests have proved that this is a correct assumption; most stable operation occurred in all cases when the iris employed was the smallest possible

The construction of the oscillator requires some novel features in design which are best illustrated by a description of the individual elements of Fig. 8. The grid-cathode cavity occupies the left portion of the assembly, the grid-plate cavity occupies the right half. The pencil triode, type 5876, is secured by its grid ring, in a molded and machined Scotch Plasticast tubing B, by pressure from the iris J. The plasticast also serves to hold four 4.7K grid resistors which are symmetrically placed about the axis. The tube makes electrical contact to the cathode cavity by means of the spring fingers shown. Plate supply voltage is furnished by means of bushing H and another set of spring fingers. The rf electrical path in the plate circuit is completed from the bushing H through a ceramic disc capacitor C of 75 mmf capacity which is maintained firmly in contact with the spring finger base by pressure exerted from spring G.

The pencil triode is often a very delicate tube type and is easily broken. For this reason, accurate alignment of supports is required. The dimensions on the sketch are shown to three decimal places allowing a ± 0.005-inch tolerance in normal shop practice.

Three models of this oscillator were constructed, all single-frequency devices. (Tuning may be accomplished by the insertion of tuning screws through the central peripheral collar or by the introduction of plungers.) The first oscillator was designed to operate in the 600- to 650-mc region, the others in the 800- and 1000-mc regions, respectively. The operating characteristics of all three were found to be essentially identical and the stability characteristics compare favorably with those of the series-capacitor oscillator, as do the power output and efficiency. Although higher frequency versions have not been constructed it is believed that this configuration may be successfully applied at frequencies as high as 1500 mc.

Minimization of Temperature Sensitivity in Coaxial Cavities

The usual coaxial cavity is constructed with a center conductor of approximately one-quarter wavelength at the desired operating frequency, while the outer conductor is made somewhat longer in order to provide an *overhang* beyond the inner conductor. This *overhang* region is a circular waveguide operating beyond cutoff if the proper design parameters are used. If the *overhang* must be made short because of material or space considerations there is a capacitance effect to the end-plate which appears in shunt with the high-impedance end of the cavity. This capacity is given as

$$C = \frac{a}{30\pi v}\left(\frac{\pi a}{4d} + \ln\frac{(b-a)}{d}\right) \text{ farads} \tag{10}$$

where v is the velocity of propagation in the cavity dielectric, b and a are the diameters of the conductors, and d is the length of overhang. It is this capacity which produced the positive excursion of Fig. 7. If the overhang is made sufficiently long the effect of C may be made negligible.

Two methods of construction are available by which to provide cavities of minimum temperature sensitivity. The most obvious is to employ a basic material, such as Invar, which has low expansivity and low thermal conductivity. Commercial Invar has an expansivity of slightly less than 1 ppm/°C., but recent reports from other activities indicate that an almost complete removal of impurities from the material can result in expansivities as low as 0.3 ppm/°C. It appears that such super-Invar, if generally available, can provide a most satisfactory basic material for construction of very stable cavity-controlled oscillators.

Two other materials which have attractive characteristics are fused quartz and certain types of ceramics. Fused quartz is reported to have an expansivity of about 0.5 ppm/°C. and has been used in the construction of cavities employed at microwave frequencies but is not conveniently employed in coaxial cavities because of difficulties in fabricating and combining the required elements. One form of ceramic, appearing under the trade name of Stupalith,[8] is potentially attractive because it is claimed to have a zero temperature-coefficient of expansion. Early tests with this material were unsatisfactory because of insufficient information relative to methods of establishing adherent silver surfaces. Techniques which led to successful plating and soldering of this completely temperature-insensitive material were developed by the authors[9] in the course of recent investigations. Inasmuch as numer-

[8] Bulletin No. 1051, Stupakoff Ceramic and Manufacturing Co., Latrobe, Pa.

[9] D. W. Fraser and E. G. Holmes, "Precision Frequency Control Techniques (500 MC and Higher)." Final Rep. Project No. 229-198. Georgia Inst. of Tech., Atlanta, Ga., Signal Corps Contract DA-36-039-sc-42590.

ous problems arose in determining optimum methods of plating, and also because considerable interest in the problem has been shown by numerous investigators, a summary of the methods of plating has been included in the appendix. Subsequent tests have indicated that cavities constructed from Stupalith may be made insensitive to temperature only if the complete cavity assembly (inner and outer conductor and end plate at the low impedance end) is homogeneous in its entire structure. No oscillators have as yet been constructed with cavities made of Stupalith.

The final method to be mentioned as a means of minimizing temperature sensitivity is the process of temperature compensation. Compensated coaxial cavities utilize a principle by which the unequal expansivities of two materials are employed to neutralize the normal effects of thermal expansion upon the inner conductor. The length of the compensating section is made equal to the product of the length of the inner conductor and the ratio of expansivities of the materials used in the inner and outer conductors. Although the principle is not new, a description of a tunable compensated cavity is included in this paper as an illustration of the principle in a model which has performed satisfactorily in practice.

Fig. 11 illustrates a tunable compensated cavity with

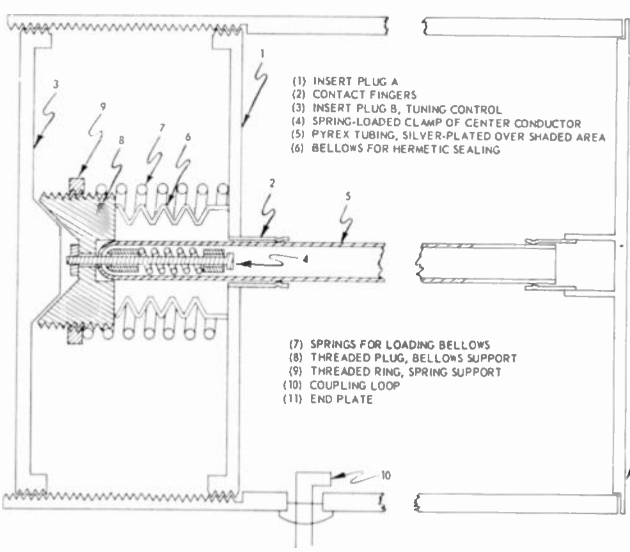


Fig. 11—Tunable compensated cavity.

brass outer conductor which is designed for the 600-mc range. The active cavity extends from the end-plate at the low impedance end of the cavity (Insert Plug *A*) to the end of the shaded portion of the Pyrex center conductor. The spring-loaded section at the left permits controlled axial motion of the center conductor within a hermetically-sealed bellows. The length of the compensating section (Plug *B* to Plug *A*) is very closely 3/20 of the active length of the center conductor, a ratio determined by the coefficient of expansion of Pyrex (3 ppm/°C.) and of brass (20 ppm/°C. This cavity is tunable from approximately 585 to 615 mc and tests conducted within this range have shown that the temperature sensitivity is very small (less than 0.3 ppm/°C.).

Compensated cavities do not readily lend themselves to utilization in oscillators in which the vacuum tube is interior to the cavity. They are, however, well suited to use in an arrangement which employs a discriminator and some means of controlling the oscillator frequency from the output of the discriminator. Numerous descriptions of various automatic frequency control devices appear in the literature and will not be discussed here.

Appendix

Methods of Plating and Soldering Materials in Cavity Resonators

Plating of Glass- or Ceramic-Based Materials

Silvered surfaces were utilized exclusively in all the resonators constructed and tested. The plating methods used on Pyrex, Stupalith, and Vycor are essentially alike. The steps involved and the precautions necessary are summarized as follows:

1) Remove any uneven or badly discolored spots on rod or tubing by a nonconductive abrasive, such imperfections will not retain a silvered surface.

2) Clean material thoroughly of grease and dirt by scrubbing with detergent. Remove detergent thoroughly by copious rinsing.

3) Bake ceramic materials at 350°F., or above, for two hours to remove all traces of absorbed moisture.

4) After the material has cooled to approximately room temperature, spray with silver-based air-drying paint such as DuPont No. 4760 Silver Paste which has been thinned by Toluol to proper consistency for application by spraying.

5) After two hours air-drying, place in oven and raise to 1250°–1350°F. for $\frac{1}{2}$ hour; cool in oven to 450°F., then remove and air-cool to ambient temperature.

6) Inspect surface with a 5 to 10 power glass. If necessary, remove hills and pits with fine abrasive. A mirror-like finish is required to insure a final satisfactory conductor.

7) Porous ceramic surfaces which are not to be plated, but will come into contact with the solutions should be masked off with a good grade of lacquer which will not be affected by HCl.

8) Electroclean 1 minute at about 8 volts. A wire soldered to an outside surface will permit this process as well as the plating to follow. Rinse with tap water.

9) *Pickle* in a 30 per cent HCl solution. Rinse again with tap water.

10) *Strike* plate at a high current (about 15 to 25 amps per square foot) for 30 seconds. The surfaces should be completely coated with Ag after strike. Rinse with tap water. The Ag *strike* solution should be composed of:

8 to 10 oz. NaCN per gallon of solution
0.1 to 0.2 oz. AgCn per gallon of solution

11) Silver plate in an agitated cyanide solution at 5 to 10 amps per square foot until the desired thickness is obtained. The solution should be held at about 30°C. Upon completion of plating, rinse with tap water again.

Plating of Invar

No special treatment is necessary other than that normally performed on ferrous materials. The steps in plating used are:

1) Remove imperfections on surface with an abrasive, clean with a strong detergent, and rinse well with tap water.

2) Electroclean 1 minute at about 8 volts. A wire soldered to an outside surface will permit this process as well as the plating to follow. Rinse with tap water.

3) Any portions (such as screw threads) that should not be plated are masked off with lacquer.

4) Follow steps 9) through 11) in previous section.

Soldering Techniques

Particular care must be exercised in soldering to silver-plated materials whose base is of glass or ceramic. Most common solders such as tin-lead, indium, silver-enriched tin-lead, Cerrobend, etc., have the undesirable property of contracting significantly during solidification. This contraction may easily strip the silvered surface from the base. The use of solder of minimum contraction reduces or eliminates this difficulty. One example of a satisfactory solder is found by the name of Cerotru (58 per cent Bi and 42 per cent Sn), which has a melting point of 281°F. This solder readily receives electroplated silver, hence soldered joints may be given a final silver surface.

This solder may be applied by a small iron which concentrates heat at a point or by flowing it onto a surface which has been raised by oven-heating to a temperature equal or greater than the melting point of solder. The preliminary application of a flux such as Nalco-14, manufactured by the National Lead Company, results in improved adhesion of the solder.

Correction

Arthur Uhlir, Jr., author of the correspondence entitled "High-Frequency Shot Noise in *P-N* Junctions," which appeared on pages 557–558 of the April, 1956 issue of PROCEEDINGS, has informed the editors of the following correction to his letter.

On the right side of (4) on page 558, the plus sign (+) should be a minus sign (−).

IRE Standards on Solid-State Devices: Methods of Testing Transistors, 1956*

(56 IRE 28. S2)

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* Approved by IRE Standards Committee, July, 1956. Reprints of this Standard, 56 IRE 28.S2, may be purchased while available from the Institute of Radio Engineers, 1 East 79th Street, New York, N. Y. at $0.80 per copy. A 20 per cent discount will be allowed for 100 or more copies mailed to one address.

1.0 General

1.1 Scope

This standard deals with the methods of measurement of important characteristics of transistors. In general, these characteristics are referred to as parameters of the devices.

Because of the youthfulness of the transistor art, methods of testing transistors will continue to change considerably before the art can be considered to have "stabilized" sufficiently for complete standardization. This standard corresponds to the current state of transistor testing methods, and its publication by the IRE is considered preferable to waiting for a future stabilization of the many rapid changes now characteristic of this field.

1.2 General Precautions

Attention is called to the necessity, especially in tests of apparatus of low power, of eliminating, or correcting for, errors due to the presence of the measuring instruments in the test circuit. This applies particularly to the currents taken by voltmeters.

Attention is also called to the desirability of keeping the test conditions, such as collector voltage and collector current, within the safe limits specified by the manufacturers. If the specified safe limits are exceeded, the characteristics of the transistors may be permanently altered and subsequent tests vitiated. When particular tests are required to extend somewhat beyond a specified safe limit, such portions of the test should be made as rapidly as possible and preferably after the conclusion of the tests within the specified safe limit.

1.2.1 Repeatability: Care must be taken that the measured parameter values are repeatable within precision of measurement after performance of any one or all tests performed on the device.

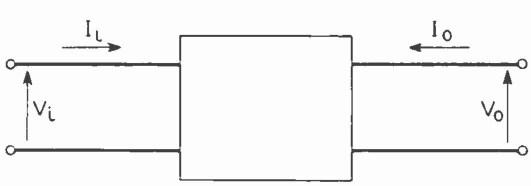


Fig. 1—General four-terminal network.

1.3 Four-Terminal Representation

Fig. 1 shows a 4-terminal network sometimes known as a 2-terminal pair. The behavior of this network may be defined in terms of the quantities V_i, V_o, I_i, and I_o. The ac input current and voltage are I_i and V_i and the output current and voltage are I_o and V_o. Similarly, the dc characteristics may be represented in terms of the input and output voltage, V_I and V_O, and the input and output current, I_I and I_O.

Since the transistor may be employed in three circuit configurations usually referred to as common base, common emitter, and common collector, there is a possible ambiguity in the definition of any one parameter unless the circuit configuration is stated definitely. Six possible sets of parameters exist for defining the 4-terminal network and the choice of the appropriate set therefore depends on the nature of the device to be characterized.

The h parameters are used throughout the standard since they are peculiarly adaptable to the physical characteristics of transistors. In previous literature these have been referred to as the series-parallel parameters, but a recent paper[1] coined the name "hybrid" which has become a significant method of identification and will, in the interests of clarity, be used throughout this standard.

A specific method of notation involving the letter subscript is used throughout this standard, but no preference over number subscripts is implied thereby. See IRE Standards on Letter Symbols for Semiconductor Devices, 1956 (56 IRE 28. S1).[2]

The three most commonly used sets of parametric equations are:

Open-circuit impedance parameters

$$V_i = z_i I_i + z_r I_o \tag{1}$$

$$V_o = z_f I_i + z_o I_o. \tag{2}$$

Short-circuit admittance parameters

$$I_i = y_i V_i + y_r V_o \tag{3}$$

$$I_o = y_f V_i + y_o V_o. \tag{4}$$

Hybrid parameter

$$V_i = h_i I_i + h_r V_o \tag{5}$$

$$I_o = h_f I_i + h_o V_o. \tag{6}$$

The input impedance h_i is the impedance between the input terminals when the output terminals are ac short-circuited

$$h_i = V_i/I_i \quad \text{when} \quad V_o = 0.$$

The voltage feedback ratio h_r is the ratio of the voltage appearing at the input terminals, when they are ac open-circuited, to the voltage applied to the output terminals

$$h_r = V_i/V_o \quad \text{when} \quad I_i = 0.$$

The forward current multiplication factor h_f is the ratio of the current flowing into the output terminals, when they are ac short-circuited, to the current flowing into the input terminals

$$h_f = I_o/I_i \quad \text{when} \quad V_o = 0.$$

The output admittance h_o is the admittance between the output terminals when the input terminals are ac open-circuited

$$h_o = I_o/V_o \quad \text{when} \quad I_i = 0.$$

[1] D. A. Alsberg, "Transistor metrology," 1953 IRE Convention Record, Part 9, pp. 39–44. Also IRE Trans., vol. ED-1, pp. 12–15; August, 1954.

[2] Proc. IRE, vol. 44, pp. 934–937; July, 1956.

LIST OF TERMS

V_i = ac input voltage.
V_o = ac output voltage.
I_i = ac input current.
I_o = ac output current.
V_I = dc input voltage.
V_O = dc output voltage.
I_I = dc input current.
I_O = dc output current.
z_i = input impedance, small signal, output open-circuited.
z_o = output impedance, small signal, input open-circuited.
z_f = forward transfer impedance, small signal, output open-circuited.
z_r = reverse transfer impedance, small signal, input open-circuited.
y_i = input admittance, small signal, output short-circuited.
y_o = output admittance, small signal, input short-circuited.
y_f = forward transfer admittance, small signal, output short-circuited.
y_r = reverse transfer admittance, small signal, input short-circuited.
h_i = input impedance, small signal, output short-circuited.
h_I = input resistance, static value, output short-circuited.
h_o = output admittance, small signal, input open-circuited.
h_O = output conductance, static value, input open-circuited.
h_f = forward current transfer ratio, small signal, output short-circuited ($=-\alpha_f$).
h_F = forward current transfer ratio, static value, output short-circuited ($=-\alpha_F$).
h_r = reverse voltage transfer ratio, small signal, input open-circuited.
h_R = reverse voltage transfer ratio, static value, input open-circuited.
z_{in} = input impedance, small signal, output termination Z_o.
z_{out} = output admittance, small signal, input termination Z_i.
V_e = ac emitter voltage.
I_e = ac emitter current.
V_c = ac collector voltage.
I_c = ac collector current.
r_e = ac emitter resistance derived from T-equivalent circuit.
r_b = ac base resistance derived from T-equivalent circuit.
r_c = ac collector resistance derived from T-equivalent circuit.
r_m = ac transfer resistance derived from T-equivalent circuit.
C_o = collector capacitance measured at collector electrode.

Note: See also IRE Standards on Letter Symbols for Semiconductor Devices, 1956 (56 IRE 28. S1).[2]

2.0 METHODS OF TEST FOR DC CHARACTERISTICS

The static characteristics of a transistor represent its performance only at zero or low frequency. The static characteristics are the input, output, and transfer. In general, these characteristics may be obtained up to the point where thermal effects become significant, or where critical voltages or currents are exceeded. High frequency or pulse methods such as those described in section 2.1.3 may be used to obtain information beyond this point.

2.1 DC Point-by-Point Method

The point-by-point method of obtaining characteristics requires the introduction of a direct voltage or current at one pair of terminals, and the measurement of the current or voltage at either the same or a different pair of terminals, depending upon the characteristics under examination. A family of characteristics can be obtained by measuring a voltage-current characteristic while another voltage or current is changed stepwise over the range of interest in accordance with usual practice. A representative arrangement for the determination of the common base characteristics of transistors is shown in Fig. 2.

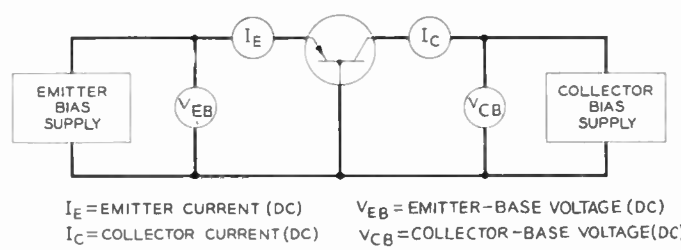


Fig. 2—General dc measurement arrangement.

2.1.1 General Precautions in Transistor Measurement: Test conditions which cause large voltage or current surges, or exceed the safe limit of dc power dissipation should be avoided. Large overloads even for a small fraction of a second may cause damage to a transistor or modify its characteristics.

The correct voltage polarity must be observed at all times. Incorrect voltage polarity may seriously damage the transistor and test equipment.

Transistors are inherently temperature sensitive devices. The effect of the ambient temperature must be taken into account, and possibly also the internal temperature rise due to the dissipation in the device occurring during the test.

2.1.2 Visual Displays: Visual displays of transistor static characteristics are useful for prediction of performance in circuits up to frequencies at which reactance effects become important. Oscilloscopic displays are useful in disclosing small irregularities in the voltage-current characteristics which may escape observation by the point-by-point method. The visual display is particularly useful for determination of trends or orders of magnitude in transistor parameters.

The transistor static characteristics normally displayed visually are: input voltage vs input current; input voltage vs output current; output voltage vs input current; output voltage vs output current; and output current vs input current.

2.1.2.1 General Precautions in Oscilloscopic Display: As a preliminary step a passive network may be used as a dummy transistor to check the over-all circuit performance before actual application to transistors, and the voltage-current characteristic may be compared with known or published curves.

Cumulative heating effects must be anticipated. If extreme care to prevent overloading is not taken a gradual shift in the observed characteristic is noted.

Instability may result if suitable series resistance is not provided, particularly in the case of point-contact transistors which are in general short-circuit unstable.

In addition, the general procedures noted in section 3.5 must be observed.

2.1.3 Pulse Methods: It is often of considerable importance to know the static characteristics of transistors beyond the normal operating range where thermal effects would be significant if point-by-point methods were used. In such cases it is necessary to employ pulse methods in which the transistor is allowed to pass currents only for short intervals of such duration and recurrence frequency that the average power dissipation is small.

Pulse methods may be employed for obtaining input, output, and transfer characteristics. The basic circuit elements required for a pulse method are pulse generators and suitable current and voltage indicators. Where one pulse generator is employed, it is usually connected to the appropriate terminals depending upon the characteristic desired, with provision for introducing bias. If more than one pulse generator is used, it is necessary to synchronize the pulses. In general, one pulse generator is adequate and simpler to employ.

A variable amplitude pulse voltage or current is applied to one pair of terminals and simultaneously the corresponding pulse amplitude of current or voltage is measured at the same or a different pair of terminals.

2.1.3.1 Precautions: Care must be taken that the original static characteristics are reproducible after the device has been pulsed.

2.2 Load (Dynamic) Characteristics

The methods used for the determination of load characteristics from static characteristic curves, and the direct measurement of load characteristics have been published.[3] Load characteristics permit calculation of the performance data for the transistor such as input power, output power, efficiency, dissipations, etc.

2.2.1 Direct Measurement of Load Characteristics: The load characteristics of a transistor can be measured directly, without resorting to calculation from the static characteristics. When reactive effects are significant, they will have considerable effect upon the load characteristic. It is therefore advisable to measure load characteristics at the frequency at which the transistor is to be used.

2.3 Maximum Electrode Voltage

When the voltage-current characteristic of a transistor is presented by any appropriate technique, marked changes in slope and/or discontinuities may be noted as a function of electrode voltage and circuit configuration. These may be due to either junction breakdown, thermal gradients, or internal instabilities. In general, a junction breakdown may be correlated with the resistivity of the material in the base layer, while that due to thermal gradients is generally much lower, and is characteristically poorly defined.

2.3.1 General: A maximum electrode voltage is measured by the potential which results in a specified change in the parameter being measured. It may also represent a potential above which destructive irreversible changes occur in the transistor. In either event it represents a locus of electrode bias voltages and currents which define maximum usable operating conditions. The maximum electrode voltage will be a function of the common electrode utilized when the characteristics are taken.

When specifying the peak voltage, even though nondestructive, the duration of the peak and the duty cycle must be specified because of the short thermal time constants of the semiconductor element.

2.3.2 Definition: A maximum electrode voltage may be defined on any of the following bases:

1) Junction voltage breakdown.
2) Maximum power dissipation capability of the transistor.
3) Nonlinearity of the electrode voltage-current characteristic.

An example of these limitations is shown for a typical collector voltage-current characteristic in Fig. 3 on the following page. In this figure it may be seen that the definitions just given will govern in different regions of the characteristic. The maximum electrode voltage for the characteristic shown will be given by V_{C2}, V_{C3}, and V_{C4}.

V_{C2} is defined as the voltage corresponding to the point of tangency of the voltage saturation tangent with the I_C-V_C curve, with input current specified as shown in Fig. 3.

V_{C3} defines the voltage at which the rated power dissipation is attained.

V_{C4} defines the voltage at which the nonlinearity of the characteristic becomes a substantial limitation to use.

[3] "IRE Standards on Electron Tubes, Methods of Testing," Proc. IRE, Part I, vol. 38, pp. 917–948, August, 1950 (see sec. 4.2, p. 925); Part II, vol. 38, pp. 1079–1093, September, 1950.

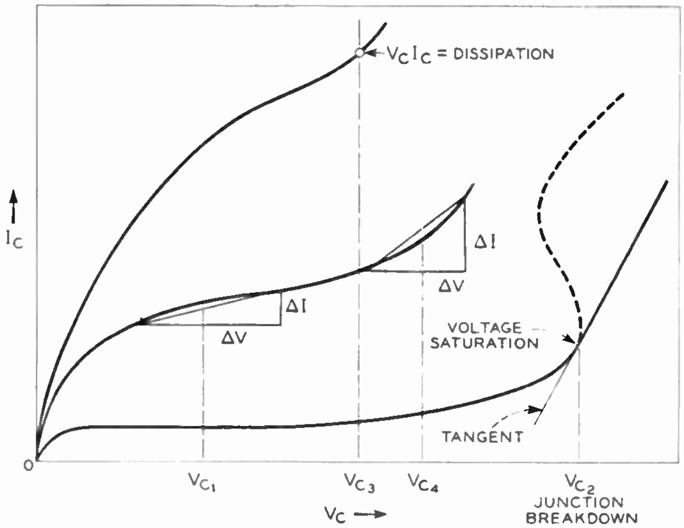


Fig. 3—Maximum electrode voltages.

2.3.3 Precautions: The following specific precautions should be followed in addition to those noted in section 2.1.1. a) High-speed oscilloscopic sweep methods may be preferable to point-by-point and other low-speed methods, because inaccuracies due to thermal gradients and incipient junction breakdown are minimized. b) The electrical characteristics must be reproducible within the margin of error after a determination of maximum collector voltage.

3.0 Methods of Test for Small Signal Applications

For purposes of this section small signal operation assumes linearity over the operating range. For linear operation the transistor is completely specified by means of four independent parameters which are in general complex quantities whose value may depend upon frequency, operating point, and environment. For linear operation the value of the parameter must be independent of the amplitude of the signal.

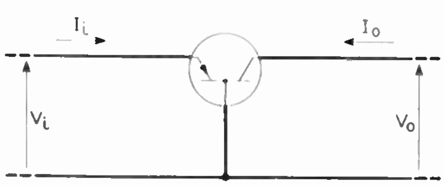


Fig. 4—Transistor four-pole representation.

Most transistors can be characterized by a 4-pole representation in which two terminals are usually common as in Fig. 4 above. It is often found that the measurement of transistors having a common base current amplification of less than unity is more practicable with either the short-circuit admittance or hybrid (y or h) parameters while with those having a common-base current amplification greater than unity the open-circuit impedance or hybrid (z or h) parameters are more practicable.

In the illustrations which follow, the common-base configuration is generally shown for purpose of economy. It will be understood that the parameters may be taken in any possible stable configuration.

3.1 General Precautions

It is necessary that the test signals employed be small enough so that the transistor operation is linear. Generally, the greater the parameter accuracy desired, the smaller the test signal must be. A method of determining whether the signal is sufficiently small is to decrease the amplitude of the test signal progressively until a further decrease in amplitude produces no change within the accuracy desired in the value of the parameter.

Methods of determining test signal amplitude include the checking of voltage or current amplification derived from combinations of 4-pole parameters with those measured experimentally. If the test signal is sufficiently small, the derived and measured values will check within the accuracy desired.

In the methods of measurement to be discussed, it is preferable to either ac short-circuit or ac open-circuit different terminal pairs to carry out the measurement. In order to be certain of the accuracy of the measured data, it is necessary to ascertain the adequacy of the ac short or open circuits employed. Stray series elements such as lead inductance may seriously alter ac short circuits. One method of ascertaining the adequacy of the ac short or open circuit employed is to change progressively by known amounts the terminal admittance or impedance while making measurements of the parameter under investigation. A graphical plot, as shown typically in Fig. 5, of the measured parameter as a function of absolute magnitude of the terminal admittance or impedance would show an asymptotic approach to the correct value.

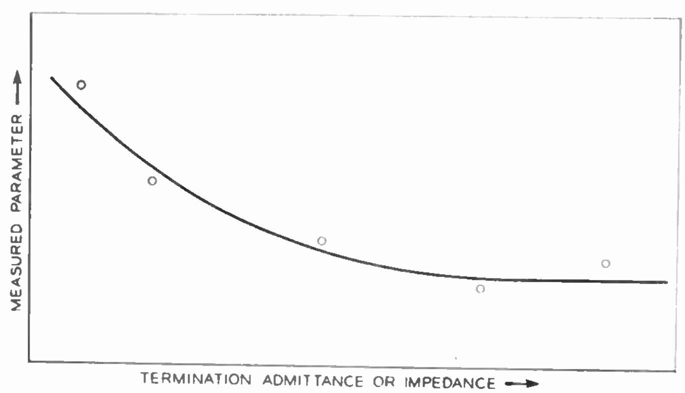


Fig. 5—Adequacy of termination.

For preliminary measurements or for approximate results, the basic idea of this check method may be applied by making certain that there is negligible change in the parameter being measured when the ac short or open circuit is changed by an appreciable amount.

Care must be exercised, particularly as a result of the inherent short-circuit or open-circuit instability of the transistor, to insure that the measurement circuit is not oscillating at either the test or some spurious frequency. The presence of oscillations will be indicated generally by abrupt changes in the curve of Fig. 5.

Another method of checking the adequacy of the ac short or open circuit is to choose reasonable and convenient values of terminating impedance. The complete measurements are then made. The results of these measurements are then used to calculate whether the initially chosen ac short or open circuits were adequate.

The results of measurements will be dependent upon circuit and environmental conditions. Until such time as these conditions become well standardized by usage, it will be necessary that the exact conditions of the measurement be specified.

1) Circuit configuration employed and quantities measured.
2) DC terminal voltages and currents (any two independent quantities are sufficient).
3) Test frequency employed.
4) Temperature.

The following test conditions may be important and may have to be specified: 1) Humidity, 2) Aging period, 3) Test socket employed and shielding configuration.

Independent measurements of ratios of parameters are useful for determining the adequacy of the ac short or open circuit and for insuring the absence of oscillations. Generally, if the independent measurement of ratios of parameters checks the computed values within a few per cent, there is reasonable assurance of linearity, adequacy of termination, and freedom from oscillations. Some of the parameter ratios that may be independently measured are listed below.

1) The forward current transfer ratio, which is the negative ratio of the alternating current at the ac short-circuited output terminal to the alternating current introduced at the input terminal

$$\alpha_f = z_f/z_o = -y_f/y_i = -h_f.$$

2) The reverse current transfer ratio, which is the negative ratio of the alternating current at the ac short-circuited input terminal to the alternating current introduced at the output terminal

$$\alpha_r = z_r/z_i = -y_r/y_o = \frac{h_r}{h_ih_o - h_fh_r}.$$

3) The forward voltage transfer ratio, which is the ratio of the alternating voltage at the ac open-circuited output terminal to the alternating voltage introduced at the input terminal

$$\mu_f = z_f/z_i = -y_f/y_o = \frac{-h_f}{h_ih_o - h_rh_f}.$$

4) The reverse voltage transfer ratio, which is the ratio of the alternating voltage at the ac open-circuited input terminals to the alternating voltage introduced at the output terminal

$$\mu_r = z_r/z_o = -y_r/y_i = h_r$$

3.2 *Open-Circuit Terminal Measurements*

Some of the transistor parameters may be defined under conditions of open-circuit termination. The transistor dc biases are applied to produce the specified operating point and the appropriate terminals are ac open-circuited and the specified measurements made. The ac open circuit is conveniently supplied by a suitable series impedance, a parallel-resonant circuit, a transmission line, or other means.[4] The circuit used to produce the open circuit must have an adequately large impedance at the frequency or frequencies of measurement. Methods of ascertaining the adequacy of the open circuit are discussed in section 3.0.

3.3 *Short-Circuit Terminal Measurements*

Other transistor parameters are defined under conditions of short-circuit termination. The transistor dc biases are applied to produce the specified operating point and the appropriate terminals are ac short-circuited and the specified measurements made. The ac short circuit is conveniently supplied by a large admittance such as a capacitor, a series-resonant circuit, a transmission line, etc. The circuit used to supply the short circuit must have an adequately large admittance at the frequency or frequencies of measurement to insure reliability. The adequacy of the short circuit may be determined by the methods discussed in section 3.0.

3.4 *Finite Termination Measurements*

Where tests for the adequacy of short or open circuit show that it is not adequate, and cannot be readily attained, then a finite termination must be used. This is often the case where measurements must be made over a large range of frequencies, where the variations of a characteristic as a function of frequency must be determined, or where circuit noise considerations impose a limitation on experimental accuracy. After the dc operating biases are applied to produce the specified operating point, the specified terminals are ac terminated by the finite impedance termination. The finite impedance is conveniently supplied by a nonreactive fixed resistor, a monocyclic[5] (frequency-independent) network, a terminated transmission line, or by an impedance of known characteristics.

3.5 *Methods of Parameter Measurement*

3.5.1 General: The characteristics of a transistor may be measured at the specified terminals under the stated

[4] "IRE Standards on Electron Tubes, Methods of Testing," Proc. IRE, vol. 38, sec. 7.3, p. 945; August, 1950.

[5] Keith Henney, "Radio Engineering Handbook," McGraw-Hill Book Co., Inc., New York, N. Y., 1st ed., 1933; C. Steinmetz, "Theory and Calculation of Transient Alternating Current," p. 117.

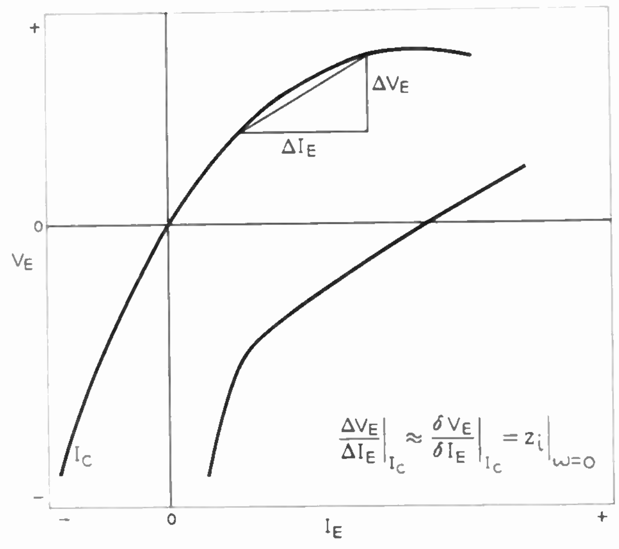


Fig. 6—Graphical determination of $z_i|_{\omega=0}$.

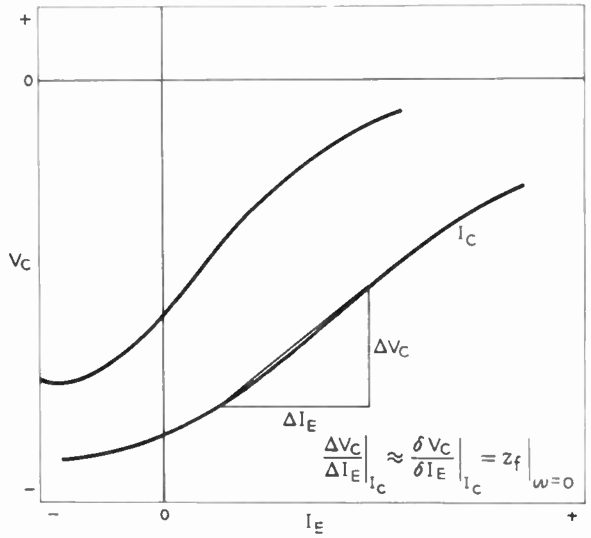


Fig. 8—Graphical determination of $z_f|_{\omega=0}$.

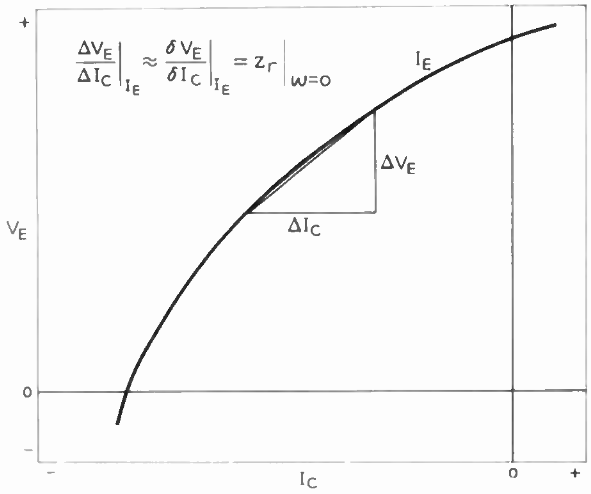


Fig. 7—Graphical determination of $z_r|_{\omega=0}$.

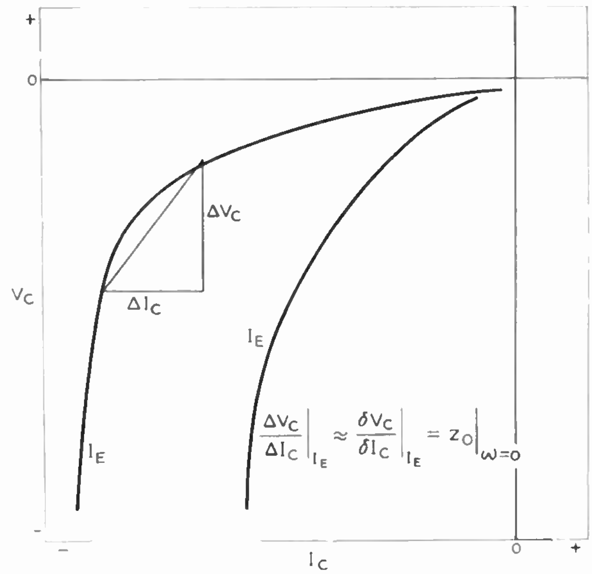


Fig. 9—Graphical determination of $z_o|_{\omega=0}$.

conditions of termination by either a voltmeter-ammeter, by bridge methods, or by graphical calculations made upon the measured static characteristics.

3.5.2 Graphical Calculations: The low-frequency values of the parameters can be determined approximately by graphical calculations on the input, output, and transfer characteristics, The data obtained from graphical methods are inherently of low precision and should be used only as an approximate check. In the common-base configuration, the static characteristics may be taken as shown in Fig. 2.

3.5.2.1 Impedance Parameters: The impedance parameters may be obtained from he static characteristics as shown in Figs. 6–9 above.

3.5.2.2 Admittance Parameters: The admittance parameters may be obtained from the static characteristics as shown in Figs. 10–13, opposite.

3.5.2.3 Hybrid Parameters: The hybrid parameters may be obtained from the static characteristics as shown in Figs. 14–17, p. 1550.

3.5.3 AC Ammeter and Voltmeter Measurements: The absolute magnitude of the measured parameter can be determined by ac ammeter-voltmeter measurements. For these measurements an adequately small alternating current or voltage of suitable frequency is injected at the input or output terminal. The alternating current of interest is then measured by measuring the voltage appearing across a small nonreactive resistor; the alternating voltage of interest is measured by a high-impedance voltmeter such as a vacuum-tube voltmeter. The magnitude of the particular parameter at the frequency chosen is determined by taking the ratio of the appropriate currents and voltages. This technique is generally applicable, particularly to sweep methods where a voltage or current is held constant; the dependent current or voltage is then presented on an oscilloscope as a function of frequency or test voltage, or test current.

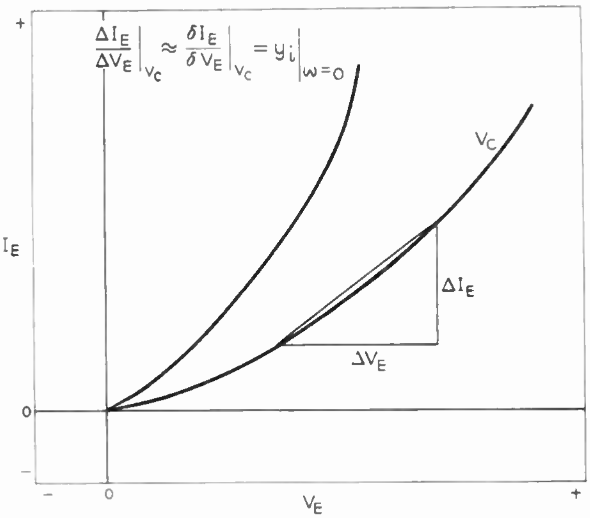


Fig. 10—Graphical determination of $y_i|_{\omega=0}$.

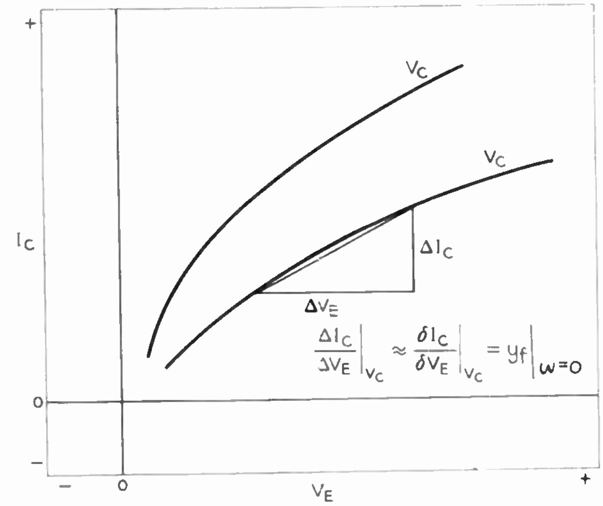


Fig. 12—Graphical determination of $y_f|_{\omega=0}$.

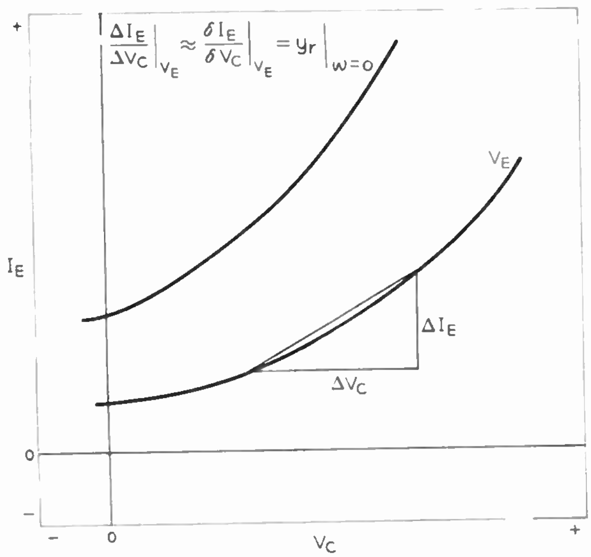


Fig. 11—Graphical determination of $y_r|_{\omega=0}$.

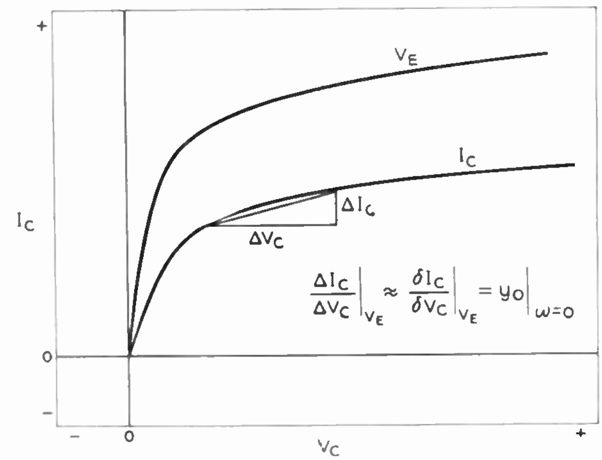


Fig. 13—Graphical determination of $y_o|_{\omega=0}$.

3.5.3.1 Voltage and Current Measurements Including Phase Angle: At frequencies where reactive and transit time effects are not negligible, phase-angle information is necessary for complete specification of the parameter. The usual practice is to compare the phase angle of voltages. Thus if the phase angle of currents is desired, the currents to be measured are applied to nonreactive resistors, and voltages proportional to and in phase with these currents are obtained.

The following methods of measurement of phase angle are most common (listed in the order of complexity of instrumentation).

3.5.3.2 Oscilloscope Method: The voltages to be compared are applied to the appropriate set of deflection plates and the phase angle is determined from the resulting Lissajous figures.[6,7] Care must be taken that the phase shift of the oscilloscope and the associated amplifiers at the frequency of measurement is negligible.

3.5.3.3 Pulse and Square-Wave Methods: The voltages to be compared are transformed into sharp pulses or into square waves. The lead or lag of the edges of the pulses or square waves can be compared on an oscilloscope, in trigger circuits, etc. Commercial phase meters generally employ this or a similar principle.[6,8]

3.5.3.4 The Harmonic Multiplier or Subdivider Method: One of the voltages to be compared is applied to a harmonic multiplier or subdivider and the fundamental and resulting harmonic or subharmonic signal is applied to pairs of deflection plates of an oscilloscope. The phase angle may be determined from the intersections of the multiple Lissajous figure.[6,9] With proper precautions high accuracies are attainable (better than 0.1 degree) at single frequencies.

3.5.3.5 The Heterodyne Method: The voltages to be compared are heterodyned in mixers with a beating

[6] F. E. Terman and J. M. Pettit, "Electronic Measurements," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 267–275; 1952.

[7] D. Bagno and A. Barnett, "Cathode ray phase meter," *Electronics*, vol. 11, p. 24; January, 1938.

[8] E. R. Kretzmer, "Measuring phase at audio and ultrasonic frequencies," *Electronics*, vol. 22, p. 114; October, 1949.

[9] M. F. Wintle, "Precision calibrator for a low frequency phase meter," *Wireless Engr.*, vol. 23, p. 197; July, 1951.

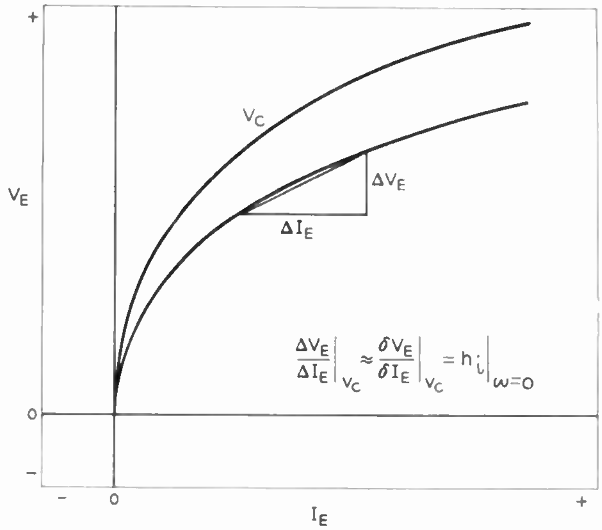


Fig. 14—Graphical determination of $h_i|_{\omega=0}$.

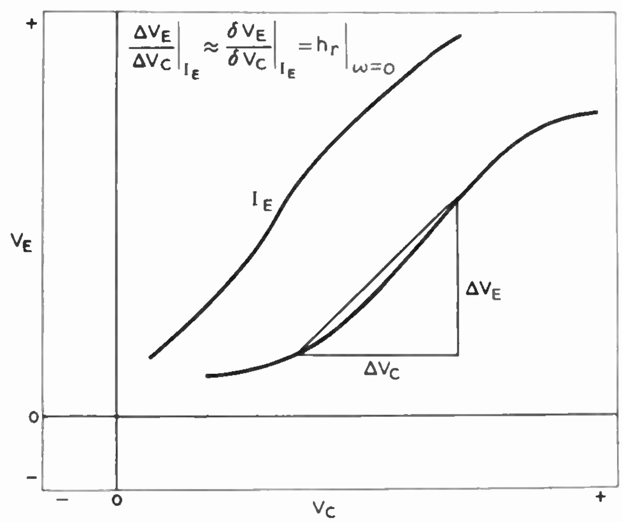


Fig. 15—Graphical determination of $h_r|_{\omega=0}$.

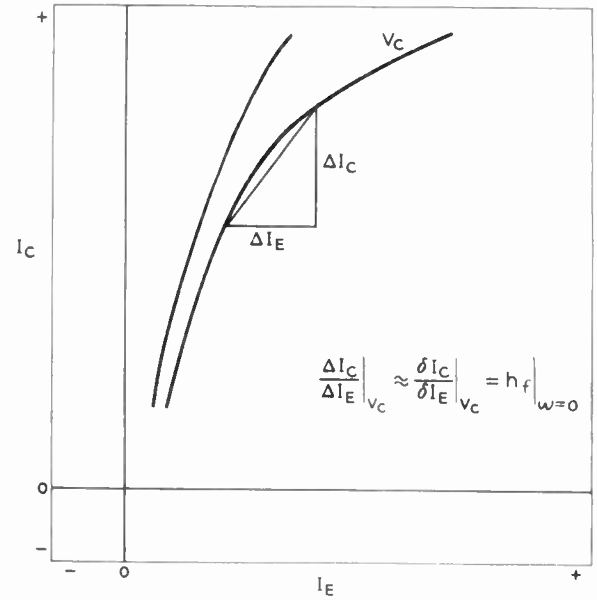


Fig. 16—Graphical determination of $h_f|_{\omega=0}$.

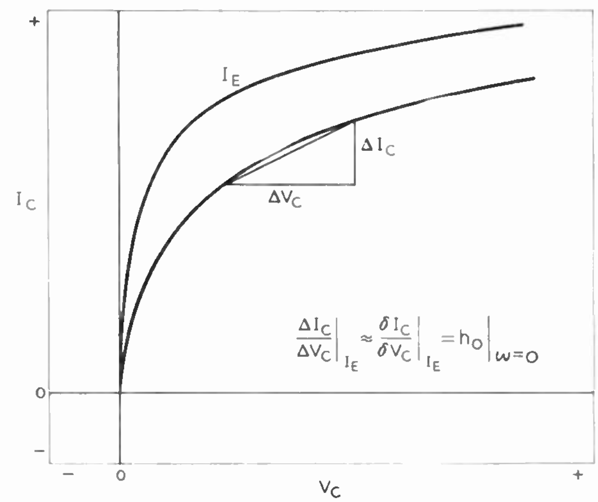


Fig. 17—Graphical determination of $h_o|_{\omega=0}$.

frequency, offset by a fixed amount from the signal frequency. The resulting beat-frequency signals are then compared as to phase at a fixed frequency. Depending on the accuracy desired, the phase comparator may employ cathode-ray oscillograph comparison, harmonic generator, pulse and square-wave methods, phase discriminators, calibrated phase shifters, etc. This method is capable of extreme accuracies and is practical over the entire frequency spectrum.[6,10–13]

3.5.4 Bridge Methods: A parameter may be measured by a suitable bridge[13] under the specified conditions of termination. In general, the bridge method is capable of determination of real and reactive components of the measured characteristics, at the specified frequency or frequencies. It is most applicable to point-by-point measurements, and does not lend itself readily to frequency sweep measurements.

3.5.5 Open-Circuit Parameters: The transistor may be described by the 4-terminal network shown in section 1.3 and in (1) and (2).

3.5.5.1 Equivalent Circuits: The device represented by the circuit equations of (1) and (2) may be represented by either one- or two-generator equivalent circuits as shown in Fig. 18.

3.5.5.2 Measurement of Input Impedance z_i: The open-circuit input impedance z_i may be measured by voltmeter-ammeter or bridge methods. A voltmeter-ammeter method from which the complex magnitude, but not the phase angle, may be derived is shown in Fig.

[10] M. Levy, "Measuring phase at audio and ultrasonic frequencies," *Elec. Commun.*, vol. 18, p. 206; January, 1940.

[11] D. A. Alsberg, "Principles and applications of converters for high frequency measurements," PROC. IRE, vol. 40, pp. 1195–1203; October, 1952.

[12] D. A. Alsberg and D. Leed, "A precise direct reading phase and transmission measuring system for video frequencies," *Bell Sys. Tech. J.*, vol. 28, pp. 221–238; April, 1949.

[13] B. Hague, "Alternating Circuit Bridge Methods," Isaac Pitman and Sons, Ltd., London, Eng., 5th ed.; 1943.

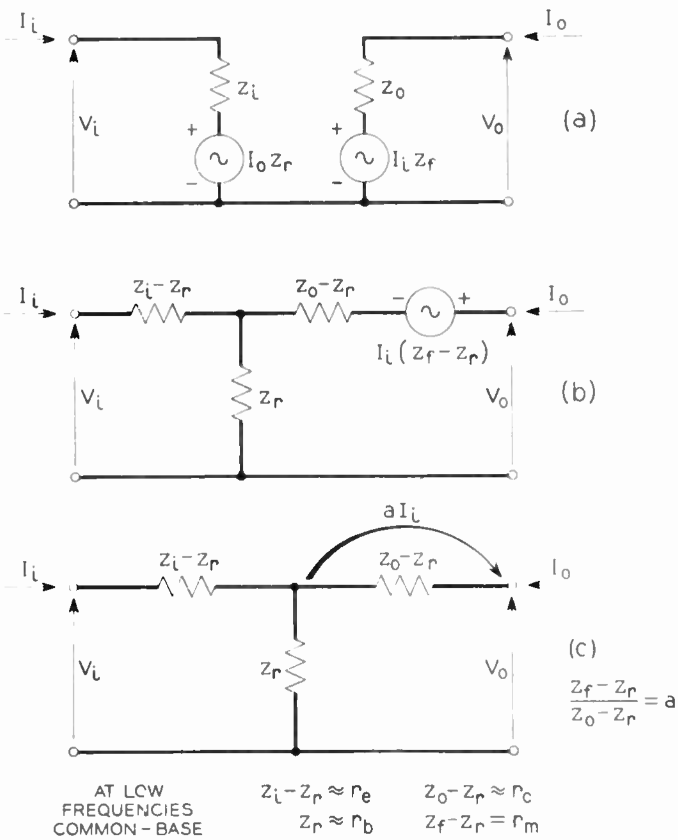


Fig. 18—Open-circuit impedance equivalent circuits. (a) Two generator. (b) One generator. (c) One generator.

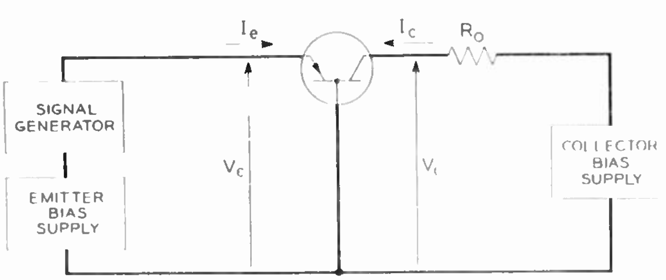


Fig. 19—Measurement of z_i or z_f.

19. It is valid only where z_o is small compared to the shunt reactance and internal impedance of the collector bias supply.

Note 1: A suitable low frequency lies in the range from 100 to 6000 cycles per second for point contact devices, and 100 to 400 cps for junction devices.

Note 2: Where z_o is very large (as in junction transistors) the measurement of z_i should not be attempted. A measurement of h_i as described in section 3.5.7 is preferable.

3.5.5.3 Measurement of Reverse Transfer Impedance z_r: The open-circuit reverse transfer impedance z_r may be measured by a bridge method, or by the voltmeter-ammeter method shown in Fig. 20.

Taking care to make the ac collector current I_c small

$$z_r = \frac{V_e}{I_c} \;(\approx r_b \text{ where the frequency is low}).$$

3.5.5.4 Measurement of Output Impedance z_o: The open-circuit output impedance z_o may be measured in the circuit shown in Fig. 20.

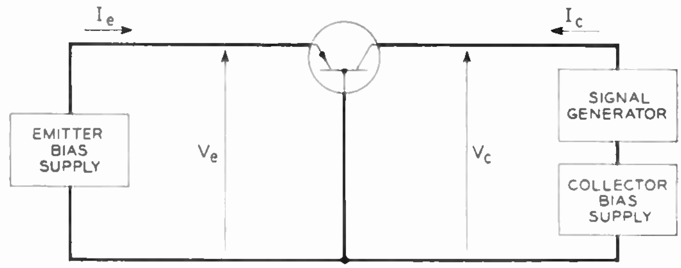


Fig. 20—Measurement of z_o or z_r.

$$z_o = \frac{V_c}{I_c} \;(\approx r_b + r_c \text{ where the frequency of measurement is low}).$$

3.5.5.5 Measurement of Transfer Impedance z_f: The open-circuit forward transfer impedance z_f may be measured in the circuit shown in Fig. 19 or by a suitable bridge method.

$$z_f = \frac{V_c}{I_e} \;(\approx r_b + ar_c \text{ where the frequency of measurement is low}).$$

Note: In the case of junction transistors the high value of z_o makes z_f and z_r measurement difficult because of stray capacitance effects. A more satisfactory method is to measure the short-circuit-current-transfer ratio h_f directly, and compute the desired value from the other three measurable parameters, since $z_f = h_f z_o$.

3.5.5.6 Measurement of Short-Circuit Current Transfer Ratio h_f (or $-\alpha_f$):

Note: The term α is also used to define the short-circuit forward current transfer ratio, but as it is subject to ambiguous interpretation h_f is used.

The short-circuit-current-transfer ratio h_f may be measured by many methods, two of which are shown in Figs. 21 and 22 (next page). For $|h_f| \leq 1$ and a low test frequency, the circuit shown in Fig. 21 may be used. Care must be taken that the phase characteristic of h_f does not cause a substantial error. The short-circuit current transfer ratio in the common base configuration h_{fb} may be expressed in terms of the short-circuit transfer ratio, common emitter, h_{fe}, where

$$h_{fb} = -\left(\frac{h_{fe} + h_{ie}h_{oe} - h_{re}h_{fe}}{1 + h_{fe} + h_{ie}h_{oe} - h_{re}h_{fe} - h_{re}}\right) \cong \frac{\frac{V_2}{V_1}}{\frac{R_2}{R_1} + \frac{V_2}{V_1}}.$$

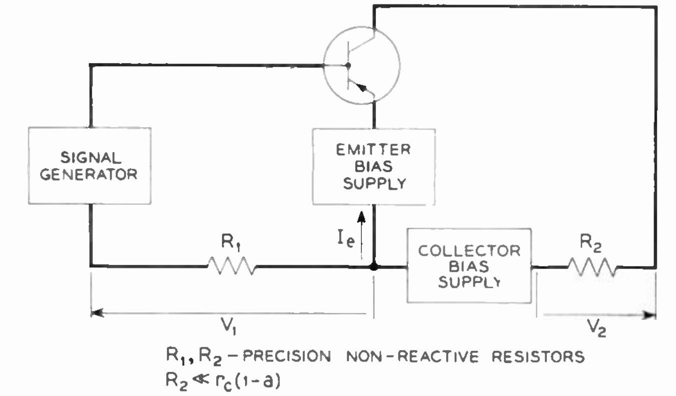


Fig. 21—Measurement of h_{fe}.

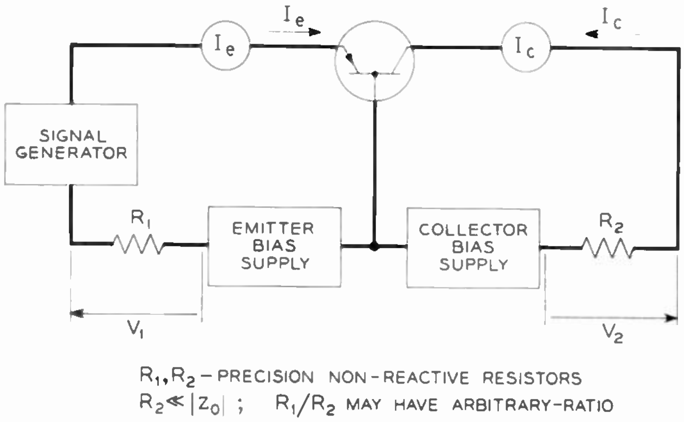


Fig. 22—Measurement of h_{fb}.

3.5.5.7 Measurement of h_f as a Function of Frequency: To measure $|h_f|$ the test circuit shown in Fig. 22 may be used, where

$$\left|h_f\right| = \frac{R_1}{R_2}\left|\frac{V_2}{V_1}\right| = \left|\frac{I_c}{I_e}\right|.$$

A common application of this measurement is to determine the frequency of α-cutoff.

3.5.5.8 Measurement of Phase Angle of h_f (Method 1): To measure the phase angle of h_f as a function of frequency, one method is to place reference voltages V_1 and V_2 on the horizontal and vertical plates of a suitable oscilloscope and by standard means convert the Lissajous figure information to a phase angle as in section 3.5.3.2.

3.5.5.9 Measurement of Phase Angle of h_f (Method 2): A second method is described in section 3.5.3.5.

3.5.5.10 Measurement of Output Capacitance C_o: The output capacitance C_o is the capacitance associated with the reactive component of h_o, which may be measured by a resonance method, as shown in Fig. 23, or by method shown in Fig. 24. C_o of the transistor is the difference in the settings of C_x when resonated with L, with the transistor in and out of the circuit, Fig. 23.

$$C_o = C_{x2} - C_{x1}$$

where C_o is a function of V_{CB}, I_C and frequency.

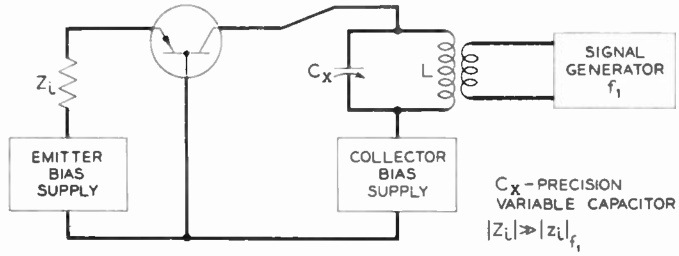


Fig. 23—Resonance method of measurement of $C_o \cdot Z_i$ must be nonreactive.

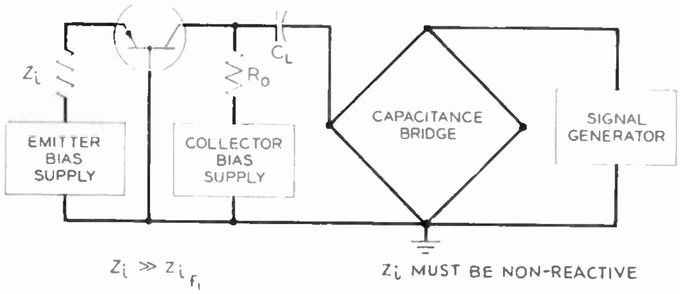


Fig. 24—Bridge method of measurement of $C_o \cdot Z_i$ must be nonreactive.

C_o of the transistor is the difference between capacitance bridge reading with the transistor in and out of circuit, Fig. 24.

3.5.6 Short-Circuit Admittance Parameters: A transistor may also be defined by the admittance equations (3) and (4).[14]

3.5.6.1 Equivalent Circuits: The input and output nodal equations (3) and (4) can be simply represented by a 2-generator equivalent circuit as shown in Fig. 25 or a 1-generator equivalent circuit as illustrated in Fig. 26.

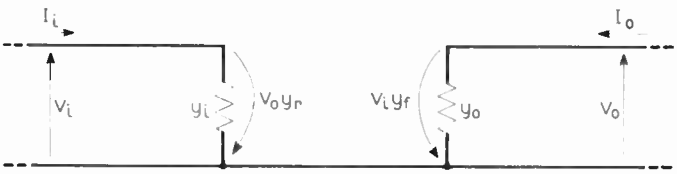


Fig. 25—Short circuit admittance, 2-generator equivalent circuit.

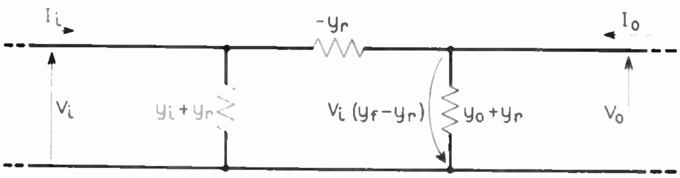


Fig. 26—Short-circuit admittance, 1-generator equivalent circuit.

3.5.6.2 Test Methods.

3.5.6.2.1 AC Ammeter and Voltmeter Measurement: The absolute magnitude of the admittance parameters

[14] L. C. Peterson, "Equivalent circuits of linear active four-terminal networks," *Bell Sys. Tech. J.*, vol. 27, pp. 593–622; October, 1948.

can be determined by ac ammeter-voltmeter measurements. For these measurements, an alternating voltage of suitable frequency is connected to the input terminal or output terminal. The appropriate alternating current is determined by measuring the voltage appearing across a small nonreactive resistor. The magnitude of the particular admittance parameter at the frequency chosen is determined by taking the ratio of the measured current to the applied voltage. This method of measurement is illustrated in Figs. 27 and 28 for the input self-admittance and forward-transfer admittance respectively. The output self-admittance and the reverse-transfer admittance may be measured by methods similar to Fig. 27 and Fig. 28 respectively.

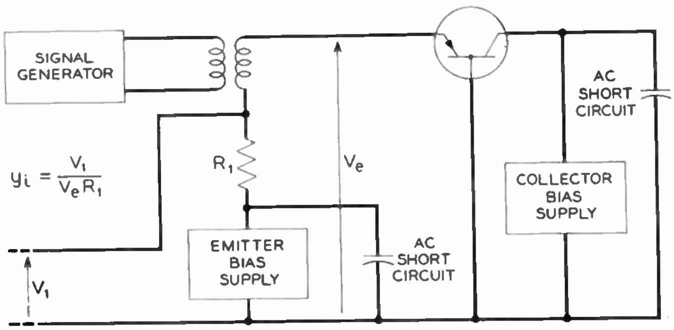


Fig. 27—Measurement of y_i.

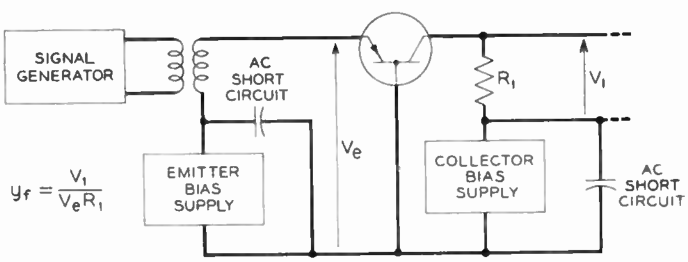


Fig. 28—Measurement of y_f.

In general, if the ac ammeter-voltmeter method is used to determine the conductance parameters, the frequency of the test signal must be chosen sufficiently low so that the susceptance parameter is negligible. Care must be taken to insure that the voltage drop across the resistor R_1 is negligibly small.

3.5.6.2.2 Bridge Measurements: The most accurate method for determination of the admittance parameters is by use of a suitable bridge.[13,15,16] For accurate measurements, the bridge circuits employed must be capable of balancing both the conductance and susceptance parameters simultaneously, although the bridge need be calibrated for only the component desired.

A typical simplified bridge circuit for measuring the admittance parameters and certain of their ratios is shown in Fig. 29. The bridge connection shown in Fig.

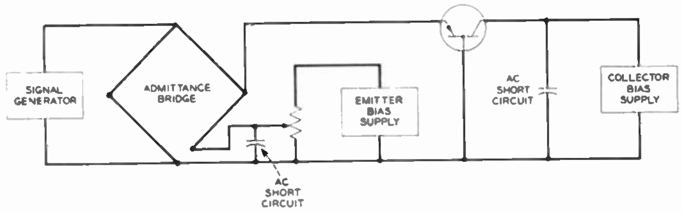


Fig. 29—Bridge method of measurement of y_i.

29 may be used to measure y_i; if the emitter and collector connections, and bias supplies, are reversed then it may be used to measure y_o. The measurement of y_f and y_r can be performed on more complex bridges.

3.5.7 Hybrid Parameters: The hybrid parameters defined by (5) and (6) are of value to all transistors. Since the measurements are based upon open-circuit terminations across low self-impedance, and short-circuit terminations across high self-impedances, the errors due to nonideal terminations are minimized.[17]

3.5.7.1 Equivalent Circuit: A convenient equivalent circuit for the device represented by (5) and (6) is shown in Fig. 30.

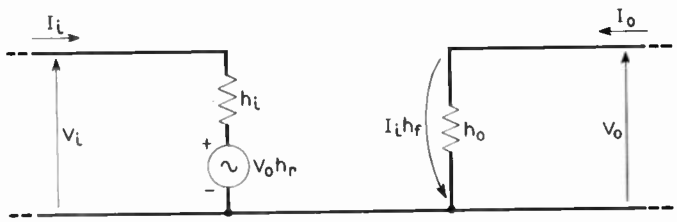


Fig. 30—Hybrid parameter, 2-generator equivalent circuit.

3.5.7.2 Measurement of Input Impedance h_i: The short-circuit input impedance h_i may be measured by the procedure of section 3.5.5.2 except that the output is ac short-circuited in place of the ac open-circuit. Or, it may be measured by the procedures of sections 3.5.6.2.1 and 3.5.6.2.2, noting that $h_i = 1/y_i$.

3.5.7.3 Measurement of Reverse Transfer Ratio h_r: The open-circuit reverse transfer ratio h_r may be measured by the procedure outlined in section 3.5.5.3, since $h_r = V_e/V_c$.

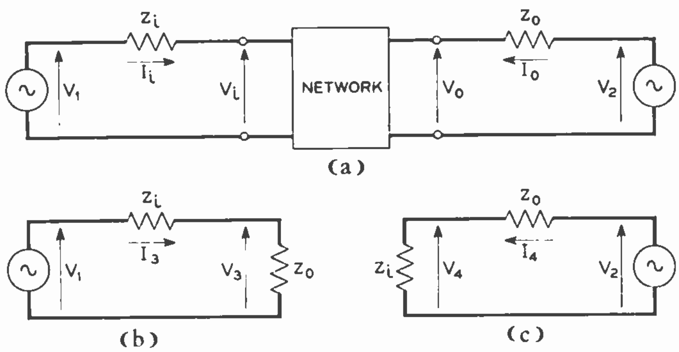
3.5.7.4 Mesaurement of Forward Transfer Ratio h_f (or $-\alpha_f$): The short-circuit forward transfer ratio h_f may be measured by the procedure detailed in sections 3.5.5.6 through 3.5.5.9. This parameter is an important physical characteristic of all transistors and the variation as a function of frequency is important in circuit application.

3.5.7.5 Measurement of Output Admittance h_o: The open-circuit output admittance h_o is measured by the procedure detailed in section 3.5.5.4 noting that $h_o = 1/z_o$, and in section 3.5.6.2.2 except that an open circuit is substituted for the short circuit across the input terminals.

[15] W. N. Tuttle, "Dynamic measurements of electron-tube coefficients," PROC. IRE, vol. 21, pp. 844–857; June, 1933.
[16] L. J. Giacoletto, "Bridges for measuring junction transistor admittance parameters," *RCA Rev.*, vol. 14, pp. 269–296; June, 1953.

[17] H. G. Follingstad, "An analytical study of *z*, *y*, and *h* parameter accuracies in transistor sweep measurement," 1954 IRE CONVENTION RECORD, Part 3, pp. 104–116.

3.5.8 Finite Termination Parameters: When it is impractical to satisfy the termination conditions of an open or short circuit (*e.g.*, in a frequency or parameter sweep) recourse may be made to a finite termination in accordance with the terminology of Fig. 31. Note that the symbols used are similar to, but not interchangeable with, those used previously.



$$\text{(a)}\quad z_{in} = \left.\frac{V_i}{I_i}\right|_{V_2=0} \qquad \text{(b)}\quad I_3 = \frac{V_1}{Z_i + Z_o} \qquad \text{(c)}\quad I_4 = \frac{V_2}{Z_i + Z_o}$$

$$y_{out} = \left.\frac{I_o}{V_o}\right|_{V_1=0} \qquad V_3 = \frac{-V_1 Z_o}{Z_i + Z_o} \qquad V_4 = \frac{-V_2 Z_i}{Z_i + Z_o}$$

$$e^{-\theta} = \left.\frac{-I_o}{I_3}\right|_{V_2=0} = \left.\frac{V_o}{V_3}\right|_{V_2=0}$$

$$e^{-\theta_r} = \left.\frac{-I_i}{I_4}\right|_{V_1=0} = \left.\frac{V_i}{V_4}\right|_{V_1=0}$$

Fig. 31—Finite termination parameter definitions.

The input impedance z_{in} is the impedance between the input terminals when the output terminals are terminated with Z_o and with $V_2=0$. The forward insertion transmission $e^{-\theta_f}$ is the ratio of the currents flowing through (or voltages across) Z_o with the transistor inserted between Z_i and Z_o and the transistor removed and replaced by a short circuit as shown in Fig. 31 and with $V_2=0$. The reverse insertion transmission $e^{-\theta_r}$ is the ratio of the currents flowing through (or voltages across) Z_i with the transistor inserted between Z_i and Z_o and the transistor removed and replaced by a short circuit as shown in Fig. 31 and with $V_1=0$. The output admittance y_{out} is the admittance between the output terminals when the input terminals are terminated with Z_i and with $V_1=0$.

3.5.8.1 Measurement of Forward and Reverse Insertion Transmission: Forward and reverse insertion transmission may be measured by methods outlined in the literature.[12,18]

3.5.8.2 Measurement of Input Impedance and Output Admittance: The input impedance z_{in} and the output admittance y_{out} may be measured by the methods outlined in sections 3.5.7.2 and 3.5.7.5 except that the output short circuit is replaced by Z_o and the input open circuit by Z_i respectively.

[18] F. E. Terman and J. M. Pettit, "Electronic Measurements," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 297–306, 312–316, 117, 177–180; 1952.

In addition, the measuring circuit required for section 3.5.8.1 may be used to measure impedance and admittance using a hybrid coil and measuring reflection coefficient and phase,[18] or by using the insertion loss and phase principle.[19]

3.5.9 Relations between z, y, and h and Finite Termination Parameters: Any set of parameters in sections 3.5.5 through 3.5.8 may be converted to any other set in those sections by the equations of Fig. 32, pp. 1555 and 1557.

3.6 Visual Displays

3.6.1 General: It is often desirable to obtain the value of small signal parameters as a function of frequency or operating point. To avoid the tedium of point-by-point measurements and to reduce the effects of instability with respect to time, curve tracer (swept) methods of measurement are used.[20,21] The requirements of these sweep methods constrain the realizability of terminating impedances more than point-by-point methods and thereby directly influence the choice of preferred sets of parameters.[17] The following major factors enter into the design of a swept measurement system.

3.6.2 Display Mechanism: Two types of display mechanism are in general use: recorders which trace the function being measured on paper using some form of stylus, such as pen and ink, chemical or pressure-sensitive styli, spark gaps, etc., and cathode-ray tube displays. Recorder display mechanisms are usually slow, but permit very high accuracies (often 0.1 per cent or better). The cathode-ray tube permits rapid displays. In accuracy it is limited by electron optics (spot size) and tube linearity. While some special cathode-ray tube types permit display accuracies in the 1 per cent and 2 per cent range, ordinary commercial cathode ray tubes are only capable of accuracies in the 5 per cent range.

3.6.3 Repetition Rates: The upper limit of repetition rates is determined by the speed of response of the display mechanism, the display bandwidth required, the termination realizability, and the frequency response of the transistor. The lower limit of repetition rates in cathode-ray tube displays is determined by flicker causing operator fatigue. A display repetition rate of less than 25 complete displays per second is usually found objectionable, Long-persistence cathode-ray tubes permit somewhat slower repetition rates, the actual rate depending on the characteristics of the phosphor used in the tube. It should be noted that in the case of the display of families of curves the entire family must be displayed within the minimum repetition rate. In recorder-type displays the lower limit to repetition rates

[19] D. A. Alsberg, "A precise sweep frequency method of vector impedance measurement," Proc. IRE, vol. 39, pp. 1393–1400; November, 1951.

[20] W. J. Albersheim, "Measuring techniques for broad-band, long distance radio relay systems," Proc. IRE, vol. 40, pp. 548–551; May, 1952.

[21] H. G. Follingstad, "A transistor alpha sweeper," 1953 IRE Convention Record, Part 9, pp. 64–71.

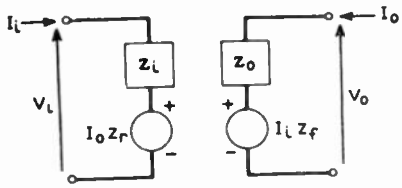
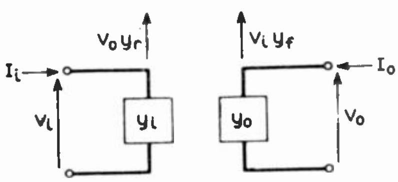
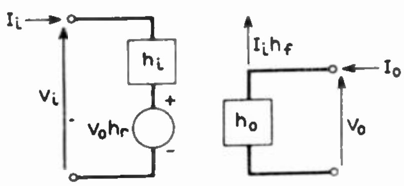
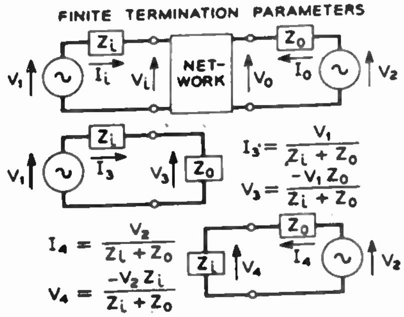
Symbolic Definitions		Word Definitions
OPEN-CIRCUIT IMPEDANCE PARAMETERS  $V_i = z_i I_i + z_r I_o$ $V_o = z_f I_i + z_o I_o$	$z_i = \left.\frac{V_i}{I_i}\right\|_{I_o=0}$	Input *Impedance* with AC Open-Circuited Output
	$z_r = \left.\frac{V_i}{I_o}\right\|_{I_i=0}$	Reverse Transfer *Impedance* with AC Open-Circuited Input
	$z_f = \left.\frac{V_o}{I_i}\right\|_{I_o=0}$	Forward Transfer *Impedance* with AC Open-Circuited Output
	$z_o = \left.\frac{V_o}{I_o}\right\|_{I_i=0}$	Output *Impedance* with AC Open-Circuited Input
SHORT-CIRCUIT ADMITTANCE PARAMETERS  $I_i = y_i V_i + y_r V_o$ $I_o = y_f V_i + y_o V_o$	$y_i = \left.\frac{I_i}{V_i}\right\|_{V_o=0}$	Input *Admittance* with AC Short-Circuited Output
	$y_r = \left.\frac{I_i}{V_o}\right\|_{V_i=0}$	Reverse Transfer *Admittance* with AC Short-Circuited Input
	$y_f = \left.\frac{I_o}{V_i}\right\|_{V_o=0}$	Forward Transfer *Admittance* with AC Short-Circuited Output
	$y_o = \left.\frac{I_o}{V_o}\right\|_{V_i=0}$	Output *Admittance* with AC Short-Circuited Input
HYBRID PARAMETERS  $V_i = h_i I_i + h_r V_o$ $I_o = h_f I_i + h_o V_o$	$h_i = \left.\frac{V_i}{I_i}\right\|_{V_o=0}$	Input *Impedance* with AC Short-Circuited Output
	$h_r = \left.\frac{V_i}{V_o}\right\|_{I_i=0}$	Reverse Transfer *Voltage Ratio* with AC Open-Circuited Input
	$h_f = \left.\frac{I_o}{I_i}\right\|_{V_o=0}$	Forward Transfer *Current Ratio* with AC Short-Circuited Output
	$h_o = \left.\frac{I_o}{V_o}\right\|_{I_i=0}$	Output *Admittance* with AC Open-Circuited Input
FINITE TERMINATION PARAMETERS  $I_3 = \frac{V_1}{Z_i + Z_o}$ $V_3 = \frac{-V_1 Z_o}{Z_i + Z_o}$ $I_4 = \frac{V_2}{Z_i + Z_o}$ $V_4 = \frac{-V_2 Z_i}{Z_i + Z_o}$	$z_{in} = \left.\frac{V_i}{I_i}\right\|_{V_2=0}$	Input *Impedance* with Output Terminated in Z_o
	$e^{-\theta_r} = -I_i/I_4\|_{V_1=0} = V_i/V_4\|_{V_1=0}$	Reverse *Insertion Transmission* Between Z_i and Z_o
	$e^{-\theta_f} = -I_o/I_3\|_{V_2=0} = V_o/V_3\|_{V_2=0}$	Forward *Insertion Transmission* Between Z_i and Z_o
	$y_{out} = \left.\frac{I_o}{V_o}\right\|_{V_1=0}$	Output *Admittance* with Input Terminated in Z_i

Fig. 32—Parameter conversion table (*cont'd next page*)

Relations between Parameters

z Parameter	y Parameter	h Parameter	Finite Termination Parameter
z_i	$\frac{1}{y_i(1-\tau)}$	$\frac{h_i}{1-\tau}$	$\frac{Y^2Z^2e^{-\theta_f}e^{-\theta_r}}{\Gamma B^2(Y-\Gamma Y_o)}+\frac{Z-\Gamma Z_i}{\Gamma}$
z_r	$\frac{-y_r}{y_iy_o(1-\tau)}$	$\frac{h_r}{h_o}$	$\frac{YZe^{-\theta_r}}{B(Y-\Gamma Y_o)}$
z_f	$\frac{-y_f}{y_iy_o(1-\tau)}$	$-\frac{h_f}{h_o}$	$\frac{YZe^{-\theta_f}}{B(Y-\Gamma Y_o)}$
z_o	$\frac{1}{y_o(1-\tau)}$	$\frac{1}{h_o}$	$\frac{\Gamma}{Y-\Gamma Y_o}$
$\frac{1}{z_i(1-\tau)}$	y_i	$\frac{1}{h_i}$	$\frac{\Gamma}{Z-\Gamma Z_i}$
$\frac{-z_r}{z_iz_o(1-\tau)}$	y_r	$-\frac{h_r}{h_i}$	$\frac{YZe^{-\theta_r}}{B(Z-\Gamma Z_i)}$
$\frac{-z_f}{z_iz_o(1-\tau)}$	y_f	$\frac{h_f}{h_i}$	$\frac{-YZe^{-\theta_f}}{B(Z-\Gamma Z_i)}$
$\frac{1}{z_o(1-\tau)}$	y_o	$\frac{h_o}{1-\tau}$	$\frac{Y^2Z^2e^{-\theta_f}e^{-\theta_r}}{\Gamma B^2(Z-\Gamma Z_i)}+\frac{Y-\Gamma Y_o}{\Gamma}$
$z_i(1-\tau)$	$\frac{1}{y_i}$	h_i	$\frac{Z}{\Gamma}-Z_i$
$\frac{z_r}{z_o}$	$-\frac{y_r}{y_i}$	h_r	$\frac{YZe^{-\theta_r}}{B\Gamma}$
$-\frac{z_f}{z_o}$	$\frac{y_f}{y_i}$	h_f	$\frac{-YZe^{-\theta_f}}{B\Gamma}$
$\frac{1}{z_0}$	$y_o(1-\tau)$	h_o	$\frac{Y_o}{\Gamma}-Y_o$
$z_i\left[1-\frac{\tau}{1+Z_o/z_o}\right]$	$\frac{1}{y_i\left[1-\frac{\tau}{1+Y_o/y_o}\right]}$	$h_i\left[1-\frac{\tau/(\tau-1)}{1+Y_o/h_o}\right]$	z_{in}
$\frac{z_rz_o\left[\frac{Z_i+Z_o}{Z_o}\right]}{z_o(1+z_oY_o)[z_i(1-\tau)+Z_i]+z_rz_f}$	$\frac{-y_ry_i\left[\frac{Z_i+Z_o}{Z_o}\right]}{y_i(1+y_iZ_i)[y_o(1-\tau)+Y_o]+y_ry_f}$	$\frac{h_r\left[\frac{Z_i+Z_o}{Z_o}\right]}{(h_i+Z_i)(h_o+Y_o)-h_rh_f}$	$e^{-\theta_r}$
$\frac{z_fz_o\left[\frac{Z_i+Z_o}{Z_o}\right]}{z_o(1+z_oY_o)[z_i(1-\tau)+Z_i]+z_rz_f}$	$\frac{-y_fy_i\left[\frac{Z_i+Z_o}{Z_o}\right]}{y_i(1+y_iZ_i)[y_o(1-\tau)+Y_o]+y_ry_f}$	$\frac{-h_f\left[\frac{Z_i+Z_o}{Z_o}\right]}{(h_i+Z_i)(h_o+Y_o)-h_rh_f}$	$e^{-\theta_f}$
$\frac{1}{z_o\left[1-\frac{\tau}{1+Z_i/z_i}\right]}$	$y_o\left[1-\frac{\tau}{1+Y_i/y_i}\right]$	$h_o\left[1-\frac{\tau/(\tau-1)}{1+Z_i/h_i}\right]$	y_{out}

Fig. 32 (*cont'd top of next page*)

Auxiliary Symbols

$$\tau = \frac{z_f z_r}{z_i z_o} = \frac{y_f y_r}{y_i y_o} = \frac{1}{1 - \dfrac{h_i h_o}{h_f h_r}} = \frac{1}{1 + \dfrac{(Z - \Gamma Z_i)(Y - \Gamma Y_o)B^2}{Y^2 Z^2 e^{-\theta_f} e^{-\theta_r}}}$$

$$Y = y_{\text{out}} + Y_O$$

$$Z = z_{\text{in}} + Z_i$$

$$B = \frac{Z_i + Z_o}{Z_o}$$

$$\Gamma = 1 + \frac{YZe^{-\theta_f}e^{-r}}{B^2}$$

The Value of Each Symbols May be Substituted in the Parameters Above

Fig. 32 (*conclusion*).

is set by the time domain stability of the transistor and the test circuit.

3.6.4 Display Bandwidth: In order to portray faithfully rapid changes in parameter value vs small changes in operating point or frequency, the test circuit and display mechanism must have adequate frequency response to respond to a sufficient number of harmonics of the repetition rate.

Insufficient display bandwidth is one of the most common failings of sweep test equipment. The display bandwidth required is determined by the maximum slope which must be faithfully portrayed. As a rule of thumb, if the maximum rise time is r, the display bandwidth required is of the order of $2/r$.[22] An approximate estimate of the rise time required may be obtained from the analysis of known transistor data such as static characteristics. The display of both forward and return trace is a safeguard against insufficient bandwidth as well as against undesired phase shifts, crosstalk, and pickup. Insufficient bandwidth is revealed by hysteresis-like separation of forward and return trace. Where transistor hysteresis is suspected, the proper operation of the test equipment may be verified by use of passive networks ("dummy transistors") having response slopes similar to the class of transistors being investigated. Sinusoidal sweep is preferred to other types as it is easily obtained and permits ready use of the retrace feature. Triangular sweeps must safeguard against ringing. Sawtooth sweeps make optimum use of available display time but do not permit use of the retrace.

3.6.5 Termination Realizability: Known, constant value terminations must be realized over broad frequency bands for parameter vs frequency measurements. Terminations must remain essentially invariant over a frequency range of twice the required display bandwidth centered on the probing signal frequency for parameter vs operating point displays. In the first case, unavoidable parasitic elements limit realizable broadband terminations; in the second case, parasitic elements and practical component size limit realizability.

[22] G. E. Valley, Jr., and H. Wallman, "Vacuum Tube Amplifiers," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., ch. 2, p. 71; 1948.

3.6.6 Parameter Vs Frequency: The parameter is selected by the choice of transistor input and output terminals, biases, and terminating conditions as outlined in sections 3.2 through 3.5. The input to output amplitude ratio and phase difference are a measure of the parameter value at the instantaneous frequency displayed.

A parameter vs frequency curve tracer consists of three basic units: the variable frequency source (oscillator), the terminating and biasing arrangement, and the detector and display mechanism; see Fig. 33.[23]

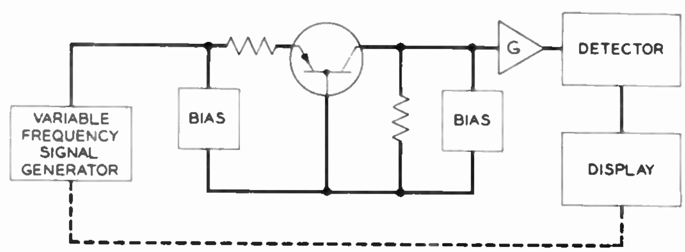


Fig. 33—Parameter vs frequency curve tracer.

3.6.6.1 Swept-Frequency Oscillator: The frequency of a swept-frequency oscillator is commonly varied by electronic or mechanical means in a sinusoidal, triangular, or sawtooth fashion. Provision is usually made to synchronize the display mechanism with the oscillator sweep rate.

3.6.6.2 Biasing and Terminating Arrangement: The transistor biasing currents or voltages may be introduced in parallel or in series with the transistor terminations. Extreme care must be taken to insure that the variations in bias circuit impedance and the associated circuits are small in effect on the parameter being measured. In the high-frequency ranges it is necessary to consider carefully the effect of all parasitic elements, which may include the transistor terminals.

3.6.6.3 Detector: The detector may be of the broadband untuned or the selective self-tuned variety. Nor-

[23] O. Kummer, "A transistor frequency scanner," 1954 IRE Convention Record, Part 10, pp. 81–87.

mally a broad-band detector is preceded by a broad-band amplifier to reduce effects of detector noise. Because of its relative simplicity the broad-band detector is used wherever possible. When the signal-to-noise ratio of the broad-band detector becomes objectionable, or when phase shift must be displayed, self-tuned heterodyne detection is used; see Fig. 34.[12]

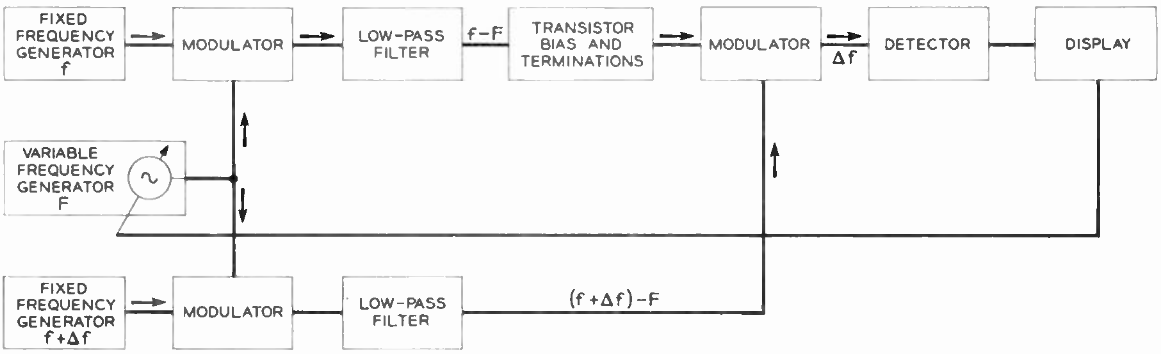


Fig. 34—Curve tracer with self-tuned heterodyne detection.

3.6.7 Parameter Vs Operating Point: All small-signal parameters may be displayed as a function of operating point using similar techniques; *e.g.*, a most common display of parameter vs operating point is that of alpha ($\alpha = -h_f$) vs emitter current.[21] The usual procedure and precautions as described in section 3.2 and 3.5 must be observed in order to make the characteristics of the unit under test as independent of the test cricuit as possible. The ratio of the input to output amplitude determines the magnitude of the parameters as a function of the operating point. The basic circuit for visual display of parameter vs operating point is shown in Fig. 35.

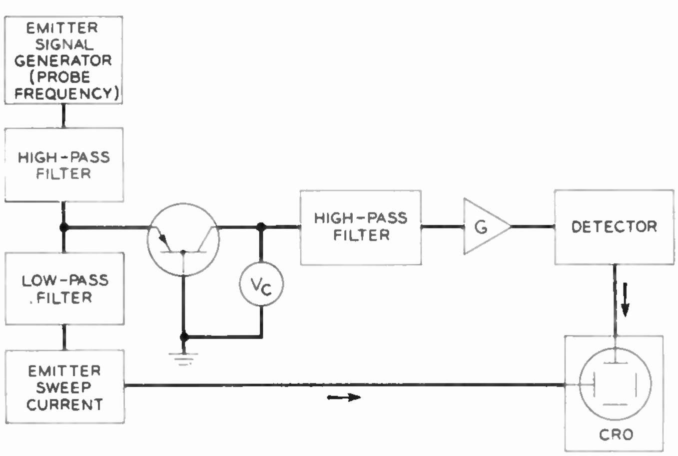


Fig. 35—Parameter vs operating point curve tracer.

3.6.7.1 Emitter Signal Current Source: The frequency of this signal generator is determined to a large extent by the display bandwidth and the filter design. A suitable probe frequency is determined by the basic repetition display bandwidth required and economic filter design. For a display bandwidth of 10 kc a probe frequency of 100 kc is suitable. Care should be taken to obtain amplitude stability and high output impedance. Since the probe signal amplitude determines in part the accuracy of the display, it should be kept as small as possible.

3.6.7.2 Emitter Sweep Current Source: The frequency of the sweep oscillator should be high enough to avoid eye fatigue (greater than 25 complete displays per second) and should not be a multiple or submultiple of line frequency. The amplitude must be sufficient to cover the range of operating points desired. Direct coupling should be used in connecting the sweep oscillator to the cathode-ray oscillograph in order to retain the display origin.

3.6.7.3 Amplifier: Since the emitter signal current source is of low amplitude and the collector load is a low impedance (in order to approach a short-circuit alpha measurement), it is necessary to amplify the collector signal before direct coupling to the cathode-ray oscillograph. A tuned amplifier of sufficient bandwidth may be used.[21]

3.6.7.4 Detector: The output from the amplifier is amplified and filtered in the detector, which is desirable, though only practical when using a high probe frequency such as 100 kc.

3.6.7.5 Calibration: Calibration of the display unit may be accomplished by connecting the driving source directly to the amplifier.

3.6.7.6 Precautions: The high and low pass filters should provide sufficient attenuation of the unwanted signal. The bandwidth of the display unit must be wide enough to prevent phase shift patterns which may be mistaken for test unit hysteresis. Suitable circuitry should be provided in order to prevent the unit under test from being swept too far into the cutoff region or any region in which the instantaneous power rating may be exceeded.

In testing point-contact transistors, care should be taken to avoid unwanted oscillations. The probe frequency should be 50 per cent or less of the cutoff frequency of the transistor for the connection used. Jitter and circuit noise should be maintained at levels which will result in a signal-to-noise ratio which is adequate for the measurement intended.

4.0 Environmental Tests

Environmental tests are performed under unique conditions of environment to obtain physical or electrical data resulting from or occurring during conditions of shock, vibration, temperature, humidity, or other environmental phenomena. Environmental tests are generally used to evaluate ratings, for the comparison of similar devices, or to determine performance relative to a specific application. The electrical tests performed on the device fall into two classes: 1) Precondition and postcondition parameter tests; 2) Tests made during a specific environmental condition.

4.1 Precautions

When electrical tests are used to evaluate mechanically-induced parameter shift, the data obtained are only as valid as the equipment and/or device repeatability. Care must be taken to minimize the effect of the environmental condition on the circuit associated with the device under test.

4.2 Temperature Coefficients

The temperature coefficient is the quotient of the difference between two parameter values divided by the corresponding temperature difference. The coefficient may be obtained over any linear portion of the parameter-temperature curve.

The time required to stabilize a parameter reading at any temperature is a function of:

1) The materials surrounding the device proper (potting wax, oil, etc.).
2) The heat conductivity of the internal connecting leads.
3) The heat generated within the device itself.

Care should be taken to avoid temperatures outside of the rated operating temperature range of the device. Permanent damage to the device under test may result if the storage temperature rating is exceeded.

4.3 Mechanical Tests

In mechanical tests a periodic or aperiodic accelerating force is applied to the device under test. The acceleration a is measured in g-units ($g=32.2$ feet per second per second).

For simple harmonic motion, a simple equation can be derived relating the acceleration a measured in g-units to frequency and displacement.

$$a = 0.0511Df^2 \text{ (}g\text{-units)}$$

where D = peak-to-peak displacement, inches, and f = frequency, cps.

Peak acceleration of aperiodic motion is usually calculated by determining the slope of the curve of velocity vs time. This can be obtained visually by photographic means or electrically by the proper choice of an accelerometer.

4.3.1 Shock Tests: In shock tests the device is subjected to a specified unidirectional acceleration for a specified time.

4.3.1.1 Orientation: To evaluate the effect of shock on the device, it is usually necessary to transmit the shock to the device along at least one direction of each of the three axes. The direction of the axis must be specified for each device configuration.

4.3.1.2 Precautions: The jig used to hold the device under test should be designed to exert the least possible constant stress. Care must be taken to limit the amount of cushioning material employed since the shock transmission characteristic of these materials is poor. Under conditions of high short-term acceleration even the metals, such as steel, must be regarded as highly viscous fluids.

When noise measurements are being made on the device under test, extreme care must be taken to avoid inducing extraneous signals in the moving lead wires. It is recommended that dummy resistive networks be used to prove out the associate test equipment before it is used to evaluate the device performance.

4.3.2 Vibration Tests: In vibration tests the device is subjected to an accelerating force whose amplitude varies sinusoidally with time. The tests are generally performed with as close to a true sine wave as practical to simplify calibration and analysis.

Vibration tests can be subdivided into three classes:

1) Fatigue vibration: tests to produce physical fatigue.
2) Mechanical resonance: tests to determine structural resonances.
3) Vibration: tests to evaluate performance under specific conditions relative to an application.

In all types of vibration tests the device should be vibrated in directions along each of the three axes.

4.3.2.1 Fatigue Vibration: The device should be tested at any specified single frequency for a specific time. Previbration and postvibration parameter tests are usually used to evaluate performance.

4.3.2.2 Mechanical Resonance: The device shall be tested over a range of the audio-frequency spectrum suitable for the intended application of the device. Mechanical resonance is determined by operating the device under test in a typical circuit and recording the noise vs vibration frequency characteristics. Distinct resonance in the device will usually result in successive noise bursts or an increase in noise figure at a specific frequency.

4.3.2.3 Vibration: The device is usually tested at a single vibration frequency of sufficient amplitude to evaluate adequately the performance relative to the application. Noise output and parameter shift are both used to evaluate performance. Frequency and amplitude are limited only by equipment considerations.

4.3.3 Acceleration Tests: Acceleration tests subject the device to a short-duration high-centrifugal acceleration. The device under test is commonly mounted in a semicompliant material (*e.g.*, nylon, teflon, etc.) to prevent excessive stresses from being generated at any point on the case or encapsulation, unless the application indicates other requirements. Care must be taken to balance the rotating wheel to minimize vibration.

4.4 Humidity Effects

The resistance to moisture penetration is primarily a function of the encapsulation. The ratio of penetration vs time is directly dependent on temperature since it is a function of the water vapor molecular activity.

4.4.1 Effects of Moisture: The effects of moisture vary with the type of semiconductor material. In general the most noticeable effect is an increase in reverse collector current with an open-circuit emitter. Such effects may be masked by any contamination present within the device and which may produce similar effects.

4.4.2 Humidity Testing: Precondition and postcondition tests will define the effect of moisture on the device. The device is usually subjected to a nominal 95 per cent relative humidity in conjunction with temperature cycling. "Dry" control lots should be run at least initially to determine separately the effect of the temperature cycling alone.

Where a true hermetic seal is not used, or where the effectiveness of a true hermetic seal is tested for check purposes, a wide-range temperature cycle and vibration are sometimes used to evaluate resistance to moisture penetration.

4.5 Radiation Susceptibility

Low-intensity radiation of short-time duration is effectively shielded by most encapsulations. Prolonged exposure to high-intensity radiation may produce permanent changes in the device parameters.

4.5.1 Types of Irradiation:

1) Electromagnetic irradiation, such as light and heat.
2) Alpha irradiation: doubly charged positive particles having a mass of 4.00, identical with helium atom nuclei.
3) Beta irradiation: high energy electrons.
4) Gamma irradiation: radiation similar to X rays but of shorter wavelength.
5) Neutron irradiation.

4.5.2 Considerations: Major effects encountered result from exposure to all frequencies in the electromagnetic spectrum. These effects may be controlled by the opacity of the encapsulation to the incident radiation.

Alpha and beta rays are not very penetrating. Gamma rays and neutrons are highly penetrating and damaging.

4.6 Pressure Effects

Pressure effects are changes in the device parameters resulting from the physical application of a stress to the device. The stress may be applied at a point or surface, or may take the form of changes in the surrounding atmosphere. The effect on the device parameters is wholly dependent on the transmission of any pressure exerted on the device encapsulation.

5.0 Noise Measurements

Deviation from Electron Devices Standards[24] are recommended only in those instances where characteristics unique to semiconductors justify such.

5.1 General

Semiconductor devices have frequency-dependent noise-producing mechanisms. This applies only to noise originating within the device under study, and should not include noise emanating from extraneous sources, such as described in the Standard cited.[24]

5.1.1 Noise Spectrum Analysis: The spectral energy distribution of the noise may be determined directly by the method shown in Fig. 36.

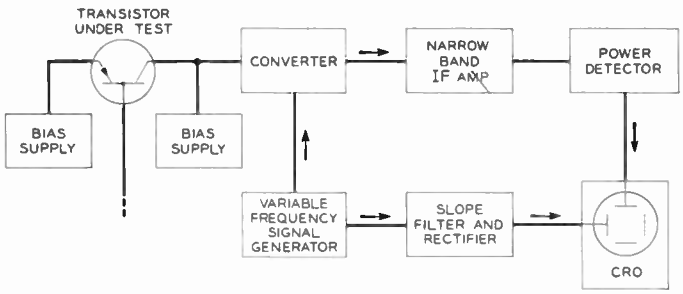


Fig. 36—Noise vs frequency curve tracer.

5.2 Spectral Energy Distribution

The spectral energy distribution must be obtained before measurement of the noise figure in order to ascertain that the center frequency and the bandwidth used to measure the noise figure will yield accurate, reproducible results. For example, a measurement of noise figure at f_1 of Fig. 37 would be misleading.

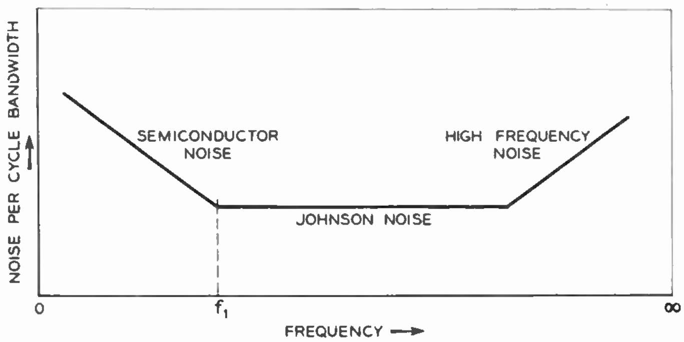


Fig. 37—Spectral distribution of noise energy.

5.3 Noise Factor

5.3.1 Measuring Time: A statement of the noise factor of a semiconductor should be accompanied by in-

[24] IRE Standards on Electron Devices; Methods of Measuring Noise, Proc. IRE, vol. 41, pp. 890–896; July, 1953.

formation as to either a) the time-constant of the measuring system or b) the time duration of observation. The latter applies to the case in which the output noise is recorded in a system of relatively rapid response with the rms level defined by equal areas above and below the median line.

5.3.2 Optimum Time Constants: Power detection should be accomplished by a vacuum thermocouple or a noise bolometer. If the time-constant adjustment is made at the power detector for noise factor measurements, the balance of the system within its own limitations may be used to indicate short-period pulses and other non-Gaussian properties of the device under test on a separate indicating or recording device.

5.3.3 Noise Bandwidth: Refer to section 10.1.2.1.1 of footnote 24. For measurement of average noise factor on devices exhibiting a large measure of dependence on frequency, the reference frequency f_o should be that frequency above and below which approximately one-half of the noise power is developed, for the specified measurement bandwidth.

5.3.3.1 Preferred Bands: Noise factors may be usually identified with some particular frequency characterizing a well-established usage application. Examples are: 1 kc, audio (both speech and music), 50 kc, typical carrier frequency, 500 kc, broadcast intermediate frequency, 1 megacycle, broadcast radio frequency, 30 megacycles, video intermediate frequency, etc. Preferred bands for average noise factor measurements cannot be recommended even for audio-frequency applications, because of the extremes in present acceptability standards. However, in specifying the noise factor, the frequency and the band over which the noise factor is measured should be specified.

5.3.4 Precautions: In addition to the precautions described in section 10.1.5 of footnote 24, the following sources of error must be noted: a substantial alteration of the noise band can result from the variation in emitter and collector capacitances in a transistor. Large measurement errors can result from the fact that the peaking factor for non-Gaussian noise exceeds that of white noise.

Common-Emitter Transistor Video Amplifiers*

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Summary—A design procedure and theory are given for the common-emitter transistor video amplifier with and without a feedback resistor in the emitter lead. In the analysis a junction transistor of the alloy type is represented by the Johnson-Giacoletto hybrid-pi equivalent circuit for the common-emitter transistor. The design theory accounts for the most significant part of the bilateralness of the transistor by adding a "Miller" capacitance term to the diffusion capacitance of the common-emitter transistor.

Gain-bandwidth products and optimum load resistors determined for a cascade of amplifier stages are reported. The figure of merit of the transistor in a cascade of identical video amplifier stages is compared with the figure of merit of a transistor used as a power amplifier.

The theory and design are given for the process of obtaining a maximally flat frequency response in a single stage by means of a capacitor shunting the feedback resistor, or by means of inductances in the amplifier interstages. Experimental results for the capacitor compensation scheme are given.

Introduction

THIS PAPER presents a theory of a design procedure for common-emitter video amplifiers. The theory is applicable and used on transistors of the alloy junction type as described by Mueller and Pankove.[1] In the analysis the transistor is represented by the Johnson-Giacoletto hybrid-pi equivalent circuit[2] for the common-emitter transistor. This circuit has proved to be both very useful and sufficiently exact for work of this kind. For optimum design of the video amplifiers considered here, the load resistors are so small that we can simplify the equivalent circuit by omitting some conductances that normally would be important in the design of audio amplifiers.

A major difficulty that often prevents simple and accurate analysis of transistor amplifiers is the bilateral nature of the device. For common-emitter video amplifiers, the bilateral portion is mostly due to the collector barrier capacitance. In this report the difficulties due to bilateralness are circumvented by lumping the *Miller* capacitance (the capacitive part of the *Miller* effect produced by the collector barrier capacitance) with the diffusion capacitance. Such lumping gives simple and accurate design procedures for video amplifiers.

* Original manuscript received by the IRE, March 14, 1956; revised manuscript received, June 28, 1956. This paper is part of a work carried out at Stanford Res. Inst. Menlo Park, California, while the writer was working on a one-year Research Fellowship granted by the National Academy of Sciences, Washington, in cooperation with Foreign Operations Agency.

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[1] C. W. Mueller and J. I. Pankove "A p-n-p alloy junction transistor for radio frequency amplification," Proc. IRE, vol. 42, pp. 386–391; February, 1954.

[2] L. J. Giacoletto, "Study of p-n-p alloy junction transistor from dc through medium frequencies," *RCA Review*, vol. 15, pp. 506–562; December, 1954.

Formulas are derived for the amplification and bandwidth of a simple resistance loaded amplifier stage and for a stage in an iterative cascaded amplifier. In the cascaded case a load resistor is found which gives the optimum gain-bandwidth product. This resistor is independent of the bandwidth wanted. When the load resistor is chosen, the bandwidth is used, within certain practical limits, in determining the emitter dc current. A figure of merit is found for the cascaded amplifier which differs from the figure of merit used for power-amplifiers. A simple design theory for the power amplifier case is also given, and a cascade of transformer-coupled amplifier stages is compared with a cascade of resistance-coupled amplifier stages.

The treatment is also extended to common-emitter amplifiers of the type having a feedback resistor connected between the emitter and ground. It is shown that a cascade of amplifier stages of this type has the same optimum load resistor for each stage irrespective of the value of the feedback resistor. In the cases where the optimum load resistor is used and the emitter dc current is given, the feedback resistor R_e determines the bandwidth. This amplifier can be compensated to have a maximally flat frequency response by shunting the feedback resistor R_e with a capacitor C_e. An amplifier stage of this type is shown in Fig. 1. This maximally flat response characteristic has approximately a 66 per cent greater 3 db cutoff frequency than the uncompensated response characteristic.

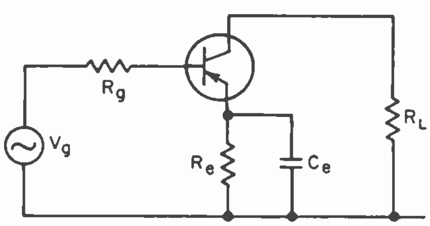

Fig. 1—Common-emitter amplifier with emitter-feedback resistor R_e and compensation capacitor C_e.

The value of the capacitor C_e can be found from (43), (43a), and (43b). The angular frequency $\omega_1 = 1/R_eC_e$ will have its upper limit 2.42 ω_a. For the examples in the text, ω_1 was approximately $2\omega_a$. Here ω_a is the cutoff frequency of the amplifier when $C_e = 0$.

A form of inductance compensation which gives a frequency response similar to the capacitance-compensated amplifier is briefly described. In this case

$$\omega_1 = \frac{R_s}{L_1} = \frac{r_{bb'}}{L_2}$$

where R_s, L_1 and L_2 are shown in Fig. 2, and ω_1 can be found from the results given in the section dealing with Inductance Compensation.

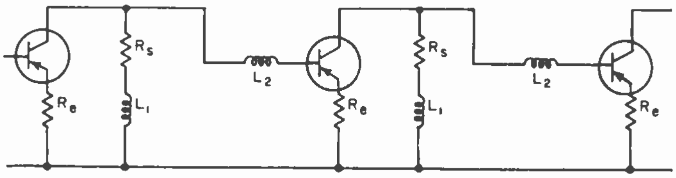

Fig. 2—Amplifier with cascaded stages—each stage given a maximally flat characteristic by coils L_1 and L_2.

The Equivalent Circuit Diagram

In Fig. 3, opposite, is shown the schematic diagram of the Johnson-Giacoletto hybrid-pi equivalent circuit for the common-emitter configuration used in this paper. The assumptions made to obtain this simplified equivalent circuit are the following:

1) The base lead resistance $r_{bb'}$ and the collector capacitance C_c are assumed to be lumped circuit parameters. These are assumptions which are fair for our alloyed type transistor and poor for a grown type transistor.
2) The only factor which makes the low-frequency value of the grounded base current gain α_0 differ from unity are assumed to be the recombination of minority carriers taking place in the base. The emitter efficiency term in the current gain and the collector multiplication factor are assumed to be unity at all frequencies.
3) The low-frequency values of the short circuit admittances Y_{22e} and Y_{12e} for the transistor with common emitter are assumed to be zero. These assumptions are generally valid when the equivalent circuit is applied in video amplifier work.

A brief discussion which leads up to the diagram in Fig. 3 is given in the Appendix.

In Fig. 3 are:

$r_{bb'}$ = base lead resistance.

α_0 = low-frequency value of the grounded base current gain.

$a_1 = 1/(1-\alpha_0)$, which is approximately the low frequency value of the grounded emitter current gain.

ω_α = alpha cutoff frequency (the frequency where the absolute value of α is 3 db below α_0).

g_{ee} = low-frequency value of the admittance y_{11} due to minority carrier diffusion in the transistor-grounded-base configuration (a more complete explanation is in Appendix.) The quantity g_{ee} is proportional to the emitter dc current I_e, and at $I_e = 1$ ma, g_{ee} equals 1/25.3 mho at 20c and 1/28 mho at 50c.

The quantities α_0, ω_α, and $r_{bb'}$ can be regarded as constant for our considerations. The capacitance C_c is inversely proportional to the square root of the col-

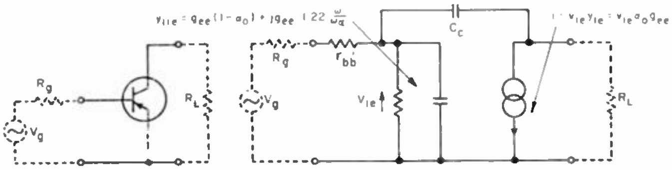

Fig. 3—Actual and equivalent circuit of an alloy junction transistor in the grounded emitter configuration.

lector dc voltage, V_c;[2] this can sometimes be an important factor in the choice of collector voltage.

Amplifiers without Emitter Resistance Feedback

Introduction

Fig. 4 contains an equivalent circuit of an amplifier stage in the case $R_e=0$.[3] In this diagram C_c has been omitted because the main effect of this capacity can be taken into account upon evaluating the capacity C_{a0} .This effect is analogous with the Miller effect in vacuum-tube amplifiers. We have

$$C_{a0} = \frac{g_{ee}1.22}{\omega_\alpha} + (\alpha_0 g_{ee} R_L + 1)C_c$$
$$C_{a0} \approx g_{ee}\left(\frac{1.22}{\omega_\alpha} + R_L C_c\right) = \frac{g_{ee}F}{\omega_\alpha} \tag{1}$$

where

$$F = 1.22 + R_L C_c \omega_\alpha. \tag{2}$$

To make the equivalent circuit a better approximation, a capacitance approximately equal to C_c should be added in parallel to R_L, but in most applications, except for some output amplifier stages where R_L may be very high, this capacitance can be neglected.

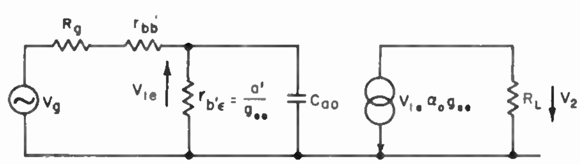

Fig. 4—Equivalent circuit for amplifier without feedback. C_{a0} = Diffusion capacitance + Miller capacitance where diffusion capacitance $= g_{ee}\ 1.22/\omega_\alpha$ and Miller capacitance $\approx g_{ee}R_L C_c$.

The frequency response of the circuit in Fig. 4 is the same as that of an RC loaded vacuum-tube amplifier stage with a dropoff at high frequencies of 6 db per octave. The 3 db cutoff frequency of the amplifier is

$$\omega_{a0} = \frac{1}{R_{a0}C_{a0}} \tag{3}$$

[3] All transistors have a finite amount of inherent emitter lead resistance. For the transistors discussed this is of the order of 1–3 ohms and will give appreciable feedback at high emitter currents. For the high emitter current case the performance will have to be evaluated by the feedback theory given below.

where

$$R_{a0} = \frac{\frac{a'}{g_{ee}}\cdot(R_g + r_{bb'})}{\frac{a'}{g_{ee}} + R_g + r_{bb'}}. \tag{4}$$

The voltage amplification at low frequencies is

$$\frac{v_2}{v_g} = \alpha_0 g_{ee} R_L \frac{\frac{a'}{g_{ee}}}{R_g + r_{bb'} + \frac{a'}{g_{ee}}}. \tag{5}$$

At high values of α_0 this reduces to

$$\frac{v_2}{v_g} = g_{ee}R_L. \tag{6}$$

Cascaded Amplifier of Identical Stages

In Fig. 5 is shown an amplifier of cascaded identical stages. By means of the equivalent circuit of Fig. 4 and (1) to (4) one can find the following expressions for gain and bandwidth. In this circuit it is assumed that the impedance loading each stage is constant with frequency. Thus the Miller-effect term in the factor F [see (2)] of this stage will be constant with frequency.

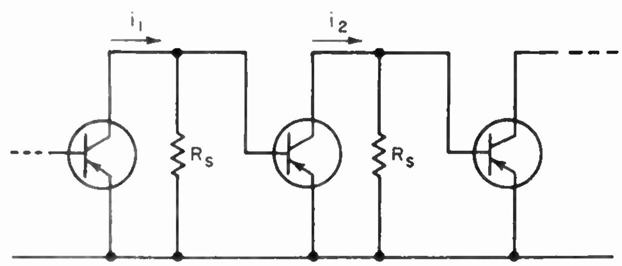

Fig. 5—Cascaded amplifier of identical stages without feedback.

This is not exactly correct because the input capacity of the following transistor is part of the load, but since the Miller term is in most practical cases (including the one which gives optimum gain) a second-order effect the approximation can be tolerated. When α_0 is close to unity, (a' is very large) so that a'/g_{ee} is very large compared with $(R_s+r_{bb'})$ the stage current gain becomes

$$\frac{i_2}{i_1} = R_s g_{ee} \tag{7}$$

and the cutoff frequency for one stage is

$$\omega_{a0} = \frac{\omega_\alpha}{(R_s + r_{bb'})F g_{ee}}. \tag{8}$$

If ω_{a0} is given, then

$$g_{ee} = \frac{\omega_\alpha}{\omega_{a0}(R_s + r_{bb'})F} \tag{9}$$

and ω_{a0} can be altered by varying the emitter current since

$$I_e = g_{ee}25.3 = 25.3\frac{\omega_\alpha}{\omega_{a0}(R_s + r_{bb'})F}. \qquad (10)$$

If ω_{a0} is given, the current amplification is

$$\frac{i_2}{i_1} = \frac{R_s\omega_\alpha}{\omega_{a0}(R_s + r_{bb'})(1.22 + R_sC_c\omega_\alpha)} \qquad (11)$$

which will be a maximum when

$$R_s = R_d = \sqrt{r_{bb'}\frac{1.22}{C_c\omega_\alpha}} \qquad (12)$$

and the amplification is, then,

$$\frac{i_2}{i_1} = \frac{\omega_\alpha}{\omega_{a0}\cdot 1.22\left(1 + \sqrt{\frac{r_{bb'}C_c\omega_\alpha}{1.22}}\right)^2}. \qquad (13)$$

The gain-bandwidth product is a constant; for a given transistor

$$\frac{i_2}{i_1}\omega_{a0} = \frac{\omega_\alpha R_d}{(R_d + r_{bb'})F} = \frac{\omega_\alpha}{1.22\left(1+\sqrt{\frac{r_{bb'}C_c\omega_\alpha}{1.22}}\right)^2}. \qquad (14)$$

This figure of merit is different from the figure of merit obtained when maximum power gain is evaluated. The maximum power gain case is discussed later.

To illustrate the use of the formulas, two examples follow.

Example 1: Transistor A has $f_\alpha = 15$ mc, $C_c = 14$ $\mu\mu f$, and $r_{bb'} = 50$ ohms. Optimum load resistance $R_d = \sqrt{50\cdot 1.22\cdot 758} = 215$ ohms.

$$F = 1.22 + \frac{215}{758} = 1.22 + 0.28 = 1.50.$$

(This illustrates a case where the Miller term part of F is less than the diffusion capacitance part.)

The gain for optimum load is

$$\frac{i_2}{i_1} = \frac{8.1}{f_{a0}}$$

where f_{a0} is in mc.

From (10) it follows that $f_{a0} = 0.95/I_e$, so emitter currents from 10 to 0.2 ma will give cutoff frequencies between 95 kc and 4.7 mc.

Example 2: Transistor B has $f_\alpha = 1$ mc, $C_c = 30\mu\mu f$, and $r_{bb'} = 200$ ohms.

$$R_d = 1140 \text{ ohms.}$$

$$F = 1.22 + 0.22 = 1.44.$$

$$\frac{i_2}{i_1} = \frac{0.6}{f_{a0}}$$

$$f_{a0} = \frac{0.013}{I_e}.$$

Emitter currents between 10 and 0.2 ma will give cutoff frequencies between 1.3 and 65 kc. As the collector-base capacitance C_c has an impedance of 4 megohms at 1.3 kc, and since this impedance is of the same order of magnitude as normal values of $1/y_{12e}$ (the minority carrier reverse transfer admittance term), the equivalent circuit and gain formulas will not be valid for the lower range of cutoff frequencies.

The above formulas were for the case where α_0 is very nearly unity; for the case where we cannot assume that α_0 is unity, we have the following

$$\frac{i_2}{i_1} = \frac{R_sg_{ee}\alpha_0}{1 + (R_s + r_{bb'})\frac{g_{ee}}{a'}} \qquad (15)$$

$$\omega_{a0} = \frac{\omega_\alpha\left[1 + (R_s + r_{bb'})\frac{g_{ee}}{a'}\right]}{(R_s + r_{bb'})Fg_{ee}} \qquad (16)$$

$$F = 1.22 + \left[R_s\|\left(r_{bb'} + \frac{a'}{g_{ee}}\right)\right]C_c\omega_\alpha. \qquad (17)$$

If the bandwidth is given, the current gain is

$$\frac{i_2}{i_1} = \frac{R_s\omega_\alpha\alpha_0}{\omega_{a0}(R_s + r_{bb'})F}. \qquad (18)$$

From (17) and (18) it will be seen that the gain-bandwidth product is only slightly modified by a'.

In a limiting case when a'/g_{ee} is small in comparison with $(R_s + r_{bb'})$ and $1/\omega_\alpha C_c$ we will have the following

$$\frac{i_2}{i_1} = \frac{R_s}{R_s + r_{bb'}}\,\frac{\alpha_0}{(1 - \alpha_0)} \qquad (19)$$

$$\omega_{a0} = \frac{\omega_\alpha}{1.22\,a'}. \qquad (20)$$

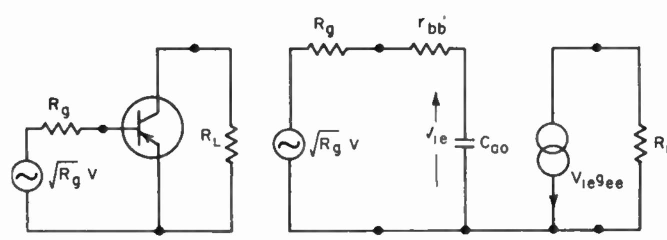

Fig. 6—Actual and equivalent circuits for a common-emitter power amplifier α_0 assumed equal to one.

Maximum Power Gain

In Fig. 6 is shown the equivalent circuit diagram for a one-stage common-emitter video power amplifier. An ideal transformer is inserted between the generator (which has an emf of one volt and an internal impedance of one ohm) and the transistor, giving a generator resistance R_g and an emf of $\sqrt{R_g}$ volts.

The low-frequency output voltage and the 3 db cut-off frequency are

$$V_2 = g_{ee}R_L\sqrt{R_g} \tag{21}$$

$$\omega_{a0} = \frac{\omega_\alpha}{(R_g + r_{bb'})g_{ee}F}. \tag{22}$$

The low-frequency output power

$$P_2 = \left(\frac{\omega_\alpha}{\omega_{a0}}\right)^2 \left(\frac{\sqrt{R_gR_L}}{(R_g + r_{bb'})(1.22 + R_LC_c\omega_\alpha)}\right)^2 \tag{23}$$

has a maximum when

$$R_g = r_{bb'} \tag{24}$$

and

$$R_L = \frac{1.22}{C_c\omega_\alpha}. \tag{25}$$

For the maximum case a figure of merit of available power gain times the square of the bandwidth

$$Vf^2_{a0} = \frac{f_\alpha}{30.6r_{bb'}C_c} \tag{26}$$

is essentially the same as those given elsewhere.[2,4]

If an amplifier is made by cascading identical power-amplifier stages with ideal transformers between the stages as shown in Fig. 7, the optimum low-frequency current gain per stage is

$$\frac{i_2}{i_1} = \frac{1}{4.41\omega_{a0}}\sqrt{\frac{\omega_\alpha}{r_{bb'}C_c}}. \tag{27}$$

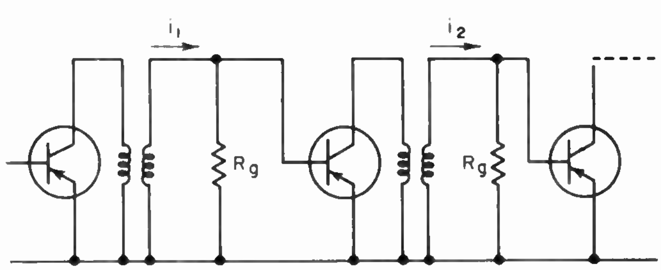

Fig. 7—Cascaded amplifier of identical stages matched for a maximum gain-bandwidth product (ideal transformers assumed).

The ratio between the current gain per stage for the case having ideal transformers and the current gain per stage for the transformerless case, (13), is

$$0.277\sqrt{\frac{1}{r_{bb'}C_c\omega_\alpha}}\left(1 + \sqrt{\frac{r_{bb'}C_c\omega_\alpha}{1.22}}\right)^2. \tag{28}$$

[4] J. M. Early, "*P-N-I-P* and *n-p-i-n* junction transistor triodes," *Bell Sys. Tech. J.* vol. 33, pp. 517–533; May, 1954.

For transistor A, mentioned earlier, this ratio is 1.64 or 4.3 db, and for transistor B the ratio is 2 or 6 db. From this we can see that, for reasons of economy, transformers should normally be considered desirable only in the cases where the required bandwidth becomes so wide that the gain per stage of the transformerless case becomes very low.

Amplifiers with Feedback

Introduction

Consider first the case of an amplifier with a feedback resistor R_e, but no shunting capacitor C_e. This kind of amplifier has a response curve of the same form as the amplifier without feedback. When the emitter current I_e (and thereby g_{ee}) is given, the cutoff frequency is about $(1+g_{ee}R_e)$ times that of the amplifier stage without feedback and the amplification is proportionally smaller, giving about the same gain-bandwidth product.

The main advantages of amplifiers with feedback are better linearity and the fact that the capacitor C_e can be added to give a flat frequency response, as described in section, "*Compensation to obtain maximally Flat Frequency Response.*" In addition, with feedback a higher emitter current can be used for a given design cutoff frequency, and in this way we can obtain a larger dynamic range for the amplifier. The following equations for cutoff frequencies and amplifications can be obtained from the formulas for the amplifier without feedback by introducing $g_{ee}/(1+R_eg_{ee})$ instead of g_{ee}. The value of F is still $1.22+R_LC_c\omega_\alpha$, but in all other places ω_α/η is introduced instead of ω_α, where η is defined as

$$\eta = 1 + \frac{R_e}{(R_g + r_{bb'})F}. \tag{29}$$

The basis for this transformation is the equivalent circuit given in Appendix. The cutoff frequency is now

$$\omega_a = \frac{\omega_\alpha\left[1 + (R_g + r_{bb'})\dfrac{g_{ee}}{a'(1 + g_{ee}R_e)}\right](1 + g_{ee}R_e)}{(R_g + r_{bb'})\eta F g_{ee}} \tag{30}$$

and for a' very large we get

$$\omega_a = \frac{\omega_\alpha(1 + g_{ee}R_e)}{(R_g + r_{bb'})\eta F g_{ee}}. \tag{31}$$

The voltage amplification of a single stage is

$$\frac{V_2}{V_1} = \alpha_0 g_{ee}R_L \frac{\dfrac{a'}{g_{ee}}}{R_g + r_{bb'} + \dfrac{a'}{g_{ee}}(1 + R_eg_{ee})}$$

$$\approx \frac{g_{ee}R_L}{1 + R_eg_{ee}}. \tag{32}$$

This differs from (5) and (6) only by the feedback factor in the denominator.

Cascaded Amplifier of Identical Stages

In a cascaded amplifier having identical stages as in Fig. 5, but with emitter resistors R_e, the 3 db cutoff frequency when a' is very large is

$$\omega_a = \frac{\omega_\alpha(1 + R_e g_{ee})}{(R_s + r_{bb'})\eta F g_{ee}}. \tag{33}$$

For example, using this formula one can decide on the value of R_e after having fixed R_s and the emitter current (and thereby g_{ee}). The amplification for one stage is

$$\frac{i_2}{i_1} = \frac{R_s g_{ee}}{1 + R_e g_{ee}}. \tag{34}$$

When ω_a is given,

$$\frac{i_2}{i_1} = \frac{R_s \omega_\alpha}{\omega_a (R_s + r_{bb'})\eta(1.22 + R_s C_c \omega_\alpha)}. \tag{35}$$

To find the optimum value of R_s it is now assumed that η is independent of R_s. Actually η is a function of F [see (29)], which is in turn a function of R_L (which equals R_s), but in most practical applications η will vary only slightly with R_s. Under this assumption the optimum load resistance becomes

$$R_s = R_d = \sqrt{r_{bb'} \cdot \frac{1.22}{C_c \omega_\alpha}}. \tag{36}$$

This is the same expression as that obtained in (12) for the simple amplifier. With $R_s = R_d$ the stage amplification is

$$\frac{i_2}{i_1} = \frac{\omega_\alpha}{\omega_a \eta \cdot 1.22\left(1 + \sqrt{\frac{r_{bb'} C_c \omega_\alpha}{1.22}}\right)^2}. \tag{37}$$

Thus, the gain-bandwidth product for this case is

$$\frac{i_2}{i_1}\omega_a = \frac{\omega_\alpha}{\eta \cdot 1.22\left(1 + \sqrt{\frac{r_{bb'} C_c \omega_\alpha}{1.22}}\right)^2} \tag{38}$$

which is the same as the gain bandwidth-product of (14) except for the factor η which normally is a little larger than one.

For the cases where a' is small, the following are more exact expressions

$$\omega_a = \frac{\omega_\alpha(1 + R_e g_{ee})\left(1 + (R_s + r_{bb'})\frac{g_{ee}}{(1 + R_e g_{ee})a'}\right)}{(R_s + r_{bb'})\eta F g_{ee}} \tag{39}$$

$$F = 1.22 + \left\{R_s \middle\| \left[r_{bb'} + \frac{a'(1 + R_e g_{ee})}{g_{ee}}\right]\right\} C_c \omega_\alpha \tag{40}$$

$$\frac{i_2}{i_1} = \frac{R_s \omega_\alpha \alpha_0}{\omega_a (r_{bb'} + R_s)\eta F}. \tag{41}$$

Maximum Power Gain

For the case where the transistor amplifiers with emitter-feedback are coupled for maximum power gain formulas result which are similar to (21) through (28). Multiplication of (23) by $1/\eta^2$ yields the expression for the output power.

The ratio between the optimum stage gain for a cascaded amplifier with ideal transformers and the stage gain for the transformerless amplifier is the same as that given by (28) multiplied by η_1/η_2 where η_1 is the value of η in the transformerless case and η_2 is for the power amplifier case.

Compensation to Obtain Maximally Flat Frequency Response

Consider now the case of a capacitor C_e shunting the feedback resistor R_e, as illustrated in Fig. 1. In this case the low-frequency amplification is not affected by the capacitor so that (32) will again serve to calculate the amplification. The value of C_e which makes the frequency response maximally flat can be found from the following formulas:

$$C_e = \frac{1}{R_e \omega_1} \tag{42}$$

where

$$\omega_1 = \frac{\eta \omega_a (1 - q^2)}{-1 + q\eta + \sqrt{\eta^2 + 1 - 2q\eta}} \tag{43}$$

where η and ω_a are given in (29) and (30) and

$$q = \frac{R_g + r_{bb'} + \frac{a'}{g_{ee}}}{R_g + r_{bb'} + \frac{a'}{g_{ee}}(1 + R_e g_{ee})}. \tag{44}$$

When $\eta \approx 1$ and $q \ll 1$ which normally will be the case, (43) reduces to

$$\omega_1 \approx 2.42\omega_a \frac{1}{1 + 0.7(\eta - 1 + q)} \tag{43a}$$

or to a further approximation

$$\omega_1 \approx 2.42\omega_a. \tag{43b}$$

As a rule ω_1 will generally be a little lower than the value given by (43b). For the case of the maximally flat characteristic the normalized amplitude response is

$$|B_0| = \sqrt{\frac{1 + \frac{\omega^2}{\omega_1^2}}{1 + \frac{\omega^2}{\omega_1^2} + \frac{\omega^4}{\eta^2 \omega_a^2 \omega_1^2}}}. \tag{45}$$

As mentioned previously some transistors have an appreciable resistance $r_{ee'}$ in series with the emitter lead that is inherent to the device. This resistance means that some feedback is present even if the external R_e is zero. To find the amplification in this situation the expressions in Part D1–D3 can be used and the same gain-bandwidth product will be obtained. For the case where the resistance R_e is introduced we can find the low frequency amplification and ω_a by including $r_{ee'}$ in R_e. If the compensating capacitor C_e is used, it shunts only the external feedback resistor. It is therefore necessary to modify (44) to

$$q = \frac{R_g + r_{bb'} + \dfrac{a'(1 + r_{ee'}g_{ee})}{g_{ee}}}{R_g + r_{bb'} + \dfrac{a'(1 + R_{e'}g_{ee})}{g_{ee}}} \tag{46a}$$

where

$$R_{e'} = R_e + r_{ee'}. \tag{46b}$$

Measurements

Introduction

The design of a maximally flat stage with a feedback resistor and capacitor as described above was tried out on several p-n-p alloy transistors.[1] The parameters of the transistors were determined in part by measuring the performance of an amplifier stage without the feedback resistor. With feedback resistors the frequency response was measured and the frequencies f_a compared with calculated figures. For the maximally flat case the response was compared with the response calculated from the transistor parameters. A typical set of measurements and calculations is given in the following paragraph. The agreement between theory and measurements in this case is similar to what is found with other transistors of the same type.

Determination of Parameters

Except for $r_{bb'}$ which was already known from other measurements, the transistor parameters were measured by means of the circuit shown in Fig. 8.

The principal measuring instruments were a signal generator and a sensitive vacuum tube voltmeter. Three measurements were made:

1) The low frequency resistance of the transistor between base and ground was determined by measuring the voltage rise at the base pin of the socket, when the transistor was pulled out of the socket.
2) The voltage amplification at low frequencies was measured.
3) The 3 db frequency cutoff frequency was measured both with the load resistance to be used in the amplifier—here 330 ohms—and with this resistance shunted down to 33 ohms.

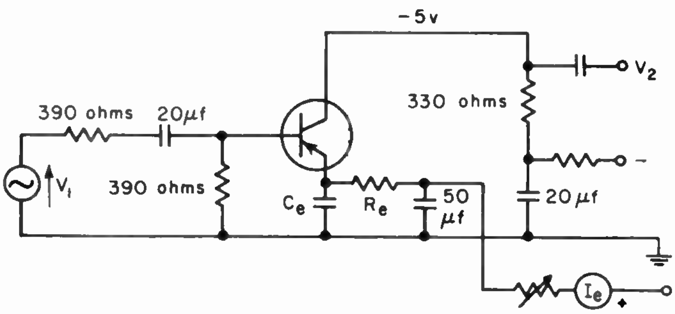

Fig. 8—Test circuit used for obtaining measurements.

The measurements were carried out at $I_e = 10$ ma and $I_e = 2$ ma with the collector voltage set at 5v. At 10 ma the collector power is 20 mw and the temperature is assumed to be 50c so $g_{ee} = 10/27$ mho. At 2 ma the temperature is assumed to be 25c and $g_{ee} = 2/25.7$ mho. The first set of measurements can be used to calculate $[a'(1+g_{ee}r_{ee'})]/g_{ee}$ and using this resistance and the amplification at low frequencies we can find $(1+g_{ee}r_{ee'})$ and $r_{ee'}$. From the measurements of the cutoff frequency with the normal load resistance we can find ω_a/Fg_{ee} and F can be found from this measurement and the measurement with low load resistance (assuming that F approaches 1.22 at low values of R_L).

Measurements

For Emitter Current $I_e = 2$ ma, $g_{ee} = 0.078$ mho:

$$\frac{a'(1+r_{ee'}g_{ee})}{g_{ee}} = 590 \text{ ohms}$$

$1 + r_{ee'}g_{ee} = 1.1$

$a' = 42$

$r_{ee'} = 1.3$ ohm. (This resistance cannot be determined very accurately at a low emitter current.)

$f_{a0} = 0.62$ mc at $R_L = 330$ ohms

$f_{a0} = 0.90$ mc at $R_L = 33$ ohms

$$\frac{Fg_{ee}}{\omega_\alpha} = 1520\ \mu\mu\text{f},\ F \approx 1.8.$$

For the case $R_e = 62$ ohms:

$1 + R_e g_{ee} = 5.9$

$\eta = 1.14$

$f_a = 2.3$ mc calculated and

$f_a = 2.2$ mc measured

$q = 0.247$, which together with the calculated values for f_a and η gives

$f_1 = 4.1$ mc and $C_e = 630\ \mu\mu f$.

Fig. 9 shows the measured responses for $R_e = 62$ ohms with $C_e = 0$, 620 $\mu\mu f$ and 820 $\mu\mu f$. For comparison the calculated response curves for the uncompensated and for the maximally flat case are shown.

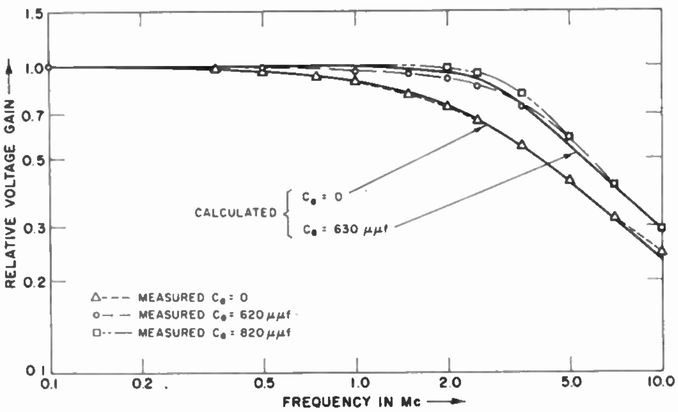

Fig. 9—Relative voltage gain of a video amplifier for different values for the compensation capacitor C_e; $I_e=2$ ma, $R_e=62$ ohms.

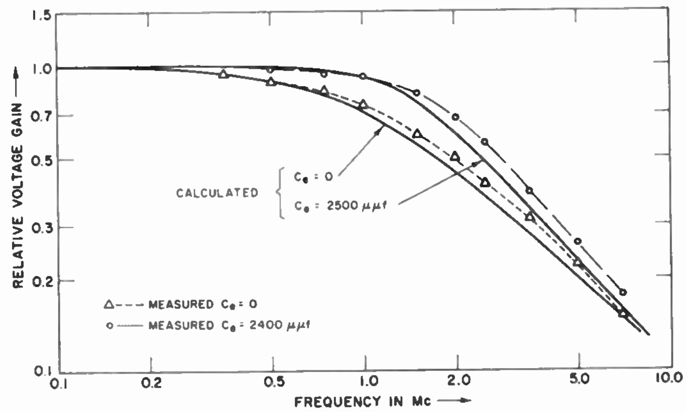

Fig. 10—Relative voltage gain of a video amplifier for different values for the compensation capacitor C_e; $I_e=10$ ma, $R_e=31.5$ ohms.

The voltage amplification V_2/V_1 (see Fig. 8) was calculated as 2.01 and the measured value was 2.10.

For Emitter Current $I_e=10$ *ma*, $g_{ee}=0.37$ *mho:*

$$\frac{a'(1+r_{ee'}g_{ee})}{g_{ee}}=168 \text{ ohms}$$

$$1+r_{ee'}g_{ee}=1.42,\ a'=44$$

$$r_{ee'}=1.1 \text{ ohm}$$

$$f_{a0}=0.28 \text{ mc at } R_L=330 \text{ ohms}$$

$$f_{a0}=0.51 \text{ mc at } R_L=33 \text{ ohms}$$

$$\frac{Fg_{ee}}{\omega_\alpha}=8000\ \mu\mu\text{f},\ F\approx 2.4.$$

Data for $R_e=31.5$, 62 and 85 ohms are shown below in Table I.

TABLE I

R_e	31.5 ohms	62 ohms	85 ohms
$1+R_eg_{ee}$	12	24.3	33
η	1.05	1.10	1.14
f_a calculated	1.01 mc	1.87 mc	2.4 mc
f_a measured	1.13 mc	1.80 mc	2.2 mc
q	0.203	0.135	0.102
f_1	2.03	3.8 mc	4.8 mc
$C_e\mu\mu$f	2500 $\mu\mu$f	670 $\mu\mu$f	390 $\mu\mu$f
V_2/V_1 calculated	4.47	2.30	1.74
V_2/V_1 measured	4.13	2.37	1.80

Fig. 10 contains the measured responses for $R_e=31.5$ ohms and $C_e=0$ and 2400 $\mu\mu f$. Calculated curves are shown for the uncompensated and for the maximally-flat case. Similar curves are shown in Fig. 11 for $R_e=62$ ohms and in Fig. 12 for $R_e=85$ ohms.

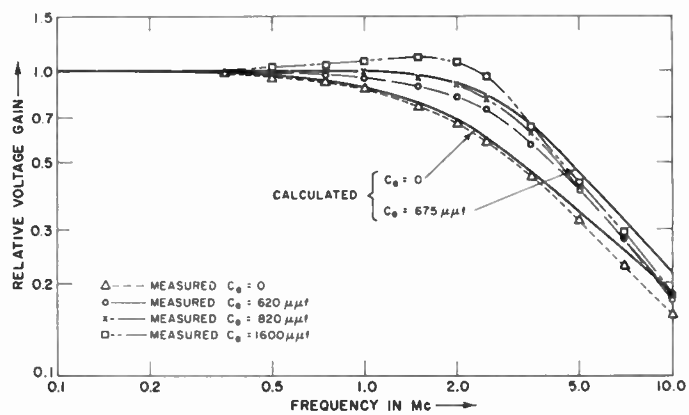

Fig. 11—Relative voltage gain of a video amplifier for different values for the compensation capacitor C_e; $I_e=10$ ma, $R_e=62$ ohms.

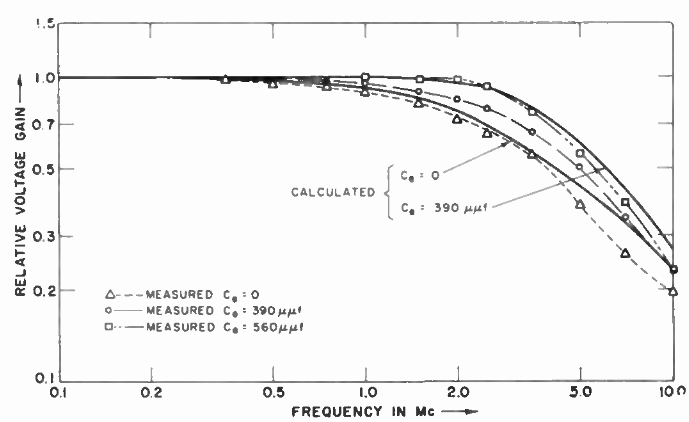

Fig. 12—Relative voltage gain of a video amplifier for different values for the compensation capacitor C_e; $I_e=10$ ma, $R_e=85$ ohms.

Inductance Compensation

In Fig. 2 is shown a cascaded amplifier where coils are used for compensation.[5] Since two coils are used in each stage there are more degrees of freedom and it will be possible to have a greater variety of frequency responses. If, on the other hand, the cases are restricted to those where

$$\frac{R_s}{L_1}=\frac{r_{bb'}}{L_2} \tag{47}$$

[5] An inductance-compensated video amplifier using surface-barrier transistor is described in the article: J. B. Angell and F. P. Keiper Jr., "Circuit application of surface barrier transistors," Proc. IRE, vol. 41, pp. 1709–1712; December, 1953.

the characteristics will be about the same as those for the capacitor-compensated case. Thus the frequency response will be maximally flat when

$$L_1 = \frac{R_s}{\omega_1} \quad \text{and} \quad L_2 = \frac{r_{bb'}}{\omega_1}. \tag{48), (49}$$

and

$$q = \frac{R_s + r_{bb'}}{R_s + r_{bb'} + \frac{a'}{g_{ee}}(1 + R_e g_{ec})} \tag{50}$$

The quantity ω_1 can then be found from (43), (43a), or (43b) and the frequency response from (45). This response will not be exactly the same as that obtained for the capacitor-compensated case because of the differences between the equations for q [see (44) and (50)].

It should be mentioned that inductance compensation of the kind described here can only be used in designing interstages which are to be driven from a constant current source. If the feedback capacitor of the driving stage is not negligible its effect is taken into account by lumping the Miller capacitance with the diffusion capacitance as described above.

Stability

A transistor of the type considered in this article is potentially unstable in the common emitter configuration. Internal feedback over the collector-base capacity C_c can produce oscillations under certain load conditions.

In video amplifiers the stability problem is only present in case of inductance compensation when the base and collector are loaded by impedances with high positive reactances. In experimental work on amplifiers of this kind, oscillations may be expected during the alignment procedure if the coils are adjusted to unnormally high inductances.

In the case of a capacity in the collector circuit being compensated by an inductance in series with the load resistance to obtain a maximally flat frequency response, the total load impedance will have a negative reactance at all frequencies, and oscillations cannot be expected. This also applies to the kind of interstage inductance compensation for a maximally flat response described in this article.

In stability considerations a reactance $j\omega L$ in series with the load resistance R_L can be taken into account as we did with the Miller capacitance, and C_{a0} in Fig. 4 will be shunted by a negative conductance $\alpha_0 g_{ee}\omega^2 LC_c \approx g_{ee}\omega^2 LC_c$. In a high alpha transistor the resistive part of the input impedance will be approximately

$$r_{bb'} - \frac{1}{\omega^2 LC_c g_{ee} + \dfrac{C_{a0}^2}{LC_c g_{ee}}}$$

with the minimum value

$$r_{bb'} - \frac{LC_c g_{ee}}{C_{a0}^2}.$$

If this resistance is positive, the transistor will not oscillate at any source impedance.

Similar considerations can be made for an amplifier stage with a feedback resistor R_e.

Appendix

Equivalent Circuit Diagram for the Common-Emitter Configuration

Fig. 13 shows the notation for a grounded-base transistor and its equivalent circuit. The three-terminal network in the rectangle contains the intrinsic part of the transistor having the short circuit admittances y_{11} y_{12}, y_{21}, and y_{22}. The quantity C_c is the collector-to-base barrier capacitance, and r_{bb}' is the base lead resistance.

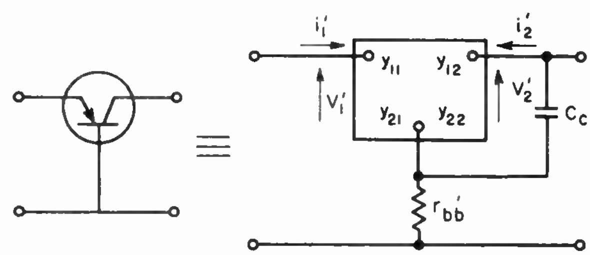

Fig. 13—Equivalent circuit of a common base transistor. The network in the rectangle represents the effect of minority carrier diffusion in the transistor.

In a transistor of the type considered here the emitter efficiency can be assumed to equal unity and for the y's we can then have approximately:[4,6,7]

$$y_{11} = g_{ee}\phi \coth \phi \tag{51}$$

$$y_{21} = -\alpha_0 g_{ee}\phi \frac{1}{\sinh \phi} \tag{52}$$

$$y_{12} = -\alpha_0 g_{cc}\phi \frac{1}{\sinh \phi} \tag{53}$$

$$y_{22} = g_{cc}\phi \coth \phi. \tag{54}$$

If $\alpha_0 > 0.9$ a good approximation will be

$$\phi = \sqrt{2.43 j \frac{\omega}{\omega_\alpha}}. \tag{55}$$

We have then

$$\alpha = \frac{y_{21}}{y_{11}} = \alpha_0 \frac{1}{\cosh \phi} \tag{56}$$

[6] H. Johnson, "Diffusion Reactances of Junction Transistors," IRE-AIEE Transistor Research Conference, Pennsylvania State College; July, 1953.

[7] The ϕ used here is the same as s used by Early, *loc. cit.*

which is 3 db down when $\omega=\omega_\alpha$. Series expansion of the amplitude and phase characteristics results in the following approximations:

$$y_{11} = g_{ee}\frac{1 + j0.97\dfrac{\omega}{\omega_\alpha}}{1 + j0.16\dfrac{\omega}{\omega_\alpha}} \approx g_{ee}\left(1 + j\frac{\omega}{\omega_\alpha}\right) \qquad (57)$$

$$y_{21} = -\alpha_0 g_{ee}\frac{e^{-j0.15(\omega/\omega_\alpha)}}{\left(1 + j0.256\dfrac{\omega}{\omega_\alpha}\right)}. \qquad (58)$$

As mentioned earlier, we can make the further approximation that $y_{12}\approx 0$ and $y_{22}\approx 0$. Converting the equivalent circuit to grounded emitter configuration produces the equivalent circuit shown in Fig. 14,

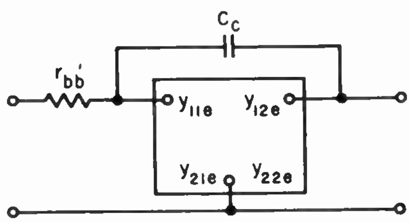

Fig. 14—Equivalent circuit of a common-emitter transistor, y_{11e}, y_{12e}, y_{21e}, and y_{22e} are the "minority carrier" admittances of the transistor.

where

$$y_{12e} \approx 0 \qquad (59)$$

and

$$y_{22e} \approx 0. \qquad (60)$$

Furthermore,

$$y_{11e} \approx y_{11} + y_{21} \approx g_{ee}(1 - \alpha_0) + g_{ee}\cdot 1.22j\frac{\omega}{\omega_\alpha} =$$

$$\frac{g_{ee}}{a'} + j1.22\frac{\omega}{\omega_\alpha}g_{ee} \qquad (61)$$

$$y_{21e} \approx -y_{21} \approx \alpha_0 g_{ee}\frac{e^{-j0.15(\omega/\omega_\alpha)}}{1 + j0.256\dfrac{\omega}{\omega_\alpha}} \approx \alpha_0 g_{ee}. \qquad (62)$$

The last approximation is permissible because the frequencies involved here are lower than ω_α.

The equivalent circuit can now be reduced as shown in Fig. 3.

Equivalent Circuit for a Common Emitter Amplifier with a Feedback Impedance in the Emitter Lead

An equivalent circuit diagram for a common emitter transistor with an impedance Z_e in series with the emitter lead is shown in Fig. 15.

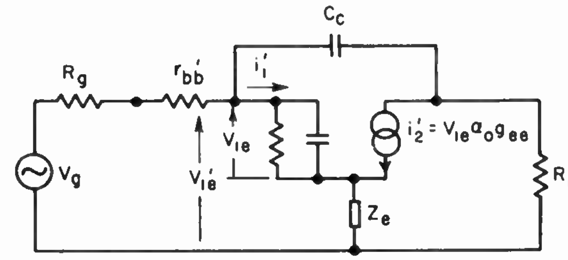

Fig. 15—Equivalent circuit of a common-emitter transistor amplifier with a feedback impedance Z_e.

In this case

$$V_{1e}' = V_{1e} + (i_1' + i_2')Z_e = i_1'\frac{1 + Z_e(y_{11e} + y_{21e})}{y_{11e}} \qquad (63)$$

$$i_2' = y_{21e}V_{1e} = \frac{y_{21e}V_{1e}'}{1 + Z_e(y_{11e} + y_{21e})}. \qquad (64)$$

and this gives the equivalent circuit in Fig. 16,

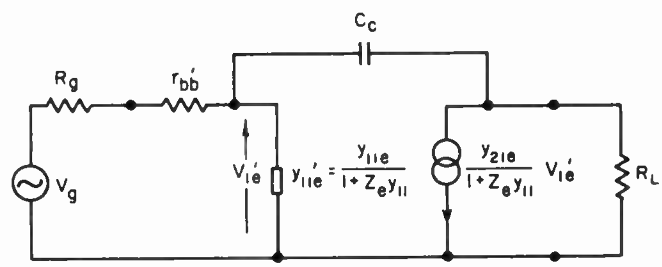

Fig. 16—Simplified equivalent circuit of a common-emitter amplifier with emitter-resistance feedback.

where

$$y'_{11e} = \frac{y_{11e}}{1 + Z_e(y_{11e} + y_{21e})} \qquad (65)$$

and

$$y'_{21e} = \frac{y_{21e}}{1 + Z_e(y_{11e} + y_{21e})}. \qquad (66)$$

From (61) and (62),

$$y_{11e} + y_{21e} \approx y_{11} \qquad (67)$$

and from (57)

$$y_{11} = g_{ee}\left(1 + j\frac{\omega}{\omega_\alpha}\right)$$

so that

$$y'_{11e} = \frac{y_{11e}}{1 + Z_e g_{ee}\left(1 + j\dfrac{\omega}{\omega_\alpha}\right)} \qquad (68)$$

$$y'_{21e} = \frac{y_{21e}}{1 + Z_e g_{ee}\left(1 + j\dfrac{\omega}{\omega_\alpha}\right)}. \qquad (69)$$

The equivalent circuit can be further reduced in the same way as the circuit of Fig. 3 was reduced to that of Fig. 4 by lumping the Miller-effect term with the diffusion capacitance. Lumping these two produces the circuit of Fig. 17,

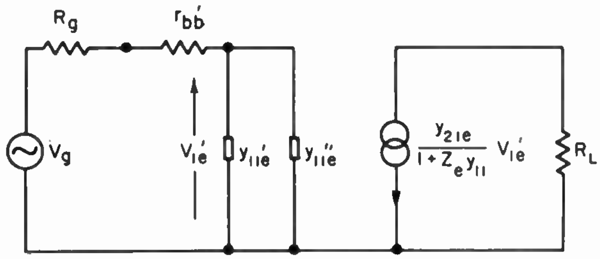

Fig. 17—Simplified circuit for a common-emitter amplifier with feedback. Admittance y_{11e}'' represents the *Miller* effect.

where

$$y''_{11e} \approx \frac{R_L y_{21e} j\,\omega C_c}{1 + Z_e g_{ee}\left(1 + j\dfrac{\omega}{\omega_\alpha}\right)}. \tag{70}$$

At this point it is seen that

$$y'_{11e} + y''_{11e} = \frac{1 + j\dfrac{\omega}{\omega_r}}{\dfrac{a'}{g_{ee}}(1 + Z_e y_{11})} \tag{71}$$

where ω_r is defined by

$$\frac{1}{\omega_r} = \frac{1.22a'}{\omega_\alpha} + a' R_L C_C = \frac{F \cdot a'}{\omega_\alpha}. \tag{72}$$

For the case where $Z_e = R_e$ the voltage amplification at low frequencies is found by means of (32).

If R_e is shunted by C_e, then

$$Z_e = \frac{R_e}{1 + j\dfrac{\omega}{\omega_1}} \tag{73}$$

where

$$\omega_1 = \frac{1}{R_e C_e} \tag{74}$$

and where, also,

$$1 + Z_e y_{11} = (1 + R_e g_{ee})\frac{1 + j\dfrac{\omega}{\omega_2}}{1 + j\dfrac{\omega}{\omega_1}} \tag{75}$$

where

$$\frac{1}{\omega_2} = \frac{1}{\omega_1(1 + R_e g_{ee})} + \frac{R_e g_{ee}}{\omega_\alpha(1 + R_e g_{ee})}. \tag{76}$$

Symbols

α_0—low frequency value of current gain for common base transistor.

Γ—available power gain, transducer gain.

$\omega_1 = 2\pi f_1 = \dfrac{1}{R_e C_e}$—reference angular frequency associated with external emitter impedance.

$\omega_\alpha = 2\pi f_\alpha$—angular alpha cutoff frequency.

$\omega_{a0} = 2\pi f_{a0}$—angular cutoff frequency of amplifier stage.

$\omega_a = 2\pi f_a$—angular cutoff frequency of amplifier stage with emitted-feedback when $C_e = 0$.

ω_r—angular frequency, defined in (72).

ϕ—a transistor parameter defined in (55).

$$\eta = 1 + \frac{R_e}{(R_g + r_{bb'})F}.$$

$a' = \dfrac{1}{1 - \alpha_0}$—approximately the current gain of a transistor in the grounded emitter configuration.

C_{a0}—diffusion capacity corrected for Miller effect, see (1).

C_c—collector-base barrier capacity.

C_e—feedback capacitor in external emitter circuit. See Fig. 1.

F—a parameter relating to the Miller effect as defined in (2).

g_{ee}—low frequency value of the short circuit admittance y_{11} of the intrinsic transistor.

I_e—emitter dc current.

L_1, L_2—compensating coils as shown in Fig. 2.

P_2—output power.

q—a parameter defined in (44) and (46) for capacitor compensation and in (50) for coil compensation.

$r_{b'e}$—a transistor parameter in the hybrid-π equivalent circuit (see Fig. 3).

$r_{bb'}$—base lead resistance in transistor.

$r_{ee'}$—emitter lead resistance in transistor.

R_{a0}—a resistance defined in (4)

$R_d = \sqrt{r_{bb'}\dfrac{1.22}{C_c\omega_\alpha}}$, an optimum load resistance [see (12)].

R_e—emitter feedback resistance.

R_g—generator resistance.

R_L—load resistance.

R_s—load resistances for cascaded stages in amplifier (see Fig. 5).

y_{11}, y_{12}, y_{21}, y_{22},—short circuit minority carrier admittances for grounded base transistor.

y_{11e}, y_{12e}, y_{21e}, y_{22e},—short circuit minority carrier admittances of grounded emitter transistor.

Z_e—feedback impedance in emitter lead.

ACKNOWLEDGMENT

The author wishes to express his gratitude to the National Academy of Sciences and The International Cooperation Administration, Washington, for the Research Fellowship which made it possible for him to spend a year at Stanford University and Stanford Research Institute. He also wishes to thank the Danish foundations, Danmark-Amerika Fondet, and The Otto Mönsteds Fond and his employers, The Royal Technical University of Denmark and The Radio Receiver Research Laboratory, for their support. Thanks are also due to Dr. Victor Grinnich and L. J. Kabell of Stanford Research Institute for their helpful suggestions and critical review of the report on which this paper is based.

During the year in Stanford the author had many interesting and inspiring discussions with J. M. Pettit who suggested Transistor Video Amplifiers as a research object and with John Linvill, Dr. R. D. Middlebrook, R. M. Scarlett and others at the Electronics Research Laboratory, Stanford University.

Hazards Due to Total Body Irradiation by Radar*

H. P. SCHWAN† AND K. LI†

Summary—**Experimental work by others at 10 cm wavelength has shown that irreversible damage to the eye is caused by electromagnetic radiation, if the energy flux is in excess of about 0.2 watt/cm². Intolerable temperature rise, due to total body irradiation may be anticipated for flux values in excess of 0.02 watts/cm². Hence a discussion of hazards due to total body irradiation is of primary interest. This paper presents data which analyze the mode of propagation of electromagnetic radiation into the human body and resultant heat development. The two quantities which are considered in detail are: 1) coefficient, which characterizes the percentage of airborne electromagnetic energy as absorbed by the body, and 2) distribution of heat sources in skin, subcutaneous fat, and deeper situated tissues. It is shown:**

1) **The percentage of absorbed energy is near 40 per cent at frequencies much smaller than 1000 and higher than 3000 mc. In the range from about 1000 to 3000 mc the coefficient of absorption may vary from 20 to 100 per cent.**
2) **Radiation of a frequency below 1000 mc will cause deep heating, not well indicated by the sensory elements in the skin and, therefore, considered especially dangerous. Radiation whose frequency exceeds 3000 mc will be absorbed in the skin. Radiation of a frequency between 1000 and 3000 mc will be absorbed in both body surface and in the deeper tissues, the ratio being dependent on parameters involved.**
3) **Arguments are advanced in support of tolerance values for total body irradiation near 0.01 watts/cm².**

Conclusions of practical value are: 1) Since sensory elements are located primarily in the skin, low-frequency radiation (f < 1000 mc) is much more dangerous than high-frequency radiation. 2) Radiation of very high frequency (f > 3000 mc) causes only superficial heating with much the same effects as infrared and sunlight. The sensory reaction of the skin should provide adequate warning.

* Original manuscript received by the IRE, December 15, 1955; revised manuscript received, June 15, 1956. These studies were aided by a contract between the Office of Naval Research, Dept. of the Navy and the Univ. of Pennsylvania, NR 119-289.

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THE PROBLEM

HEALTH HAZARDS, resulting from the exposure of mankind to strong sources of electromagnetic radiation, have been discussed by several investigators in recent years. Earlier investigations, concerned with the possible harmful effects of electromagnetic radiation, stated negative results,[1] while more recent investigations are indicative of the possible harmful effects of such radiation.[2–6] This may be due to the fact that only recently sources of sufficient power to establish a health hazard have become available. A more detailed discussion of the presently available literature has been given by us elsewhere[7] and is, therefore, omitted. The harmful effects of excessive amounts of radiation result either from a general rise in total body temperature or are limited to selective temperature rise of sensitive parts of the body, such as testicles or especially the eyes. Table I compares data pertaining to these types of damage. It has been assumed here that in the case of total body irradiation

[1] L. E. Daily, "A clinical study of the results of exposure of laboratory personnel to radar and high frequency radio," *U. S. Naval Bull.*, vol. 41, p. 1052; 1943.

[2] D. B. Williams, J. P. Monahan, W. J. Nicholson, and J. J. Aldrich, "Biologic effects studies on microwave radiation: Time and power thresholds for the production of lens opacities by 12.3-cm microwaves," IRE TRANS., PGME-4, pp. 17–22; February, 1956.

[3] T. S. Ely and D. E. Goldman, "Heat exchange characteristics of animals exposed to 10-cm microwaves," IRE TRANS., PGME-4, pp. 38–43; February, 1956.

[4] H. M. Hines and J. E. Randall, "Possible industrial hazards in the use of microwave radiation," *Elec. Engrg.*, vol. 71, p. 879; 1952.

[5] J. F. Herrick and F. H. Krusen, "Certain physiologic and pathologic effects of microwaves," *Elec. Engrg.*, vol. 72, p. 239; 1953.

[6] S. I. Brody, "Operational hazard of microwave radiation," *J. Aviation Med.*, vol. 24, p. 328; 1953.

[7] H. P. Schwan and G. M. Piersol, "The absorption of electromagnetic energy in body tissues: a review and critical analysis," *Amer. J. Phys. Med.*, part I, vol. 33, p. 371; 1954; part II, vol. 34, p. 425; 1955.

TABLE I*

	Critical temperature elevation	Estimated critical flux	Experimental evidence at 10 cm
Eye damage (Cataract)	10°C.	0.1	0.2
Testicular damage	1°C.(?)	0.01	?
Total body irradiation	1°C.	0.01	0.02

* Experimental evidence for critical energy flux in watt/cm² to cause intolerable effects of microwaves is compared with estimated values and resultant temperature elevation. Critical temperature rise is arbitrarily defined for case of total body irradiation and testes. Experimental evidence so far only obtained at 10 cm wavelength is given for the rabbit's eye by Williams[2] and total body irradiation by Ely and Goldman.[3] All tolerance values pertain to infinite exposure. Estimated critical flux levels refer to biologically effective, *i.e.*, absorbed energy. Experimental values refer to air prior to exposure. The latter values must be larger due to reflection from the body surface.

fever corresponding to a temperature rise in excess of 1°C. is considered intolerable. Cataract in the eye is produced when a temperature elevation of about 10°C. is caused, possibly due to denaturation of various macromolecular components.[2] It is seen from the Table that 1) estimates for critical flux levels, obtained as described later in this paper and experimental evidence, agree approximately, 2) critical flux values and temperature elevation are in proportion to each other, and 3) in total body irradiation eye damage is not limiting the flux value. We conclude that significant body temperature rise is the more serious hazard. This is true at least, whenever substantial parts of the body are exposed so that conditions of *total body irradiation* are approximated. The present investigation, therefore, is primarily concerned with the total body's absorption of electromagnetic radiation.

Presently available literature does not give any indication that other than purely thermal considerations are to be applied, *i.e.*, it is justified to assume that the effects of electromagnetic radiation are caused by the heat, which is generated by the mechanism of absorption.[7] Of primary interest in a discussion of the total body's response to electromagnetic radiation are the questions: 1) What percentage of airborne radiation is absorbed by the human body? 2) Where in the human body is the absorbed energy converted into heat? The answers to these questions, in combination with a knowledge of the amount of heat which can be tolerated by the human body, permits tolerance dosage recommendations. Previous discussions pertaining to this problem apply especially to lower frequencies, where the influence of skin may be neglected.[8,9] However, at frequencies above 1000 mc. which are of interest here, the assumption of a negligible skin influence is no longer justified.[10] We have undertaken, therefore, an investigation of the above formulated two problems under special consideration of the effects of skin. The discussion covers the total frequency spectrum from 150 to 10,000 mc. It covers the total range of practically interesting thickness values of subcutaneous fat and skin and it considers variability of results as function of temperature and dielectric data of tissues.

Method of Procedure

An enormous number of difficult experiments would be required on a purely experimental basis, in order to obtain conclusions of general value, extending over the total range of variability of parameters involved. A more theoretical approach seems, therefore, indicated. This approach utilizes the fact that the dielectric properties of various tissues involved and the arrangement of tissues of different dielectric properties in the body determine uniquely the mode of propagation of airborne electromagnetic energy into the human body. For simplicity, we assume plane electromagnetic radiation, propagating perpendicular to the surface of the body. This case will be approximated roughly by the trunk of a person facing the source of radiation. We can state that the percentage of absorbed energy will be a maximum under such conditions. Any conclusions drawn from such an approach will give, therefore, highest possible values of absorbed energy as should be considered in an attempt to establish tolerance dosage figures. The body itself is approximated by a triple layer arrangement as indicated in Fig. 1. The justification for this model is derived from the following facts: 1) All internal body tissues with high water content have comparable dielectric data similar to those of muscle and may, therefore, be represented by one type

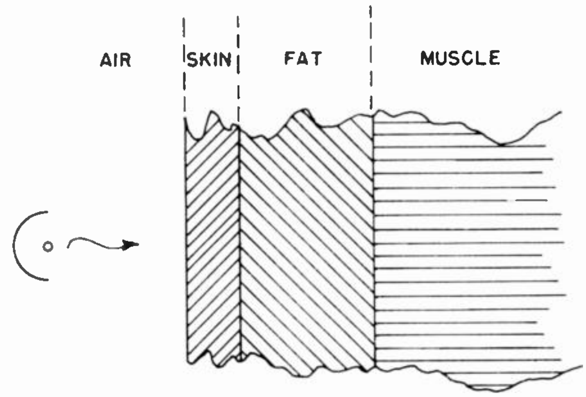

Fig. 1—Phantom arrangement simulating the human body's characteristics for absorption of uhf radiation.

[8] H. P. Schwan and E. L. Carstensen, "Application of electric and acoustic measuring techniques to problems in diathermy," *Trans. AIEE*, (Communications and Electronics) p. 106; May, 1953.
[9] H. P. Schwan, E. L. Carstensen, and K. Li, "Heating of fat-muscle layers by electro-magnetic and ultrasonic diathermy." *Trans. AIEE*, (Communications and Electronics) p. 483; September, 1953.
[10] T. Foelsche, "The energy distribution in various parts of the body due to irradiation with dm- and cm-waves," *Z. Naturf.* vol. 9b, p. 429; 1954.

of dielectric.[11] 2) Tissues of low water content inside the body are existent in the form of bone structures. But, they establish only a small part of the total body volume and only occasionally appear within the reach of the radiation and, if at all, only for lower frequencies than predominantly of interest today.[12] 3) Depth of penetration in the deep tissues inside the body is sufficiently small for electromagnetic radiation to permit approximation of the deep tissue complex by an infinitely extended medium as shown in Fig. 1. 4) Both subcutaneous fat, separating body surface and deep tissues, and skin have different dielectric data than the deep tissues.

Most of the dielectric data required for the calculations have been obtained throughout the total frequency range of interest by several investigators, notably by Schwan and Li between 100 and 1000 mc[11] and by Herrick, Jelatis, and Lee from 1000 to 10,000 mc.[13] They are well understood and explained in terms of the various body constituents.[6,11] An analysis of all data has shown that we may distinguish between two different classes of tissue, namely muscular tissues and body organs such as heart, liver, lung, kidney, etc., on the one hand and fat and bone tissue on the other hand.[6] These two classes of tissues have greatly different dielectric constants and losses. The first group of tissues is characterized by a rather high water content of about 70 to 80 per cent and a protein content of about 20 per cent in weight. Its electrical properties are found to be rather reproducible, due mainly to the constancy of the water-protein distribution. The fatty and bone tissues on the other hand have dielectric constants and conductivities which are about tenfold smaller than the data of the first group of tissues. The fat material varies strongly in its dielectric properties from sample to sample and perhaps from one type of animal to another. This may be due to the fact that it contains only small amounts of water. This water content is variable and accordingly effects the dielectric properties strongly due to the high dielectric constant and conductivity of water. A summary of the dielectric data of the two types of tissues is given in Table II. The *wet* fat in the table, with somewhat higher water content, represent horse fat, while the *dry* fat has been found more characteristic of pork. Human fat values are somewhat in between. Skin tissue so far has been investigated only by a few authors. Its dielectric data are slightly lower than those for muscular and similar tissues. Some of our skin measurements recently obtained with techniques described elsewhere[14,15] and those obtained by others at higher frequencies[16,17] are given in Fig. 2. The vari-

TABLE II*

ε	37°C.			50°C.	20°C.
	Muscle	Wet fat	Dry fat	Wet fat	Wet fat
mc					
150	66	7.6	3.8	7.6	7.6
400	58	6.8	3.4	6.8	6.8
900	54	6.1	3.05	6.1	6.1
3000	54	4.4	2.2	4.4	4.4
10,000	45	3.3	1.65	3.3	3.3

$10^3\kappa$	37°C.			50°C.	20°C.
	Muscle	Wet fat	Dry fat	Wet fat	Wet fat
mc					
150	10	0.66	0.33	1.32	0.33
400	10	0.78	0.39	1.56	0.39
900	11	0.91	0.45	1.81	0.45
3000	22	1.18	0.59	2.35	0.59
10,000	125	2.63	1.31	5.26	1.31

* Dielectric properties of muscle, characteristic for all tissues with high water content, and fat for various frequencies and temperatures. The muscle and *wet* fat data are from actual measurement. The data are simplified in the *dry* case since the total variability due to variation in water content is characterized by a factor of two. This seems an optimal value based on the limited available material as obtained from horse, pork, and human autopsy material. Temperature dependence of dielectric constant is small and has been neglected in the idealized data

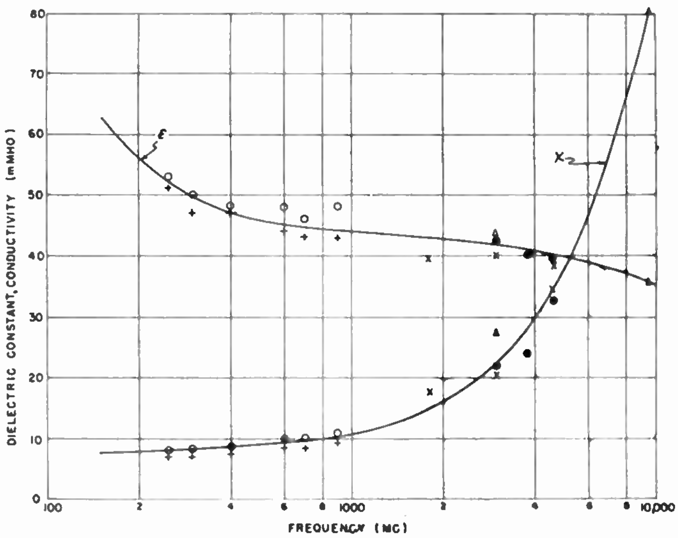

Fig. 2—Dielectric constant (ε) and conductivity (κ) (in mMho) of skin tissue. The results are obtained by the authors (0, +), England (Δ)[16] and Cook (●, x)[17]. The curves define the values used in all numerical discussions.

[11] H. P. Schwan and K. Li, "Capacity and conductivity of body tissues at ultrahigh frequencies," PROC. IRE, vol. 41, p. 1735–1740; December, 1953.

[12] This does not exclude the possibility that bone structures sufficiently near the body surface may occasionally cause standing wave pattern in the tissues separating body surface and bone due to impedance mismatch between bone and other body tissues. However, the influence of such effects on the total amount of absorbed energy is likely to be small and, therefore, does not deserve special consideration in a discussion of the thermal aspects of total body irradiation.

[13] J. F. Herrick, D. G. Jelatis, and G. M. Lee, "Dielectric properties of tissues important in microwave diathermy," *Fed. Proc.*, vol. 9, p. 60; 1950; and personal communication.

[14] H. P. Schwan, "Method for the determination of electrical constants and complex resistances, especially biological material," *ZS. f. Natur.*, vol. 8b, p. 1; 1953.

[15] H. P. Schwan and K. Li, "Measurement of materials with high dielectric constant and conductivity at ultrahigh frequencies," *Trans. AIEE*, (Communications and Electronics), p. 603; January, 1955.

[16] T. S. England, "Dielectric properties of the human body for wavelengths in the 1–10 cm range," *Nature*, vol. 166, p. 480; 1950.

[17] H. F. Cook, "Dielectric behavior of some types of human tissues at microwave frequencies," *Brit. J. Appl. Phys.*, vol. 2, p. 295; 1951.

ability of the skin data is characterized by two out of altogether five sets of measurements by us (200–1000 mc) and the deviation of the curve from the single observations. It is small enough to neglect it in the following discussions.

The calculations proceed as follows. The characteristic wave impedance of all tissues concerned, *i.e.*, *deep* tissues, fatty tissues, and skin, is obtained by dividing the wave impedance of air (377 ohm) by the square root of the complex dielectric constants $\varepsilon^+ = \varepsilon - j60\lambda\kappa$ where λ is the wavelength of radiation in air. The propagation constants of the radiation $\gamma = \alpha + j\beta$ are obtained from

$$\gamma = j\frac{2\pi}{\lambda}\sqrt{\varepsilon^+}.$$

Denoting by single subscripts each material (M muscle, F fat, S skin, A air) and by double subscripts interfaces, the following equations hold for characteristic impedances, input impedances, and complex reflection coefficients $p = \rho e^{j\Phi}$:

$$Z_M = Z_{FM} = \frac{377}{\sqrt{\varepsilon_M{}^+}}$$

$$P_{FM} = \frac{Z_{FM} - Z_F}{Z_{FM} + Z_F} = \frac{\sqrt{\varepsilon_F{}^+} - \sqrt{\varepsilon_M{}^+}}{\sqrt{\varepsilon_F{}^+} + \sqrt{\varepsilon_M{}^+}}$$

$$Z_{SF} = Z_F \frac{1 + P_{FM}e^{-2\gamma_F d_F}}{1 - P_{FM}e^{-2\gamma_F d_F}}$$

$$P_{SF} = \frac{Z_{SF} - Z_S}{Z_{SF} + Z_S}$$

$$Z_{AS} = Z_S \frac{1 + P_{SF}e^{-2\gamma_S d_S}}{1 - P_{SF}e^{-2\gamma_S d_S}}$$

$$P_{AS} = \frac{Z_{AS} - 377}{Z_{AS} + 377}$$

where the d symbolizes thickness of material under consideration. The field strength E in each tissue is resultant from incident waves and waves reflected from the interfaces between the tissue layers. Hence[9]

$$E = E_0[e^{-\gamma x} + pe^{+\gamma x}]$$

where x is the space coordinate ($x = 0$ at interface which is responsible for reflected wave). The parameters E_0 are determined by the boundary conditions, that no potential jump is permissible at the interfaces. From the field distribution, heat development per cm length is obtained:[9]

$$I = \frac{E_0{}^2}{2}\kappa\left[e^{-2\alpha x} + \rho^2 e^{2\alpha x} + 2\rho\cos(2\beta x + \Phi)\right].$$

Finally, the integrals of heat development per cm

$$\int_0^d I dx = \frac{E_0{}^2}{2}\kappa\left[\frac{1-\rho^2}{2\alpha}(1 - e^{-2\alpha d}) + \frac{\rho}{\beta}(\sin[2\beta d + \Phi] - \sin\Phi)\right]$$

are determined for the three layers and compared with each other. They give total heat development in skin, fat, and deep tissues.

Results

Coefficients which give total percentage of absorbed energy and heat development in skin, fat, and muscle have been determined for all parameters of interest as follows:

1) frequencies of 150, 400, 900, 3000, and 10,000 mc.
2) thickness values of subcutaneous fat of 0, 0.5, 1, 1.5, 2, 2.5, and 3 cm.
3) thickness values of skin of 0, 0.2, and 0.4 cm. The range of 0.2 to 0.4 cm skin thickness covers values of practical interest. The value 0 is included in order to permit judgment of what happens if skin is neglected.
4) dielectric constant and conductivity data of fat with high water content and low water content in order to investigate the effect of the variability of fat properties. Variability of the dielectric data of deep tissues and skin is very small and need not be considered.
5) dielectric data for fat of a temperature near 50°C. and fat of a temperature near 20°C. Influence of the temperature variation is demonstrated in the case of fat, since fat has been shown to vary its dielectric parameters with the temperature more strongly than any other tissue.[11] However, variation in temperature of subcutaneous fat causes no very pronounced effects as will be shown below. It has been necessary, therefore, to choose lower and upper temperature limits out of the range of physiological interest to demonstrate temperature influence. The dielectric data which have been used are summarized in Table II. Variation with temperature throughout the range of practical interest involves predominantly change in conductance and only to a smaller degree change in dielectric constant.[11] It is justified, therefore, to represent the change from 20 to 50°C. by a change of the conductivity by a factor of two, while the dielectric constant data are not varied.

Some statements are necessary to justify the neglecting of a discussion of the temperature dependence of the skin and deep tissue layer. It follows from numerical calculation not demonstrated here that the input impedance of the deep tissue complex varies only to a small degree with the temperature of the deep tissue layer. This is due to the fact that the dielectric constant is nearly temperature independent[11] and that the conductivity, whose temperature coefficient is about 2 per cent/°C., has practically no effect on the input impedance. The input wave impedance for the deep tissues is, furthermore, quite different from the characteristic wave impedance in the fatty layer, resulting in a pronounced reflection of energy from the fat-deep tissue interface. Variation in the input impedance of the deep tissue has, therefore, only a small affect on the standing wave pattern in front of the deep tissue layer. We conclude that the development of heat in all layers is practically independent of the temperature in the deep tissues and, hence, also the coefficient which characterizes the percentage of airborne radiation absorbed by the body.

The temperature of skin has some effect on its absorption of energy. It does not affect the ratio of heat developed in fat and deep tissue layer, but its own consumption of energy is found to vary by about 2 per cent per °C. This is small enough so that the percentage of incident energy which is absorbed by the body, and the skin's own heat development are affected only to a minor extent.

In view of the number of parameters, which are involved, a great amount of numerical data has been obtained. Space does not permit the presentation of all the material.[18] We will, therefore, restrict ourselves to part of the material which seems to be most characteristic.

Percentage of Absorbed Energy

Table III gives percentages of total absorbed energy for the frequencies 150 to 400 mc. The data demonstrate the simple situation prevailing at frequencies below 1000 mc. Comparison of the data shows practically no influence of the degree of wetness of the fatty material for thickness values of the subcutaneous fat layer up to 2 cm. The same applies for variation in temperature.

TABLE III*

Fat	Skin thickness	150 mc	400 mc
Wet 37°C.	0	26–27	36–42
	0.2 cm	26–31	36–54
	0.4 cm	27–32	37–60
Dry 37°C.	0	27–29	37–40
	0.2 cm	27–32	37–52
	0.4 cm	29–34	38–61
Wet 50°C.	0	26–28	37–43
	0.2 cm	26–30	37–52
	0.4 cm	27–33	37–59
Wet 20°C.	0	26–27	37–44
	0.2 cm	26–31	36–55
	0.4 cm	27–33	37–62

* Percentage of absorbed energy for frequencies of 150 and 400 mc. The data pertain to somewhat *wet* fat and rather *dry* fat at body temperature. The small effect of temperature is shown by listing data applying to extremes of temperature (20°C. and 50°C.). The ranges, which are given, are covered monotonously as the thickness of the subcutaneous fat increases from 0 to 2 cm.

The percentage of absorbed energy increases monotonously with fat layer thickness. The lower figures pertain, therefore, to zero-thickness and the higher figures for a thickness of 2 cm of fat. The increase is negligible for 150 mc for all values of skin thickness. It is still small for 400 mc. if the skin thickness is neglected, but becomes more pronounced as the skin thickness increases. The percentage values are equal to 30 per cent and vary only by ± 4 per cent at 150 mc. At 400 mc they are near 50 per cent ± 10 per cent. The increase in absorption becomes rapidly more pronounced as the fat thickness or thickness of skin increase beyond the values discussed in the table. The material indicates that at low frequencies skin and subcutaneous fat have only minor influence on the absorption percentage. This must be so, since for the thickness values under discussion, both skin and fat layer are considerably smaller than one quarter of a wavelength in either type of material and, therefore, almost transparent for the electromagnetic energy. However, the amount of absorbed energy is seen to be frequency dependent and to vary from amount 26 to 35 per cent if skin and fat layer are neglected and when frequency increases from 150 to 400 mc. These percentage values are characteristic for a semi-infinite layer of muscular tissue, hit by airborne radiation. Further details of this frequency dependence are shown in Fig. 3 in more detail.

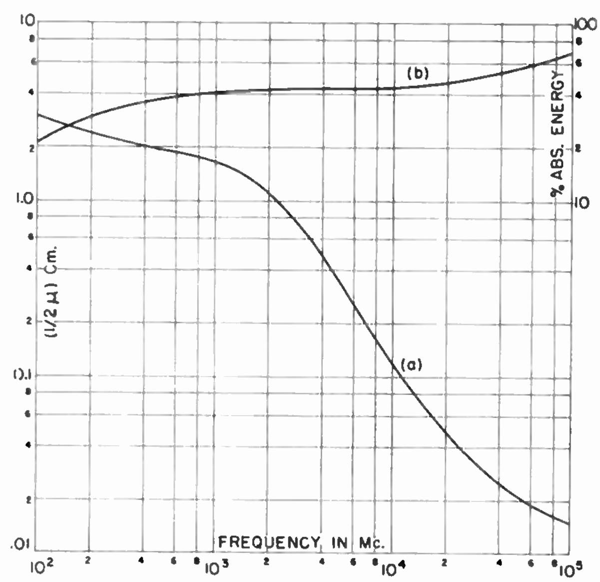

Fig. 3—Depth of penetration a) and percentage of absorbed energy b) of airborne electromagnetic radiation in tissues with high water content as function of frequency. The depth of penetration is defined as the inverse of 2μ (μ absorption coefficient) and characterizes the thickness required to diminish the field intensity to $1/e$ of its original value.

Fig. 3 demonstrates with curve *b* the continuous increase which the percentage of energy absorbed by a single semi-infinite layer of muscle shows with increasing frequency. A plateau exists where the percentage is nearly frequency independent from 600 to 10,000 mc and equal to about 40 per cent. The explanation of this plateau-effect has been given at another place.[6] Curve *a* shows a strong decrease of the depth of penetration ($\frac{1}{2}\mu$) with increase of frequency.

[18] Those interested in the detailed results may request them from the authors.

Fig. 4 has been chosen to demonstrate the tremendous influence which skin can have on the absorption, if its thickness becomes comparable to or greater than $\lambda/4$ in skin material. *Wet* fat has been assumed. This type of fat has the unusual property that it matches the input impedance of the deep tissue layer to the wave impedance of air, if it is a quarter of a wavelength thick. For this thickness, which is 1.25 cm for 3000 mc., 100 per cent energy absorption results if skin thickness is neglected. However, at a thickness of 4 mm the skin itself establishes $\lambda/4$ transformer, causing a very large mismatch. The result is an absorption of only 20 per cent. At other values of fat layer thickness than 1.25 cm, the dependence of percentage of absorbed energy on skin thickness is less pronounced.

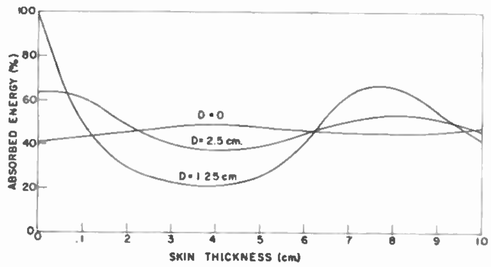

Fig. 4—Critical influence of skin thickness on the amount of absorbed energy is demonstrated for various fat layer thickness values at a frequency of 3000 mc.

Fig. 5 gives the coefficient of absorption in per cent for the frequencies 900 mc, 3000 mc, and 10,000 mc. Thickness of subcutaneous fat is varied between 0 and 3 cm and skin thickness values from 0 to 0.4 cm. These are the ranges of predominant practical interest.[19] The total material is affected by the occurrence of standing wave patterns in the fatty and skin layers. The corresponding periodicity of the absorption coefficient is obtained in good approximation if the wavelength in air $\lambda = 3.10^{10}$/frequency is divided by the square root of the dielectric constant values as given in Table II. The periodic behavior is further affected by the losses in the subcutaneous fat, which cause the curves to become less modulated as the fat layer thickness increases. The effect of skin and subcutaneous fat layer is pronounced at 900 and especially at 3000 mc. It affects the coefficients of energy absorption strongly whenever the thickness of these layers matches multiples of $\lambda/4$ as has been demonstrated for skin already in Fig. 4. The 10,000 mc values, on the other hand, are rather constant and fluctuate between 40 and 50 per cent for values of skin thickness of practical interest. The values are quite similar to the value taken from Fig. 3 for incident radiation of the same frequency hitting deep tissue material, but quite different from the values which are obtained at 10,000 mc if skin is neglected completely (curve $K=0$). This is explained by the similarity of skin tissue to the deep tissue components in regard to dielectric properties (compare Fig. 2 with Table II)

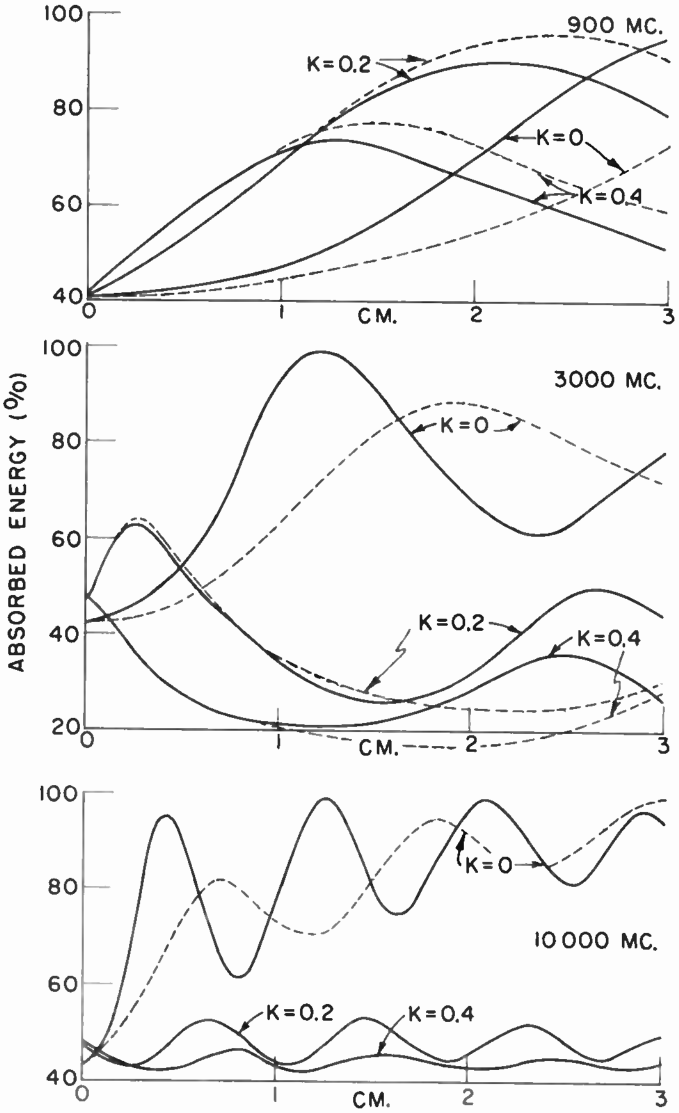

Fig. 5—Percentage of absorbed energy as function of thickness of subcutaneous fat layer. The parameter K refers to thickness of skin and is varied from 0 to 0.4 cm. Frequencies of 900, 3000 and 10,000 mc are considered. Wet fat (solid curves) and dry fat (dashed curves) of body temperature are assumed.

and the fact that the depth of penetration of 10,000 mc radiation in such tissues is extremely small. Its value is about 0.1 cm as demonstrated in Fig. 3, which means that the energy is almost completely absorbed in the skin. Only if skin is neglected, the percentage of absorbed energy must fluctuate due to pronounced variations of the input wave impedance at the body surface with the fat layer thickness.

The effect of variation in water content is illustrated in Fig. 5 by the dashed curves which pertain to *dry* fat, while the solid curves hold for *wet* fat. Available

[19] Larger values for the thickness of the subcutaneous fat may occur occasionally. However, such layers will almost never cover a substantial part of the human body's surface and will, therefore, be of no concern in a study of the effects of total body irradiation. The general relationships indicated in this study permit furthermore, judgment how the curves may be extrapolated to higher values of subcutaneous fat thickness.

dielectric data do not permit us to make final statements where human dielectric properties fall in Table I. However, it is almost certain that human fatty tissue is to be placed between the *wet* and *dry* type of fat recorded in the Table II.[20] The curves which characterize the coefficients of energy absorption in mankind will fall, therefore, somewhere between the solid and dashed curves given in Fig. 5. The general picture in the case of dry and wet fat is the same, with the peaks and minima of the curves presented in Fig. 5 appearing for dry fat at about $\sqrt{2}$ times higher subcutaneous fat thickness values. This is due to the decrease of the fat dielectric constant by a factor of 2 in the *dry* case. The strong temperature coefficient of the conductivity in fatty tissue, reported elsewhere,[11] makes it also necessary to investigate the dependence of above presented results from temperature. Such investigations have been conducted so far only for the case where the thickness of the skin may be neglected.[21] The result of this work shows that the effects of temperature variation are small. If skin is not neglected and at the very high frequency of 10,000 mc neither water content nor temperature of the subcutaneous fat can influence the coefficients of absorbed energy, since almost all the energy is absorbed already in the skin (Fig. 3). The same applies to the low frequencies of 150 and 400 mc as demonstrated above in Table III. Here the major part of the energy is absorbed in the deep tissues, indicating that skin and subcutaneous fat are rather transparent for the radiation. At 900 and 3000 mc, however, one may suspect a somewhat stronger influence of the temperature of fatty material. Table IV presents for these two frequencies the results obtained for temperatures of 20 and 50°C. for skin thickness values from 0 to 0.4 cm. It shows that even at 900 and 3000 mc the effect of temperature is not pronounced. Similar results are obtained for 20 and 50°C. even though a temperature variation had to be assumed of physiologically unreasonable magnitude, in order to demonstrate temperature effects.

The results presented above may be summarized as follows: At low frequencies well below 1000 mc and at high frequencies well above 3000 mc simple conditions exist. The percentage of absorbed energy is nearly independent of skin and subcutaneous fat thickness and near 40 per cent. However, in the range from 1000 to 3000 mc complicated conditions exist. Here the percentage of airborne energy, which is absorbed by the body, may vary between 20 and 100 per cent, depending upon how thick skin and subcutaneous fat are. The explanation for this fact is the ability of both skin and fat to transform the input wave impedance of the deep tissues over a considerable range of impedance values. This causes, depending on conditions, all possibilities from complete mismatch to almost exact impedance match with air with corresponding variability of the percentage of airborne energy absorbed by the body. Since tolerance considerations must be conservative, up to 100 per cent energy absorption must be assumed in an establishment of tolerance dosage for frequencies between 1000 and 3000 mc; and up to 50 per cent for frequencies either well below 1000 mc or well above 3000 mc.

[20] Only a restricted number of samples of horse fat, pork fat, and human autopsy material has been investigated by us. Horse fat and pork fat established so far the extremes in dielectric values. However, the statistical fluctuation is sufficiently great to render it impossible to make final statements in regard to the dielectric properties of human fatty tissue.

[21] H. P. Schwan and K. Li, "Variations between measured and biologically effective microwave diathermy dosage," *Arch. Phys. Med. and Rehab.*, vol. 36, p. 363; 1955.

TABLE IV*

	K=0 cm		K=0.2 cm		K=0.4 cm	
d_{cm}	20°C	50°C	20°C	50°C	20°C	50°C
900 mc						
0	41	41	41	41	42	42
1	48	48	71	68	71	69
2	70	69	91	86	66	66
3	98	89	77	79	50	55
3000 mc						
0	42	42	45	45	48	48
0.5	54	55	51	52	27	28
1	91	93	33	35	22	23
1.5	86	95	25	29	20	24
2	59	76	26	32	29	30
2.5	54	72	52	46	40	33
3	72	83	44	42	25	28

* The table gives percentage of incident radiation absorbed by the arrangement shown in Fig. 1 for 900 and 3000 mc and for two temperatures of subcutaneous fat (20°C. and 50°C.). Thickness of subcutaneous fat d is varied from 0 to 3 cm and thickness of skin K from 0 to 0.4 cm. The table illustrates the small effect of temperature variation on the percentage of absorbed energy.

Distribution of Heat Sources in Various Tissues

The following figures and tables explain where the energy, absorbed by the body, is transformed into heat. Fig. 6 gives heat developed in skin, subcutaneous fat, and deep tissues in per cent of the total energy, which is penetrating into the body. The results are given for 400, 900, 3000, and 10,000 mc. The upper, middle, and lower rows apply to skin thickness values of 0, 0.2, and 0.4 cm. The data are presented as function of thickness of subcutaneous fat layer over the range from 0 to 3 cm. Assumed is *wet* fat of body temperature (solid curves) and *dry* fat (dashed curves). The amounts of heat developed in fat and skin increase, of course, with the thickness of either type of tissue. Both heat in fat and skin are seen to increase also with frequency. The ratio of the amount of heat developed in fat to that in deep tissues is independent of the thickness of the skin layer, since it can be shown to be determined completely by the dielectric properties of both tissues and their thickness.[9] The amount of heat developed in the deep tissues is small at 10,000 mc. Even the amount of energy available in fat is small at 10,000 mc, unless skin is

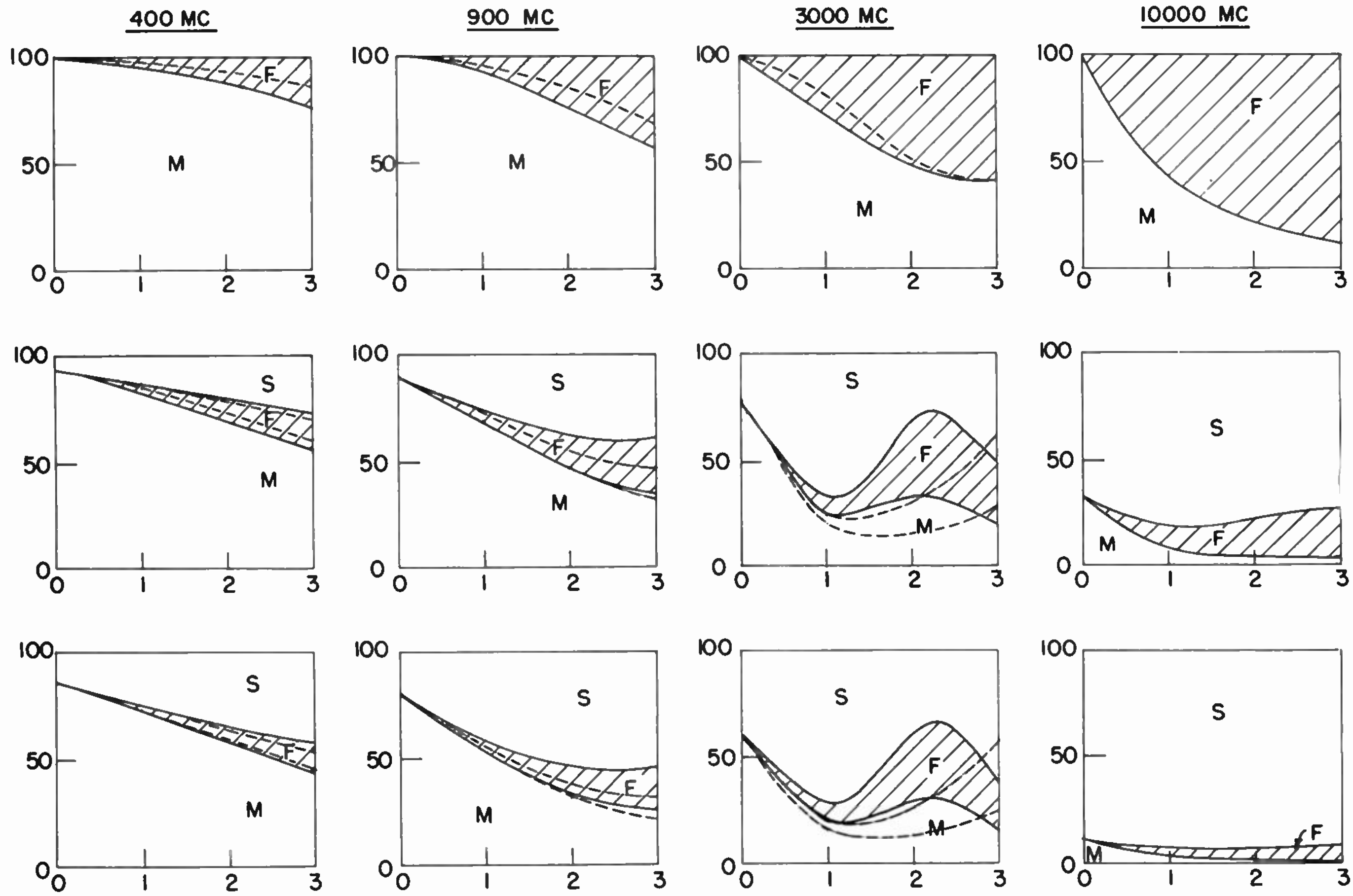

Fig. 6—Heat development in skin (S), subcutaneous fat (F) and deeper situated tissues (M) are given in per cent of total energy absorbed by the body as function of thickness of subcutaneous fat in cm. The upper row of graphs holds for a skin thickness K=0, the middle row for K=0.2 cm and the lower one for k=0.4 cm The solid curves pertain to fat with high water content and the dashed curves to dryer fat. The shaded areas emphasize the heat developed in fat in the wet case. For any particular combination of values of frequency, thickness of skin and fat the sum of all heat contributions developed in the three layers is 100 per cent in this presentation.

neglected. This demonstrates again that 10,000 mc radiation is absorbed completely in the surface of the body. This holds even more so at higher frequencies due to the continuous decrease of depth of penetration into skin as frequency increases. At the lower frequencies of 900 and 3000 mc a more complex situation exists. In general, the values characterizing heat developed in skin, fat layer, and deep tissues are somewhat more comparable with each other than at 10,000 mc, at least for the range of skin thickness values of practical interest. However, the values fluctuate strongly with all parameters involved. At 400 mc, almost all of the energy reaches into the deep tissues and the same applies, of course, at 150 mc even more so as our evaluations not demonstrated here show. In summary: heat development occurs predominantly in the deep tissues below 900 mc and at the body surface at frequencies above 3000 mc. The range from 1000 to 3000 mc establishes a transition period where more difficult relationships apply.

The effect of temperature of the subcutaneous fat on the results presented in Fig. 6 are discussed in Fig. 7. Here, results are given which pertain to wet fat of 20 and 50°C. It is demonstrated that the amount of heat, which is developed in the subcutaneous fat increases by about a factor of two as the temperature increases from 20 to 50°C. The curves pertaining to the lower temperature are placed in the areas characteristic for fat heating as obtained at higher temperatures, almost in all instances. This means that both skin and deep situated tissues benefit from the decrease of energy consumption in fat at lowered temperature. Since the range of temperature variation which is of physiological interest is at least five times smaller than the temperature range discussed in Fig. 7, effects of temperature variation of the subcutaneous fat can be neglected.

Conclusion

The thermal heat conductivity of subcutaneous fat is known to be about twofold smaller than the heat conductivity of deep tissues. The relatively poor blood supply of the fat tissue only emphasizes its ability to establish a thermal barrier, separating body interior from exterior. Noticeable temperature elevation inside the body is, therefore, necessary before sufficient temperature gradient across the subcutaneous fat is established to balance heat generation inside with escape

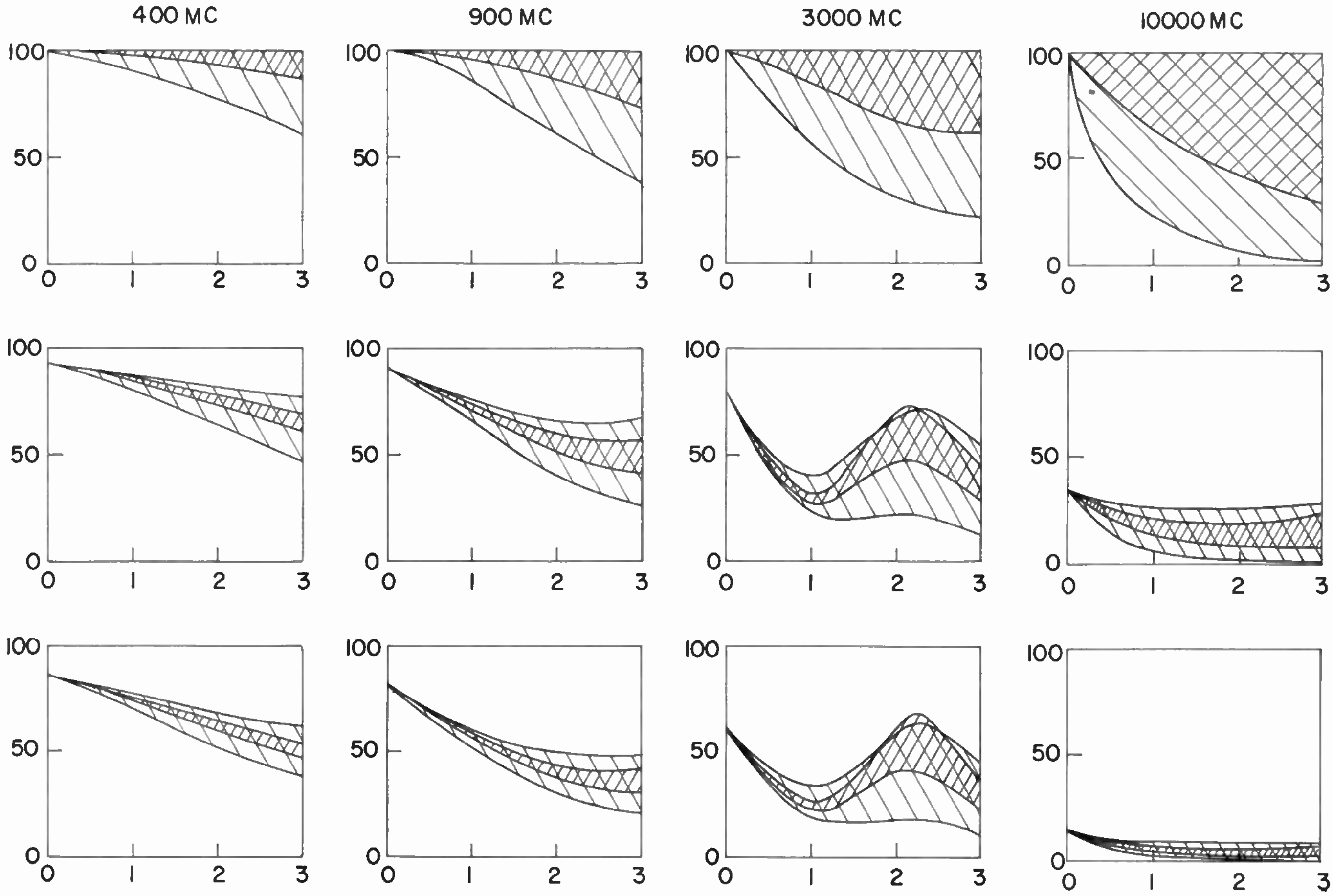

Fig. 7—Same as Fig. 6, except that influence of temperature variation of subcutaneous fat is shown. The densely shaded areas refer to a temperature of 20°C and the lightly shaded area (lines from upper left to lower right) to 50°C. Fat with high water content is assumed throughout.

from body surface. On the other hand, heat developed in the skin, *i.e.*, at the body surface will have difficulty in penetrating inside the body and will rather escape by means of the usual effective mechanism of heat regulation (radiation, evaporation, heat conduction). Thus, we recognize that radiation of such high frequency, that it is developing its heat in the body surface, must be much less apt to cause intolerable temperature elevation than radiation of lower frequencies. Presently available knowledge makes it difficult to state how much more heat, generated at the body surface, is tolerable than heat generated in the deep tissues. A more differentiated dosage statement must wait, therefore, until more research has been done concerning this aspect of heat physiology.

However, the human body's ability to tolerate heat may be estimated as follows:

1) Data are available pertaining to irradiation of restricted parts of the human body as performed for example in clinical practice,[5] (see also the detailed study by Cook[22]). The frequency in this case was 2500 mc. The experiments show that application of about 100 watts of radio frequency energy to an area of approximately 100 cm² results in temperature rise of about 5°C. in the first five minutes. From above discussions of coefficients of absorption, we know that this temperature increase of 1°C. per minute corresponds to an absorbed flux figure of 0.1 to 1 watt/cm². This experimental result is in fair agreement with numerical estimates, which utilize knowledge of the depth of penetration in deep tissues (about 1 cm, see Fig. 3) and give transient temperature rise of the volume which is defined by exposed area and depth of penetration. Clinical experience has shown, furthermore, that the considerable extension of blood vessels, which occurs when significant temperature elevation has happened, provides an effective means of counteracting excessive temperature rise.[4,5] Under such circumstances blood carries a good part of the developed heat away, *i.e.*, the mass of the total body becomes available as a cooling reservoir for the restricted part of the body irradiated in clinical practice. This means in effect that a steady state temperature is achieved which can be tolerated, *i.e.*, rapid temperature rise of 1°C. per minute in

[22] H. F. Cook, "A physical investigation of heat production in human tissues when exposed to microwaves," *Brit. J. Appl. Phys.*, vol. 3, p. 1; 1952.

the beginning of the transient period is soon replaced by a tolerable steady state temperature elevation. It is obvious that the steady state temperature elevation must depend critically on the ratio of irradiated part of the body surface to total body surface. It is safe to predict that it will increase with the area of irradiation. This means that a flux figure of about 0.3 watts/cm² must result in intolerable temperature rise when the irradiated area is larger than 100 cm². If we assume linearity between tolerance flux figure and ratio of nonirradiated to irradiated body surface, a figure of 0.03 watts/cm² is found dangerous if at least half of the body (*i.e.* about 1 m²) is exposed.

2) Average heat dissipation under normal circumstances is about 0.005 watts/cm². This figure is based on an energy uptake in form of food of 3000 Kcal per day, an efficiency of somewhat below 30 per cent and a body surface of about 2 m². Only under unusually fortunate circumstances is the body surface able to handle tenfold higher heat flux figures. However, double the above rate seems well within the capacity of the human body. This means that it is permissible to develop inside the human body an additional amount of energy which corresponds to 0.005 watt/cm², averaged over the total body surface. In view of the fact that, at most, only half the body can be subjected to radiation, a figure of 0.01 watt/cm² absorbed energy appears as tolerable and is, therefore, suggested as a tolerance dosage. This value should not be exceeded except under unusual circumstances, where cooling efficiency of body surface is excellent.

3) An attempt must be made to supplement a tolerance statement with regard to energy flux by a total tolerance dosage, *i.e.*, a statement of optimal tolerable product of exposure time and energy flux absorbed. This is of particular interest for short time exposure to very high intensities where heat flow is not very effective, *i.e.*, whenever time of exposure is small compared with the time constants which characterize heat exchange in the human body. In such cases, we operate in the transient period where temperature rise is linear with time. For a 10 cm radiation, penetrating into muscular tissue, it has been mentioned already that a temperature rise of 1°C. per minute must be considered for a flux of about 0.3 watts/cm². If we consider temperature elevation of more than 1°C. intolerable in the case of total body irradiation we derive a figure of 0.3 watt minutes/cm² as limiting value. Since depth of penetration of radiation decreases with increasing frequency, this figure is to be replaced by higher values at frequencies below 1000 mc, 2000 mc and lower values above 3000 mc.

Taking 0.01 watt/cm² for long time exposure and 0.01 watt hour/cm² for short exposures as tolerance figures, both not to be exceeded in case of total body irradiation, and incorporating the above discussed values for percentage of absorbed energy and location of energy exchange, the following conclusions seem justified:

1) Frequencies substantially below 1000 mc (500 mc and lower): We deal with true deep heating. Coefficient of absorption is about 30 to 40 per cent. This means that incident energy flux figures of less than 0.03 watt/cm² can be tolerated.
2) Frequencies from 1000 to 3000 mc may be absorbed completely. Skin, subcutaneous fat, and deep tissues participate in this absorption and conversion into heat in a complex manner. Hence, 0.01 watt/cm² is considered as a recommendable tolerance statement.
3) Frequencies in excess of 3000 mc are absorbed in the surface of the body. Heat dissipation to the outside is, therefore, excellent. The coefficient of airborne energy, which is absorbed is 40 to 50 per cent. Hence, more than 0.02 watt/cm² are tolerated by the body.

An Analysis of Pulse-Synchronized Oscillators*

GASTON SALMET†

Summary—The present extent of the number of radio communications has led to an overcrowding of transmitting frequencies. It is therefore desirable, especially in variable frequency transmitters or receivers, to be able to make, according to circumstances, a swift choice of the proper working frequency.

On the other hand, in phase shift telegraphy and in single sideband transmitters, a very high long-term accuracy in the carrier wave is needed.

Hence, the main problem is the design of an easily tunable, high-precision variable oscillator with, of course, a limited number of crystals.

In this respect, probably the most famous present technique consists in the use of a variable oscillator frequency synchronized on pulse harmonics issued by a quartz oscillator.

This system is referred to, hereafter, as an "Impulse Governed Oscillator" (IGO).

This paper gives a mathematical analysis of the above system together with its circuits. The main difficulties met in its design as well as the way to overcome them are examined.

Introduction

The original network, designed a few years ago at the Philips Laboratories, enabled a variable frequency oscillator, covering the range of, say, 5 to 10 mc, to be synchronized on any harmonic of a 100 kc quartz pulse generator. Out of a single quartz oscillator, fifty different frequencies could thus be obtained, the different frequencies being as precise and stable as those of the crystal.

However, this result, though interesting, does not offer the possibility of obtaining out of an oscillator all the frequencies included in the range of 5 to 10 mc or even very close signals separated from each other by, say, 10 kc. In the latter case, a 10 kc quartz should be used, and accordingly, the oscillator would be synchronized on quartz harmonics ranging from the five-hundredth to the one-thousandth harmonic. Obviously, this is impracticable.

Thus, we were led to design an indirect synchronization network derived from the "Impulse Governed Oscillator" (IGO) system and described hereafter. This circuit offers the possibility of obtaining either a continuous range of frequencies or a large number of synchronization points. Precise frequency deviation in frequency telegraphy can also be obtained through this system, with deviation altogether independent of the working frequency.

This circuit, in avoiding the use of a quartz per working frequency, resolves very satisfactorily problems of precision frequency control with transceivers, and numerous applications, interesting for their relative low cost, were realized on this ground at the Télécommunications Radioélectriques et Téléphoniques (TRT) (see Figs. A, B, C, and D).

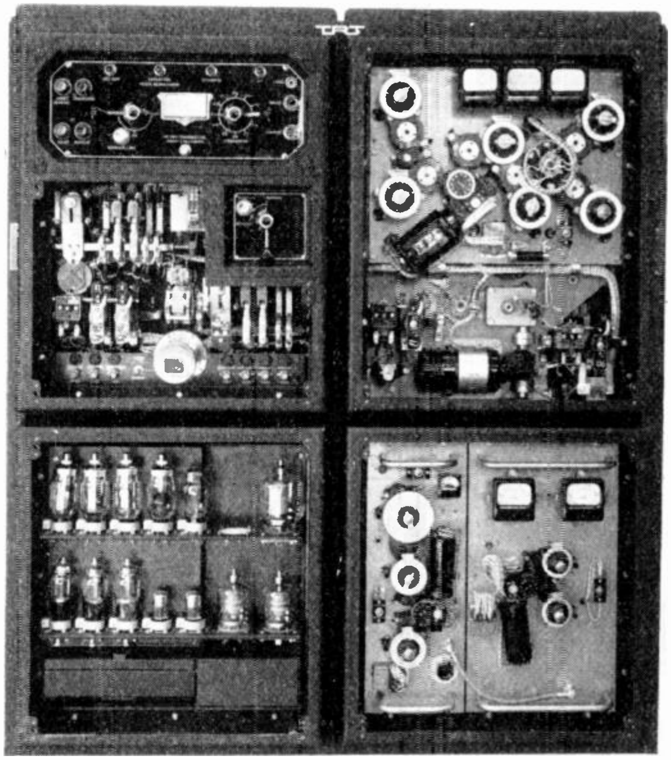

Fig. A—TRT 400 w mobile military transmitter with IGO master oscillator unit. Frequency range: 2 to 24 mc. Preset frequencies: 12 with remote control. Types of transmission: $A_1-A_2-A_3-F_1$.

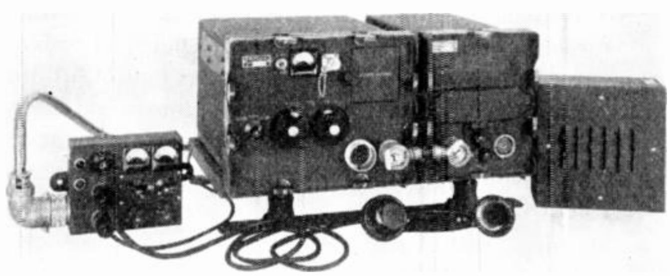

Fig. B—TRT 20 w military transceiver set. Main Features: Type of frequency control: indirect synchronization systems. Frequency range: 27 to 38.9 mc. Number of channels: 120 immediately available from a remote control box. Modulation: frequency modulation. Frequency carrier accuracy: $\pm 10^{-4}$. Crystal: one 166.66 kc crystal.

The IGO circuit offers also the possibility of being frequency modulated by phase modulation of the synchronization pulses: the ratio of the oscillator's frequency to that of the pulses is usually very sufficient to obtain in the 300–3000 cps audio frequency range, the required deviation with an acceptable phase shift.

However simple the principles of the IGO circuit may

* Original manuscript received by the IRE, July 11, 1955; revised manuscript received, July 10, 1956.

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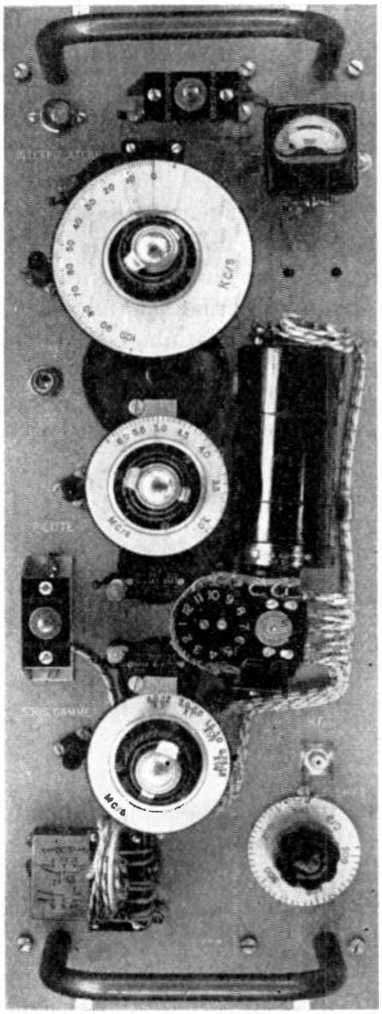

Fig. C—TRT master oscillator unit for transmitters (automatic version). Similar to the manual type except that 12 preset frequencies are available from a remote control base.

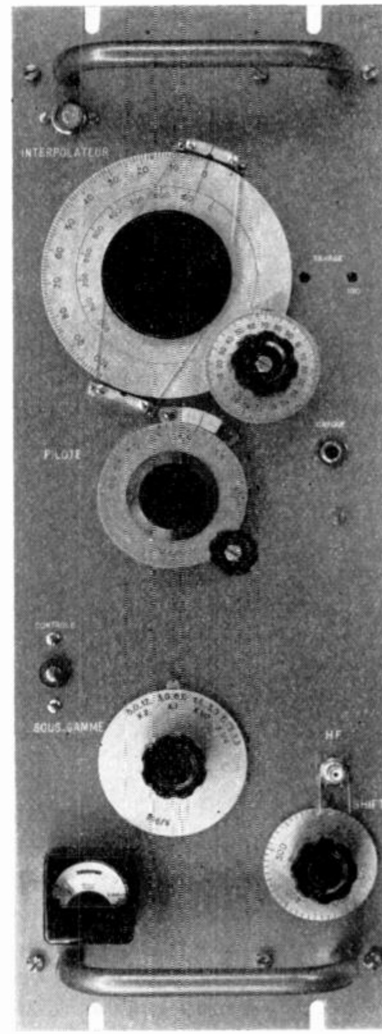

Fig. D—TRT master oscillator unit for transmitters (manual version). Principle: IGO system with indirect and phase follower synchronization. Main features: 1) Frequency range: 0.75 to 12 mc. 2) Frequency accuracy: $\pm 5, 10^{-6}$. 3) Output level: about 6 v with 500 Ω loading impedance. 4) Shift telegraphy: adjustable from 100 to 500 cps. 5) Crystal: one 100 kc crystal.

seem, there often appear in the design secondary phenomena which make the realization of such networks a difficult art calling for a very particular skill.

The following difficulties must usually be overcome in the design of an IGO circuit.

1) Appearance of spurious oscillations.
2) Insufficiency of oscillator's tuning limits within which synchronization may take place however large is the range within which it is possible to maintain an established synchronization.
3) Parasitic phase modulation (especially in the indirect synchronization system).
4) Poor low-frequency curve of response and loss of synchronization at high frequencies in phase modulated oscillators.

An analytical examination of the IGO circuit would allow, of course, a thorough examination of these deficiencies.

Unfortunately, the basic differential equation of the IGO system is nonlinear; hence, direct and complete analysis is impossible. Thus, the principal aim of this study is to show as simply as possible, through certain simplifications and valid hypotheses, the main features of the synchronized oscillator theory. It is also desired to underline the advantages of a new circuit known as the "Phase Follower Synchronization" system, which reduces substantially the above inconveniences.

It should be emphasized that the results obtained hereafter do not apply merely to the IGO circuit but, more broadly, to any frequency regulated system through phase comparison.

Principles of the IGO Circuit

The essential parts of an IGO circuit are (Fig. 1):

1) A single-frequency, high-precision oscillator (usually a crystal oscillator).
2) A pulse generator, synchronous with the crystal.
3) A variable frequency oscillator to be synchronized on a harmonic of the pulse generator.
4) A reactance modulator circuit allowing a frequency variation to be obtained in the above oscillator through application of a signal on its in-

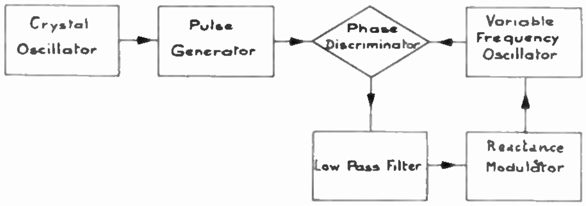

Fig. 1.

put, regardless of tuning elements (variable capacitor, for instance).

5) A phase discriminator circuit, supplying a dc voltage, the value of which depends on the phase difference between the pulses and oscillator signals. The pulses and the sinusoidal voltage are added and the result rectified.

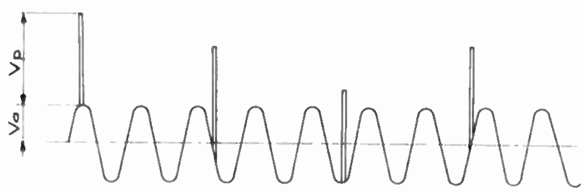

Fig. 2.

It appears from Fig. 2 that if Va is the peak value of the sine wave, and Vp, pulse amplitude, and if $Vp > 2Va$, with pulse duration much shorter than half a sinusoidal cycle, rectified voltage will vary, according to the phase, between $Vp+Va$ and $Vp-Va$. A symmetric network such as shown in Fig. 3 would enable eliminating Vp in the final result, supplying a dc voltage ranging from $+Va$ to $-Va$.

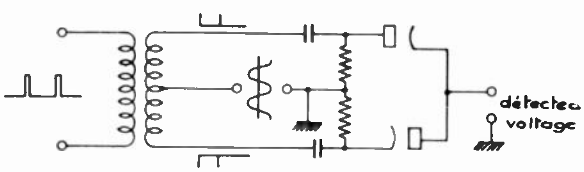

Fig. 3.

6) A low-pass filter eliminating hf components of the voltage issued by the phase discriminator and delivering only the dc signal to the reactance modulator circuit.

It should be noted that in most practical cases, the discriminator constitutes a voltage generator of high internal impedance. Accordingly, components entering the filter are limited to resistances and capacities and one should not overlook the time constant introduced by the rectifying elements of the discriminator.

Assuming in Fig. 1 that the link between the phase discriminator and the reactance modulator network is opened, if frequency Fa of the variable oscillator comes close to nFq (nth harmonic of the pulse oscillator Fq) then the phase discriminator will deliver (Fig. 2) a low frequency $(Fa-nFq)$ sinusoidal voltage with Va as peak value.

When the above loop is reconnected, regardless of its sense, there will always be a favorable moment for frequency correction since the slope of Va inverts periodically. In other words, when the loop is closed, it is always possible to find a moment when a variation of Fa produces, through the phase discriminator, a voltage variation on the input of the reactance modulator tending to oppose the very variation of Fa.

The network can then be stabilized on a value Vc of the signal applied on the input of reactance modulator circuit so that $Fa = nFq$. The phase of the pulse signal relative to the sinusoidal voltage is hence constant and of the appropriate sign of dVc/dFa (Fig. 4).

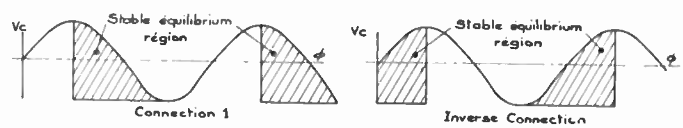

Fig. 4.

Now, if the frequency of the oscillator is varied slightly from the tuning position, the oscillator's frequency will remain synchronous of nFq, the variation in the tuning being balanced by a proper variation in the phase discriminator's output.

Of course, for every tuning position of the oscillator there is a different value of the phase, and the tuning limits where a synchronism already set up is maintained are referred to as the "synchronization range." These limits depend solely on the frequency variation that could be obtained, on one hand, by proper action on the input of the reactance modulator, and on the other hand, by the phase discriminator's maxima and minima outputs. This zone may be evaluated in kc, and it represents the frequency variation that will be registered by the oscillator with no voltage correction.

Another important factor in the theory of the IGO circuit is the "catching range," which is the zone where the system passes from the nonsynchronous to the synchronous range. It may be illustrated as follows.

Assume the oscillator is so detuned as to be out of synchronism: say, on a lower frequency than nFq. If Fa is approaching of nFq, for a given value F_1 of Fa, synchronization will take place. Now, if the operation is resumed with the oscillator detuned on a higher frequency than nFq, synchronization will be reached for a frequency $Fa = F_2 \neq F_1$. The difference $F_2 - F_1$ actually represents the "locking range." It is obvious that this zone is narrower than, or at most equal to, the synchronization range since it depends upon the filter's static characteristics and also upon the transmitting filter's attenuation for an ac signal in the absence of any synchronization.

TRT Indirect Synchronization System[1]

In this system, as shown in Fig. 5 (where for clearness' sake figures are given), the oscillator voltage to be regulated is not directly compared in the phase discriminator to the reference pulses, but is mixed with the voltage supplied by an interpolator covering a frequency range equal to the difference between two adjacent harmonics of the pulse frequency.

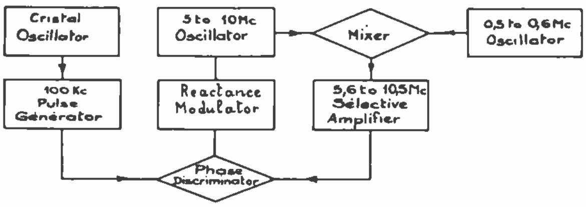

Fig. 5.

Through a selective amplifier, a proper frequency is chosen (in the case of Fig. 5, the sum).

It is this frequency which is added to the pulses in the phase discriminator. The latter's dc output is, hence, applied on the reactance modulator input, which circuit controls the oscillator frequency. It appears, from Fig. 5, that when the principal oscillator's frequency varies, there will be synchronization points whenever the sum of the frequencies of the principal and interpolation oscillators are equal to the frequency of a pulse harmonic. Accordingly, as the interpolator covers a frequency range equal to the pulse frequency, it will be possible, by an appropriate choice of the interpolator frequency and of the harmonic upon which the mixer operation is synchronized, to tune the principal oscillator on any frequency included in the above range.

The frequency precision of such a network is excellent. If the reference quartz circuit is carefully designed, the total error is practically reduced to that of the interpolator, the frequency range of which is much more reduced than the principal oscillator's. The accuracy is of the order of $5 \cdot 10^{-5}$; error is mainly due to the calibration of the interpolator which can be synchronized, according to the IGO principle, by subharmonic pulses of Fq ($Fq/10$ for instance). These can easily be obtained from a multivibrator or a blocking oscillator synchronized by the crystal.

In the case of Fig. 5, there will be 500 synchronization points for the principal oscillator without any great tuning accuracy since when one of the interpolator's 10 frequencies is chosen, there will only remain 50 possible tuning points for the principal oscillator.

Basic Equation of the IGO Circuit

In order to simplify the final expression, we shall assume that the oscillator's frequency variation is a linear function of the voltage applied on the input of the reactance modulator. This, though introducing some quantitative error, does not change the main feature of the physical phenomena, nor does it lead to wrong conclusions.

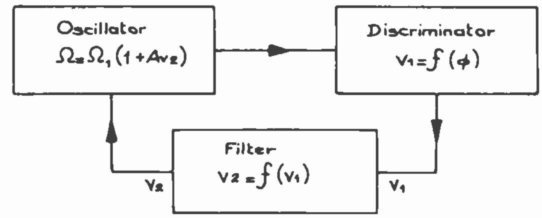

Fig. 6.

The IGO circuit, as shown in Fig. 6, can be reduced to a three-block diagram.

1) A variable oscillator, frequency-regulated by the reactance modulator's input signal.
2) A phase discriminator supplying a voltage, the amplitude of which is a junction of the oscillator's phase relative to the reference pulses.
3) A linear passive filter.

The oscillator's angular velocity is, in accordance with the above assumption,

$$\Omega = \Omega_1(1 + Av_2) \tag{1}$$

where A is the sensitivity of reactance modulator and v_2 is the voltage applied on input of reactance modulator. It can easily be seen that, assuming pulse duration is small compared to the sinusoidal period, the phase discriminator's output v_1 will be

$$v_1 = B\left[\sin \int_0^t (\Omega - \Omega_0)dt + \phi_0\right] \tag{2}$$

where B is a constant of proportionality, Ω is oscillator's angular frequency, $\Omega_0 = n\Omega_q$ angular frequency of pulse harmonic nearest to Ω, and ϕ_0 = initial constant.

Eq. (2) simply shows that v_1 is a sinusoidal function of the instantaneous phase difference between the pulses and the oscillator's voltage, and if the pulses' frequency is supposed to be constant, it may also be written, in introducing operational notations,

$$\int_0^t \cdots dt = \frac{1}{p}$$

$$v_1 = B \sin\left(\frac{\Omega}{p} - \Omega_0 t\right). \tag{3}$$

The filter being linear, it follows that

$$v_2 = v_1 f(p). \tag{4}$$

From (1), (3), and (4), the following equation can be deduced:

$$p\phi - AB\Omega_1 f(p) \sin \phi = \Omega_1 - \Omega_0 \tag{5}$$

[1] French Patent No. PV 624,171; U.S. Patent No. 337,592.

where

$$\phi = \int_0^t (\Omega - \Omega_0)dt + \phi_0.$$

This is the general equation of the IGO system.

Owing to the term in $\sin\phi$ no general solution can be given to this expression. A particular solution can be found, if ϕ is assumed to be constant.

Hence

$$p\phi = \Delta\Omega = 0 \tag{6}$$

showing that the angular velocity difference $(\Omega - \Omega_0)$ is zero; *i.e.*, the oscillator is synchronized when ϕ equals ϕ_1, such as

$$\sin\phi_1 = \frac{\Omega_0 - \Omega_1}{AB\Omega_1} \tag{7}$$

since for $\Delta\Omega = 0$, $f(p) = 1$ (filter attenuation for dc assumed to be equal to zero).

The phase is hence a function of the difference between the synchronizing frequency and the oscillator frequency, the oscillator frequency being that existing where no corrective signal is applied on the reactance modulator input.

It appears from (7) that synchronization is only possible for a relative angular velocity $|\Omega_0 - \Omega_1| < |AB\Omega_1|$. This condition is mainly imposed by the phase discriminator; however, in practice, it is the reactance modulator whose action is often reduced to a rather narrow frequency range, and not the phase discriminator, which introduces the above restriction. Theoretically, the synchronization range would then extend to a value of $AB\Omega_1/\pi$.

Negative Feedback Analysis of the IGO Circuit

In order to obviate the difficulties introduced by (5), we shall assume the IGO circuit to be synchronized. It can, hence, be considered as a simple amplifier, with negative feedback. Accordingly, for its analysis, a theory of closed loop systems will be applied.

Let, then, a small sinusoidal voltage be applied on the amplifier input. Since the amplitude of the signal is low, (5) may be considered as linear and

$$\sin\phi = K\phi.$$

Further, as K varies with the phase discriminator output (upon which the oscillator is synchronized), K may be assimilated to B, the discriminator slope, hence

$$\sin\phi \cong \phi. \tag{8}$$

On the other hand, $j\omega$ may be substituted to p, as only steady-state performances are considered. The circuit-block diagram is then as shown in Fig. 7.

Evaluating the feedback-coefficient $\mu\beta$ which is equal to the open circuit gain of the loop, we have, in Fig. 7, when the circuit is opened in points a and b:

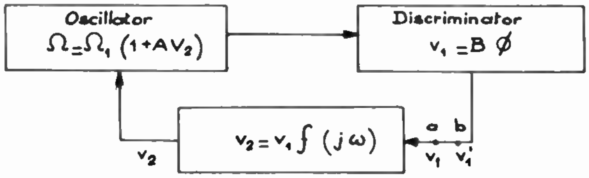

Fig. 7.

$$\mu\beta = \frac{v_1'}{v_1} = \frac{AB\Omega_1}{j\omega} f(j\omega). \tag{9}$$

Conditions of Stability in a Pulse-Synchronized Oscillator

With the feedback coefficient $\mu\beta$ being known, and given the filter response curve, it is now possible to determine the network conditions of stability using a Nyquist diagram.

Simple filter networks will only be considered so that the imaginary part of $\mu\beta$ will have a single zero as frequency varies from 0 to ∞. The system will then be stable if $\mu\beta$ is less than $+1$ at the frequency for which the imaginary part vanishes.

The circuits considered will be made up of resistances and capacitances only, since the phase discriminator has, in general, a high internal impedance and it is rather hard to design filters containing high-value, large Q inductances, due to the risks of spurious resonances.

Examining (9), clearly it is harder to comply with stability conditions in an IGO system than in a simple negative feedback amplifier or even in a frequency (but not phase) stabilized oscillator. The difficulty arises from the term $j\omega$ in the denominator which increases by $-\pi/2$ the phase margin of $f(j\omega)$. As $f(j\omega)$ tends towards zero, for high values of ω, its phase will tend towards a negative value, to which the $-\pi/2$ phase lag will be added. Obviously this will increase the chances of undesired oscillations.

Since feedback correction can only take place with A and B coefficients of opposite sign, $\mu\beta$ can be written

$$\mu\beta = \frac{-K}{j\omega} f(j\omega) \tag{10}$$

with

$$K = |AB\Omega_1|. \tag{11}$$

The simplest network between the oscillator and the reactance modulator is a single RC circuit shown in Fig. 8 and should this network be used, the system would be stable since the $-\pi$ phase lag would be reached only for $\omega = \infty$; v_2/v_1 will then be equal to 0.

Unfortunately, the simplest form of acceptable network is made up of 2 RC cells in chain (Fig. 9), the first, R_1C_1, representing the detection time constant; the second, R_2C_2, being the separator between the discriminator and the reactance modulator. Attenuation of such a circuit is given by the following relation:

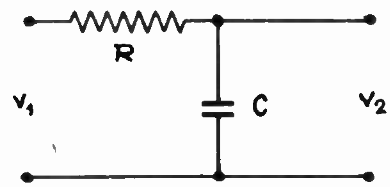

Fig. 8.

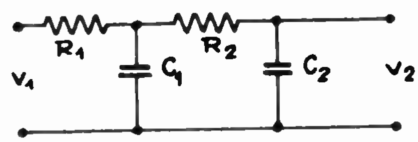

Fig. 9.

$$\frac{v_2}{v_1} = \frac{1}{1 - \tau_1\tau_2\omega^2 + j(\tau_1 + \tau_2 + \tau_{12})\omega} \tag{12}$$

where

$$\begin{aligned} \tau_1 &= R_1C_1 \\ \tau_2 &= R_2C_2 \\ \tau_{12} &= R_1C_2. \end{aligned} \tag{13}$$

From (10), it follows:

$$\mu\beta = \frac{K}{(\tau_1 + \tau_2 + \tau_{12})\omega^2 + j\omega(\tau_1\tau_2\omega^2 - 1)} \cdot \tag{14}$$

The imaginary part is cancelled for

$$\omega_0^2 = \frac{1}{\tau_1\tau_2} \tag{15}$$

for which frequency

$$\mu\beta = \frac{K\tau_1\tau_2}{\tau_1 + \tau_2 + \tau_{12}} \cdot \tag{16}$$

The circuit will be stable for $\mu\beta < 1$; *i.e.*, for

$$K < \frac{\tau_1 + \tau_2 + \tau_{12}}{\tau_1\tau_2} \cdot \tag{17}$$

Eqs. (16) and (17) suggest the following remarks.

1) $\mu\beta$ increases as the value of τ for equal ratios of time constants.
2) $\mu\beta$ increases proportionally to the value of K, *i.e.*, the synchronization zone, for equal values of time constants. It is then obvious that the higher the frequency, the more difficult is the design of an IGO system; as a matter of fact, the synchronization range ought to represent a given constant minimum percentage of the working frequency and this implies that the synchronization zone will be proportional to the frequency. However, time constants cannot be indefinitely reduced.
3) The increase of a single time constant entails the increase of $\mu\beta$, hence, the chances of unwanted oscillations. This is opposite to what was obtained with frequency-stabilized oscillators where increasing any single time constant reduces eventual instabilities.

Stabilizing Networks

It is possible to reduce oscillation risks of a circuit in introducing some elements intended to increase the attenuation of the quadripole filter at the value of ω for which the imaginary part of $\mu\beta$ vanishes.

One of the simplest circuits fulfilling this purpose is shown in Fig. 10. It is similar to that shown in Fig. 8 except for the 2 components (resistance R_3 shunted by capacitance C_3) that were added in series. A complete analysis of the system would show that the value to be given to the time constant $\tau_3 = R_3C_3$ is, approximately,

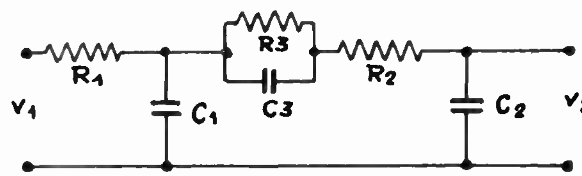

Fig. 10.

$$\tau_3 = \tau_1 - \frac{\tau_{32}}{4} + \sqrt{\left(\tau_1 + \frac{\tau_{32}}{4}\right)^2 + \tau_1\tau_2}. \tag{18}$$

The corresponding value of $\mu\beta$ is

$$\mu\beta = \frac{K}{\tau_1\tau_2C_2\alpha\omega_0^4 + (\tau_1 + \tau_2 + \tau_{12} + C_2\alpha)\omega_0^2} \tag{19}$$

with

$$\omega_0^2 = \frac{1}{\tau_3^2 - 2\tau_1\tau_3} \tag{20}$$

and

$$\alpha = \frac{R_3}{1 + \tau_3\omega_0^2} \cdot \tag{21}$$

Assume, as an example, that

$$\tau_2 = \tau_1; \qquad R_1 = R_2$$
$$R_3 = 5R_2,$$

hence

$$\tau_{32} = 5\tau_1.$$

It follows from (18) that $\tau_3 = 2.2\tau_1$, and from (19),

$$\mu\beta \cong \frac{K\tau_1}{11.7} \cdot$$

From (16), the value of $\mu\beta$ (without stabilizing network but with the same values of τ_1 and τ_2) is found

$$\mu\beta = \frac{K\tau_1}{3} \cdot$$

Obviously, on cancellation of the reactive component, $\mu\beta$ is, with the stabilizing network, 3.3 times weaker for the same time constants τ_1 and τ_2 than without, other conditions remaining unchanged.

According to the value of the τ_2/τ_1 ratio the stability margin gain is not always as high, but is appreciable, however, in most practical cases.

As an example, suppose

$$\tau_2 = \frac{\tau_1}{5}$$

with $R_1=R_2$ and $R_3=5R_2$ as before. It follows that

$$\mu\beta = \frac{K\tau_1}{11.6}$$

with stabilizing network, and

$$\mu\beta = \frac{K\tau_1}{7}$$

without stabilizing network.

Stability margin gain has then decreased.

However, if τ_2 is made equal to $10\tau_1$, it follows that with stabilizing network,

$$\mu\beta = \frac{K\tau_1}{57},$$

and without,

$$\mu\beta = \frac{K\tau_1}{2}.$$

Accordingly, not only is the stability margin gain important, but, furthermore, $\mu\beta$ is now, by far, the weakest value found.

The conclusions reached in the foregoing paragraphs, where it was underlined that no advantage resulted in increasing any particular time constant in a 2 time-constant circuit, do not hold hence when an error controller is introduced in the system. Obviously, it is very desirable to have τ_2 large with respect to τ_1.

Practically, the detection time constant τ_1 will be reduced to the utmost, while τ_2 will do the filtering.

Frequency Modulated IGO Circuit

As mentioned before the possibility of using an IGO circuit as a high-stability frequency-modulated oscillator, with synchronization pulses phase modulated, seems very attractive.

For pulse harmonic $n\Omega_q$ the frequency deviation is

$$\Delta\Omega_0 = j\omega n\varphi \qquad (22)$$

where ω is the modulation frequency, and φ is the instantaneous phase variation of the crystal oscillator.

The maximum phase deviation that may be expected with a simple phase modulator is of, approximately, ± 1 radian. Hence, the maximum frequency deviation at the lowest modulation frequency ω_0 is

$$\Delta\Omega_0 = n\omega_0. \qquad (23)$$

Usually, the multiplication power n is adequate to obtain the desired frequency deviation in the audio frequency band (300–3000 cycles). Accordingly, to have a frequency deviation regardless of the modulation frequency, one should modulate the phase of the pulses through an inversely proportional gain-to-frequency amplifier.

If, for whatever frequency, K, the coefficient of efficiency in the control circuit, was large enough, and the filter attenuation was zero, the oscillator would follow synchronization pulses without any sensible phase lag. Hence, the modulation response curve would be linear which is not the case in practice, unfortunately.

We shall only consider a single time-constant quadripole. Reverting to Fig. 7, where the phase discriminator delivers a signal proportional to the phase difference between the oscillator and the pulse generator outputs, and if $\Omega_m/2\pi$ is the nth harmonic of the frequency modulated pulses,

$$\Omega_m = \Omega_0 + Me^{j\omega t} \qquad (24)$$

hence,

$$\phi_m = \Omega_0 t + \frac{m}{j\omega} \qquad (25)$$

where m is the instantaneous value of $Me^{j\omega t}$.

However, due to the circuit's feedback, the relative phase value ϕ_r in the discriminator will be different from (25).

The value of ϕ_r, the oscillator being synchronized, is:

$$\phi_r = \frac{m}{j\omega}\left(\frac{-1}{1-\mu\beta}\right). \qquad (26)$$

The negative sign before ϕ_r arises from $\phi_r=\phi_1-\phi_m$ where ϕ_1 is the oscillator's phase.

As may easily be seen, this phase variation will produce a frequency modulation of the oscillator such as

$$\Delta\Omega = \frac{mKf(j\omega)}{j\omega(1-\mu\beta)} = -m\frac{\mu\beta}{1-\mu\beta} = \frac{m}{1-\dfrac{1}{\mu\beta}}. \qquad (27)$$

In the case of a single time-constant circuit, we have

$$\mu\beta = \frac{-K}{j\omega(1+j\tau\omega)}, \qquad (28)$$

hence,

$$\Delta\Omega = \frac{m}{1-\dfrac{\tau\omega^2}{K}+\dfrac{j\omega}{K}} \qquad (29)$$

and if

$$K\tau = U, \qquad (30)$$

$$\Delta\Omega = \frac{m}{j\omega\tau\left[j\left(\dfrac{\tau\omega}{U}-\dfrac{1}{\tau\omega}\right)+\dfrac{1}{U}\right]}. \qquad (31)$$

The quantity between the brackets is seen to be equivalent to the impedance of a series RLC network whose Q, at resonance, is $\sqrt{U}$.

Now the product of the synchronization range by the time constant is usually high, since, for safety sake, synchronization range is made as large as possible, *i.e.*, close to the pulse frequency. On the other hand, in order to have an interesting filtering, $1/\tau$ must be at least 10 times less than the synchronization range.

It is conformable to express (31), in terms of the ratio

$$\frac{\omega}{K} = \frac{\text{modulation frequency}}{\text{synchronization range}}$$

and assuming

$$\frac{\omega}{K} = X,$$

it follows

$$\Delta\Omega = \frac{m}{\sqrt{(1 - UX^2)^2 + X^2}} \cdot \tag{32}$$

Three curves of $\Delta\Omega/m$ for 3 values of U ($U = 10$, 30, and 100) are shown in Fig. 11.

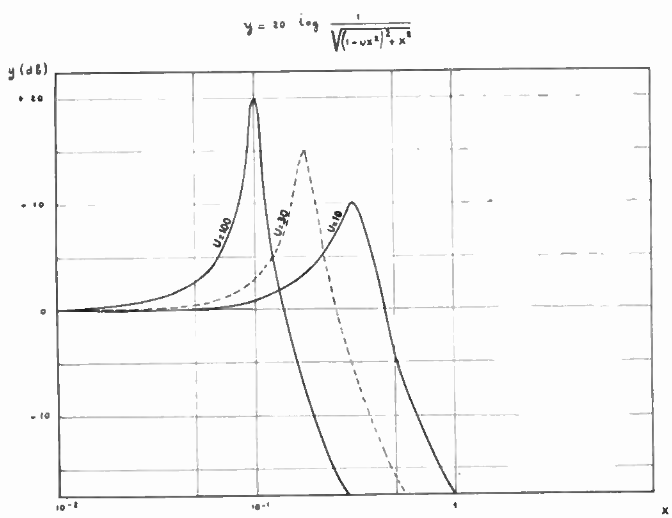

Fig. 11.

It appears therefrom that if, for instance, $U = 100$ and $K = 2\pi \times 50$ kc, resonance takes place for a 5 kc modulation frequency. This case corresponds actually to a time constant of about $\tau = 3 \cdot 10^{-4}$.

Since ω/K is small, it can be stated, from (29), that modulation will always be fairly linear up to the highest modulation frequency $\omega_2/2\pi$ provided

$$\tau \ll \frac{K}{\omega_2^2} \cdot \tag{33}$$

This condition will be more easily satisfied as the synchronization zone will be larger.

It should be also noted that the curve of response could scarcely be corrected through action on the audio frequency amplifier, coefficient K being an active element whose value depends on tube characteristics, working frequency, discriminator and reactance-modulator outputs, etc. . . .

Obviously, on one hand, if in increasing τ_2, in the case of a 2 time-constant filtering network and a phase corrector, a better stability, hence, a less acute resonance is obtained, now, on the other hand, it lowers the resonance frequency and may bring it within the transmitted af band. This underlines the difficulty in obtaining a satisfactory compromise under certain circumstances.

Another difficulty that was not yet dealt with may appear when the frequency excursion is large and modulation frequency high. It is the network desynchronization. If x volts are to be applied on the reactance modulator input to obtain the desired frequency deviation at modulation frequency $\omega/2\pi$, and if the attenuation of the quadripole is N, in absolute value, for this frequency, the circuit would work satisfactorily as long as the discriminator output is x/N volts.

Spurious Frequencies in IGO Circuits

In the simple IGO system illustrated in Fig. 1, spurious frequencies may appear either on the oscillator or on the modulator reactance network, owing to the presence of pulse residues. Such parasitic oscillations may easily be suppressed.

However, more complex problems appear in indirect synchronization circuits.

Reverting to Fig. 5, should the selective amplifier be inadequate to completely eliminate the spurious frequencies F, $F-f$, $F\pm 2f$, etc. . . . , the desired component $F+f$ will be amplitude modulated at the frequency f. As a matter of fact, when f is close to a harmonic (n) of the pulse frequency, the discriminator output will contain a component in $(f-nFq)$. This frequency may be very low, hence, pass through the transmitting filter and, by action on the input of the reactance modulator, produce a spurious phase modulation on the master oscillator.

The phenomena will be more acute as the ratio between the transmitter and the master oscillator working frequencies is large; spurious phase deviation is obviously increased in the same ratio.

Practically, a single combined tuning is provided for the master oscillator and the selective amplifier. This implies, of course, that the latter will have a bandwidth equal at least to the crystal frequency, increased by the synchronization range, say, about 200 kc in the case of Fig. 5. Obviously, elimination of undesired parts of the mixture is not thus facilitated.

Analytically, relations obtained in the preceding sections can be used in substituting $m/j\omega$, phase modulation, by r, percentage of spurious modulation of the mixing frequency.

If $\omega = 2\pi(f - nFq)$, (27) can be written

$$\Delta\Omega = \frac{j\omega r}{1 - \dfrac{1}{\mu\beta}} \cdot \tag{34}$$

In the case of a single time-constant circuit, it follows by changing (31),

$$\Delta\Omega = \frac{r}{\tau\left[j\left(\frac{\tau\omega}{U} - \frac{1}{\tau\omega}\right) + \frac{1}{U}\right]}. \tag{35}$$

Eq. (35) can also be written in the following form:

$$\Delta\Omega = \frac{rK}{j\left(\tau\omega - \frac{K}{\omega}\right) + 1}. \tag{36}$$

At the resonance,

$$\Delta\Omega = rK. \tag{37}$$

Eqs. (36) and (37) illustrate the fact that the spurious frequency modulation is proportional to the product of the spurious modulation ratio of the mixture by the synchronization range. Maximum takes place for $\omega = \sqrt{K/\tau}$ and the value of $\Delta\Omega$ is then independent of the time constant.

It is advisable to make τ large. Even if the maximum value of $\Delta\Omega$ does not change, the bandwidth of maximum noise is now reduced.

The product rK may become important.

If $K = 2\pi \cdot 100$ kc and $r = 1$ per cent, $\Delta\Omega/2\pi = 1$kc. Hence, filtering of the mixing should be very efficient to reduce eventual effects of this phenomenon.

When ω is an audible frequency, the spurious modulation appears at the receiver in the form of an undesirable whistle if the receiver is very selective, or as a double tonality in the case of A_1 working of the transmitter.

When ω has a greater value, the sidebands produced by modulation are the causes of trouble. In fact, as in this case, the modulation index $\Delta\Omega/\omega$ is usually small, there are practically only two sidebands, the amplitude of which compared to carrier is

$$\left|\frac{v_2}{v_1}\right| = \left|J_1\left(\frac{\Delta\Omega}{\omega}\right)\right| = \sim \left|\frac{\Delta\Omega}{2\omega}\right|$$

$$= \frac{1}{2}\,\frac{rK}{\sqrt{(\tau\omega^2 - K)^2 + \omega^2}}. \tag{38}$$

When $\omega/2\pi$ is much higher than the resonance frequency, we get

$$\left|\frac{v_2}{v_1}\right| = \sim \frac{1}{2}\,\frac{rK}{\tau\omega^2} = r\,\frac{\omega_0^2}{\omega^2} \tag{39}$$

with

$$\omega_0 = \sqrt{\frac{K}{\tau}}.$$

Analysis of the Catching Zone

A complete mathematical analysis of the "Catching Zone" is given in the Appendix.

It results therefrom that it is very difficult to have a catching zone as large as a synchronization band, at least when the latter has a bandwidth near to the pulse frequency. The required condition is, hence, $K\theta$ next to one; then if K is about half the pulse angular velocity, this condition will not comply with the pulse detection and filtration requirements.

In fact, the maximum that can usually be obtained is a catching zone of about 30 per cent of the pulse frequency. In the band covered by the variable oscillator, there will be large spaces without any certain synchronization. Accordingly, should the number of channels be high, the oscillator tuning will be a delicate matter and its working, unreliable.

To obviate the difficulty, we have designed an electromechanical system intended to catch the synchronization when the oscillator is brought within the synchronization band, regardless of the size of the catching zone.

Phase Follower Synchronization System[2]

The foregoing study has shown that besides its good possibilities, the IGO system has also, owing to its very principle, some inconveniences, summarized as follows.

1) Tendency to instability, especially at high frequencies.
2) Lack of handling ease.
3) Appearance of spurious frequencies.
4) Critical tuning.

For a given network, these inconveniences can be overcome, at least partly, through an appropriate technical skill, but satisfactory results are scarcely painless.

Accordingly we were lead to design a new circuit known as the "Phase Follower Synchronization System" having the IGO circuit advantages without its inconveniences.

Obviously, the fault with the IGO system is the small angle included between $-\pi/2$ to $+\pi/2$ where phase control exists, as compared to the frequency band for which synchronization is desired.

The Phase Follower Synchronization System precisely increases, in substantial proportions, the phase control angle.

This result is obtained by phase modulating the pulses as from the discriminator output and in such a way that the pulses will follow to correct an eventual phase difference between the oscillator and the pulses.

Assuming that the phase modulation of the pulses is of the order of $\pm\pi/2$, equivalent modulation compared to the oscillator period will be multiplied by n, the harmonic ratio between oscillator and pulse frequency.

[2] French Patent No. PV 681,875; U.S. Patent No. 553,132.

As n is usually very large (often of the order of 50 or 100), phase control will increase substantially.

Analysis of the Phase Follower Synchronization System

Let us first consider the synchronized states, and evaluate the new value of the coefficient $\mu\beta$. The new diagram is shown in Fig. 12. It differs from the classical IGO system by the appearance of a new feedback chain acting directly on the discriminator in order to alter the coefficient B.

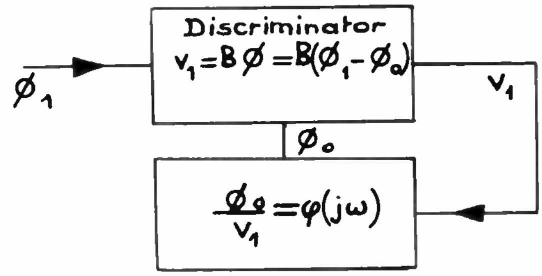

Fig. 12.

If ϕ_d is the oscillator phase with respect to pulse signal, ϕ_1 oscillator phase, ϕ_m pulse phase compared to a cycle of oscillator signal, and $\varphi(j\omega)$ transfer function of the transmission and phase modulation system, the value $\mu_1\beta_1$ of the said feedback chain is

$$\mu_1\beta_1 = B\varphi(j\omega). \tag{40}$$

For a given variation ϕ_1 of oscillator phase, there follows a relative phase ϕ_d variation:

$$\phi_d = \frac{\phi_1}{1 + \mu_1\beta_1} = \frac{\phi_1}{1 + B\varphi(j\omega)}$$

as ϕ_m is of opposite sign to ϕ_1.

This is then equivalent to a phase discriminator with a coefficient B_1, such as:

$$B_1\phi_1 = B\phi_d = \frac{B\phi_1}{1 + B\varphi(j\omega)}. \tag{41}$$

Since maximum value of ϕ_d is $|\pi/2|$, and maximum modulation of the pulse phase is also of about $\pi/2$, it is obvious that, should both maxima come to coincidence, (at least for small values of ω_0) the system's advantages will be used to the utmost. (It is not practically necessary to transmit dc component of ϕ_0).

In other words, since the pulse maximum phase compared to the oscillator time cycle is $n\pi/2$, it follows that

$$\mu_1\beta_1 = nf_2(j\omega) \tag{42}$$

where $f_2(j\omega)$ will be close to one for small frequency values.

It follows then:

$$B_1 = \frac{B}{1 + nf_2(j\omega)}. \tag{43}$$

If, in (10) B is replaced by B_1, the feedback transfer function of the Phase Follower Synchronization System is

$$\mu\beta = \frac{-Kf(j\omega)}{j\omega[1 + nf_2(j\omega)]}. \tag{44}$$

Practically, it is interesting to choose $f_2(j\omega)$ near one for all values of ω so that $f(j\omega)$ is not close to 0. In other words, if $f_2(j\omega)$ represents a time-constant circuit, this should be as small as possible. On the other hand, as

$$n \gg 1$$

hence

$$\mu\beta = \sim -\frac{K}{n}\frac{f(j\omega)}{j\omega}. \tag{45}$$

This equation is similar to (10) with K replaced here by K/n.

The following advantages of the phase follower system are then deduced.

1) Risks of instability are largely decreased. For instance, from (16) $\mu\beta$, when positive and real, is here divided by n.
2) Undesirable lateral bands appearing for large values of ω will here be n times weaker as may easily appear from (39) and (40). Further, from (37) maximum value of spurious frequency modulation is also divided by n.
3) Less obvious in this system are the advantages of the oscillator pulse phase modulation, however, a more suitable modulation process is available here: modulating the signal directly on the reactance modulator input. This is possible since the pulses can follow the wide phase deviation corresponding to the frequency modulation which avoids synchronization from vanishing.

If m_1 is the instantaneous frequency deviation with synchronization's loop open, the actual frequency deviation when synchronization exists is

$$m = \frac{m_1}{1 - \mu\beta}. \tag{46}$$

For a single time-constant circuit, it follows:

$$|M| = M_1\sqrt{\frac{\omega^2 + \tau^2\omega^4}{\omega^2 + \left(\frac{K}{n} - \tau\omega^2\right)^2}}. \tag{47}$$

For large enough values of ω, M is approximately equal to $|M_1|$; the time constant τ will have to be chosen so that for the lowest modulation frequency $\omega^2 \gg K/n\tau$ so as to obtain a sensible linear modulation; in the telephonic band, this condition can be easily obtained as n is generally high.

4) Coefficient K seems to be reduced; now the catching zone has substantially increased.

Obviously, the Phase Follower Synchronization System does not change the synchronization range since the discriminator's maximum output signal has not changed.

The new value of the catching zone will not be evaluated here; it should be noted, however, that, according to what is seen in the Appendix, the condition for which the catching zone equals the synchronization range ($4K\tau = 1$), corresponding to the critical damping in (56a) becomes, in the Following Phase Synchronization System, $4K\tau/n = 1$.

As the synchronization range has not changed, everything takes place just as if the time constant τ was n times weaker.

For large values of n, as is usually the case, there is no difficulty in obtaining a catching zone very close to the synchronization band.

The above list of advantages of the Phase Follower Synhcronization System shows that this circuit, though more complicated, offers greater possibilities than the conventional one, and is of interesting use in many cases, especially when associated with indirect synchronization systems.

Appendix

Analysis of the Catching Zone

Reverting to (5), the basic equation of the IGO system, can be written.

$$p\phi + Kf(p) \sin \phi = \Omega_d \tag{48}$$

with

$$\Omega_1 - \Omega_0 = \Omega_d \tag{49}$$

and

$$AB\Omega_1 = -K \tag{50}$$

Ω_d being, as already shown, the difference between the oscillator angular velocity without any correction signal, and the angular velocity of the pulse harmonic considered.

The catching zone will then be equal to twice the value of $\Omega x/2\pi$; Ωx, being the limit value of Ω_d, so that for the time t infinite, (48) still has a solution that is not a constant. When $t = \infty$, $d\phi/dt$ cannot be infinite; hence, it will be a periodic time function, and the phase ϕ, the sum of a periodic function and a linear time function.

For a single time constant τ, the value of $f(p)$ is $1/(1+p\tau)$ and (48) can then be written:

$$p\phi + \frac{K \sin \phi}{1 + p\tau} = \Omega_d$$

or, again, by multiplication of $(1+p\tau)$

$$\tau p^2\phi + p\phi + K \sin \phi = \Omega_d \tag{51}$$

as Ω_d being constant, we have $p\tau\Omega_d = 0$ which can be written in classical notation:

$$\tau\phi'' + \phi' + K \sin \phi = \Omega_d. \tag{52}$$

The limit value of Ω_d for which (52) accepts, at infinity, a periodic term cannot be calculated from the above equation. However, an approximate value can be obtained by making (52) linear during a whole cycle. Then let

$$K \sin \phi = K\phi \qquad \text{for} \quad \phi = -\frac{\pi}{2} \text{ to } \frac{\pi}{2}, \tag{53a}$$

$$K \sin \phi = K(\pi - \phi) \quad \text{for} \quad \phi = \frac{\pi}{2} \text{ to } \frac{3\pi}{2}. \tag{53b}$$

The function $f(\phi)$ determined by (53a) and (53b) and replacing $\sin \phi$ will have the graphic form of Fig. 13.

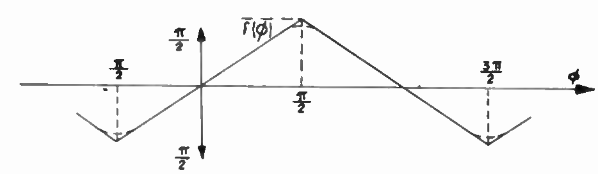

Fig. 13.

If $f(\phi)$ is a periodic function repeated to infinity, (52) can then be transformed into a system of 2 linear equations alternately valid according to the value of ϕ

$$\tau\phi_1'' + \phi_1' + K\phi_1 = \Omega_d \qquad \phi = -\frac{\pi}{2} \text{ to } \frac{\pi}{2} \tag{54a}$$

$$\tau\phi_2'' + \phi_2' + K(\pi - \phi_2) = \Omega_d; \quad \phi = \frac{\pi}{2} \text{ to } \frac{3\pi}{2} \tag{54b}$$

and, generally, for the $(n+1)$ cycle,

$$\tau\phi_1'' + \phi_1' + K(\phi_1 - 2n\pi) = \Omega_d;$$

$$\phi = 2n\pi - \frac{\pi}{2} \text{ to } 2n\pi + \frac{\pi}{2} \tag{55a}$$

$$\tau\phi_2'' + \phi_2' + K[-\phi_2 + (2n+1)\pi] = \Omega_d;$$

$$\phi = 2n\pi + \frac{\pi}{2} \text{ to } 2n\pi + \frac{3\pi}{2}. \tag{55b}$$

It is obvious, then, that from (55a) and (55b) ϕ and $f(\phi)$ are, necessarily, continuous time functions.[3]

Accordingly, the initial conditions enabling to determine the proper particular solution of (55b) will also be the final conditions given by (55a), and reciprocally.

Function $f(\phi)$ being supposed periodic with respect to time, in order to appreciate the actual value of the

[3] From those two equations, it can easily be proved that the first and second derivatives of ϕ with respect to time are also continuous time functions.

catching zone, the limit value of Ω_d, for which this function is no more periodic, *i.e.*, ϕ tends towards a constant value, has to be appreciated.

General solutions for (54a) and (54b) are, respectively,

$$\phi_1 = e^{-\alpha t}(A \sin \omega_1 t + B \cos \omega_1 t) + \frac{\Omega_d}{K}$$

$$\phi_1 = -\frac{\pi}{2} \text{ to } \frac{\pi}{2} \tag{56a}$$

$$\phi_2 = e^{-\alpha t}(Ce^{\omega_2 t} + De^{-\omega_2 t}) + \pi - \frac{\Omega_d}{K}$$

$$\phi_2 = \frac{\pi}{2} \text{ to } \frac{3\pi}{2} \tag{56b}$$

with:

$$\alpha = \frac{1}{2\tau}, \qquad \omega_1 = \sqrt{\omega_0^2 - \alpha^2}$$

$$\omega_2 = \sqrt{\omega_0^2 + \alpha^2}, \qquad \omega_0 = \sqrt{\frac{K}{\tau}} \tag{57}$$

Solution for (54a) has been supposed, at first, sinusoidal as, otherwise, with $\alpha \geq \omega_0$, ϕ_1 would have been asymptotic to Ω_d/K. Accordingly, except for the value of $\Omega_d/K > \pi/2$, ϕ would not reach the value of $\pi/2$. Hence, in this case, ϕ_1 is necessarily moving towards a stable state when the oscillator frequency is within the zone where synchronization is possible. In other words the catching zone is then equal to the synchronization range.

It could also be shown that the periodicity limit of $f(\phi)$ is given, by function ϕ_2 only, but for $\alpha \geq \omega_0$ which was examined above.

The limit of ϕ_2 is given by the value of the coefficient C of the exponential increasing term. As a matter of fact, let

$$\phi_2 = e^{-\alpha t}(\epsilon e^{\omega_2 t} + De^{-\omega_2 t}) + \pi - \frac{\Omega_d}{K} \tag{58}$$

and if ϵ is an infinitely positive small quantity of the first order, when t becomes infinite, ϕ_2 tends also towards infinite; for, from (57), $\omega_2 > \alpha$ and, on the other hand if $\epsilon = 0$, ϕ_2 is, at most equal to $\pi - \Omega_d/K$, hence, less than $3\pi/2$.

Eq. (58) will then be taken as the limit for which ϕ allows still a periodic solution.

For $t = 0$, $\phi = \pi/2$ from which $D = -\pi/2 + \Omega_d/K$, (58) then becomes

$$\phi_2 = e^{-\alpha t}\left[\epsilon e^{\omega_2 t} - \left(\frac{\pi}{2} - \frac{\Omega_d}{K}\right)e^{-\omega_2 t}\right] + \pi - \frac{\Omega_d}{K} \tag{59}$$

for $t = 0$ we get

$$\phi_2'(0) = (\omega_2 + \alpha)\left(\frac{\pi}{2} - \frac{\Omega_d}{K}\right). \tag{60}$$

For $\phi_2 = 3\pi/2$, $\epsilon e^{\omega_2 t}$ must tend towards a finite value, hence, $e^{+\omega_2 t}$ towards an infinite value. Accordingly $e^{-\omega_2 t}$ tends towards zero. There finally remains

$$\phi_2(t_1) = \frac{3\pi}{2} = \epsilon e^{(\omega_2 - \alpha)t_1} + \pi - \frac{\Omega_d}{K} \tag{61}$$

as well as

$$\phi_2'(t_1) = (\omega_2 - \alpha)\epsilon_e^{(\omega_2 - \alpha)t_1}. \tag{62}$$

Replacing in (62) $\epsilon e^{(\omega_2 - \alpha)t_1}$ by its value drawn from (61), we have

$$\phi_2'(t_1) = \left(\frac{\pi}{2} + \frac{\Omega_d}{K}\right)(\omega_2 - \alpha).$$

Analyzing function ϕ_1, it becomes, for $t = 0$, $\phi_1 = -(\pi/2)$ and, in accordance with what was stated before, on the function's continuity:

$$\phi_1'(0) = \phi_2'(t_1) = (\omega_2 - \alpha)\left(\frac{\pi}{2} + \frac{\Omega_d}{K}\right).$$

These two initial conditions determine the values of A and B in (56a), hence,

$$\phi_1 = e^{-\alpha t}\left(\frac{\pi}{2} + \frac{\Omega_d}{K}\right)\left(\frac{\omega_2 - 2\alpha}{\omega_1} \sin \omega_1 t - \cos \omega_1 t\right) + \frac{\Omega_d}{K}, \tag{63}$$

then

$$\phi_1' = -\alpha\left(\phi_1 - \frac{\Omega_d}{K}\right) + e^{-\alpha t}\left(\frac{\pi}{2} + \frac{\Omega_d}{K}\right)[\omega_1 \sin \omega_1 t + (\omega_2 - 2\alpha) \cos \omega_1 t]; \tag{64}$$

when t takes a value t_1 such as

$$\phi_1 = \frac{\pi}{2}$$

we must have

$$\phi_1'(t_1) = \phi_2'(0) = (\omega_2 + \alpha)\left(\frac{\pi}{2} - \frac{\Omega_d}{K}\right).$$

Eqs. (63) and (64) have, then, the respective forms

$$\left(\frac{\pi}{2} - \frac{\Omega_d}{K}\right) = \left(\frac{\pi}{2} + \frac{\Omega_d}{K}\right)e^{-\alpha t_1} \cdot \left(\frac{\omega_2 - 2\alpha}{\omega_1} \sin \omega_1 t_1 - \cos \omega_1 t_1\right) \tag{65}$$

$$(\omega_2 + 2\alpha)\left(\frac{\pi}{2} - \frac{\Omega_d}{K}\right) = \left(\frac{\pi}{2} + \frac{\Omega_d}{K}\right)e^{-\alpha t_1}[\omega_1 \sin \omega_1 t_1 + (\omega_2 - 2\alpha) \cos \omega_1 t_1]. \tag{66}$$

Dividing (66) by (65) and after simplification, it becomes,

$$\tan \cdot \omega_1 t_1 = -\sqrt{\left(\frac{\omega_0}{\alpha}\right)^4 - 1}. \tag{67}$$

So $w_1 t_1$ can be evaluated in terms of ω_0/α which is equal, on the other hand, to $2\sqrt{K\tau}$ [from (57)].

It should be noted that, except for the value of the ratio ω_0/α less than 4 or 5, $\omega_1 t_1$, is very close to $\pi/2$.

On the other hand, multiplying (65) by ω_1, squaring (65) and (66) and adding the result, we get, after simplification, and substitution of ω_1 and ω_2 by their values in terms of ω_0 and α:

$$\left[\frac{\dfrac{\pi}{2}-\dfrac{\Omega_d}{K}}{\dfrac{\pi}{2}+\dfrac{\Omega_d}{K}}\right]^2 = e^{-2\alpha t_1} \frac{\left(\dfrac{\omega_0}{\alpha}\right)^2 - 2\sqrt{\left(\dfrac{\omega_0}{\alpha}\right)^2+1}+2}{\left(\dfrac{\omega_0}{\alpha}\right)^2 + 2\sqrt{\left(\dfrac{\omega_0}{\alpha}\right)^2+1}+2}. \quad (68)$$

Letting $x=\omega_0/\alpha$, $y=\Omega_d/K$, and $u=\omega_1 t_1$ (67) and (68) can be written

$$_2\tan \cdot u = -\sqrt{x^4-1} \quad (69a)$$

$$\left[\frac{\dfrac{\pi}{2}-y}{\dfrac{\pi}{2}+y}\right] = e^{-2u/\sqrt{x^2-1}} \frac{x^2-2\sqrt{x^2+1}+2}{x^2+2\sqrt{x^2+1}+2}. \quad (69b)$$

For a given value of x, the corresponding value of u can be found from (69a), and, bringing into (69b) values of u and x, the value of y, ration between the synchronization and catching zone, can be deduced.

If in (69b), the right part of the equation is called R, we get

$$y = \frac{\pi}{2}\,\frac{1-\sqrt{R}}{1+\sqrt{R}}. \quad (70)$$

Fig. 14 gives the curve of Ω_d/K in terms of α/ω_0 for values of α/ω_0 included between 0 and 1.

In assimilating the sinusoidal function to its maximum slope, the ratio Ω_d/K reaches the maximum value of $\pi/2$, and not 1. Hence synchronization range extends to twice $K/2\pi$ and not twice $K/4$ (see the section entitled "Basic Equation of the IGO Circuit").

When x is large with respect to 1, the following simplified relation is found:

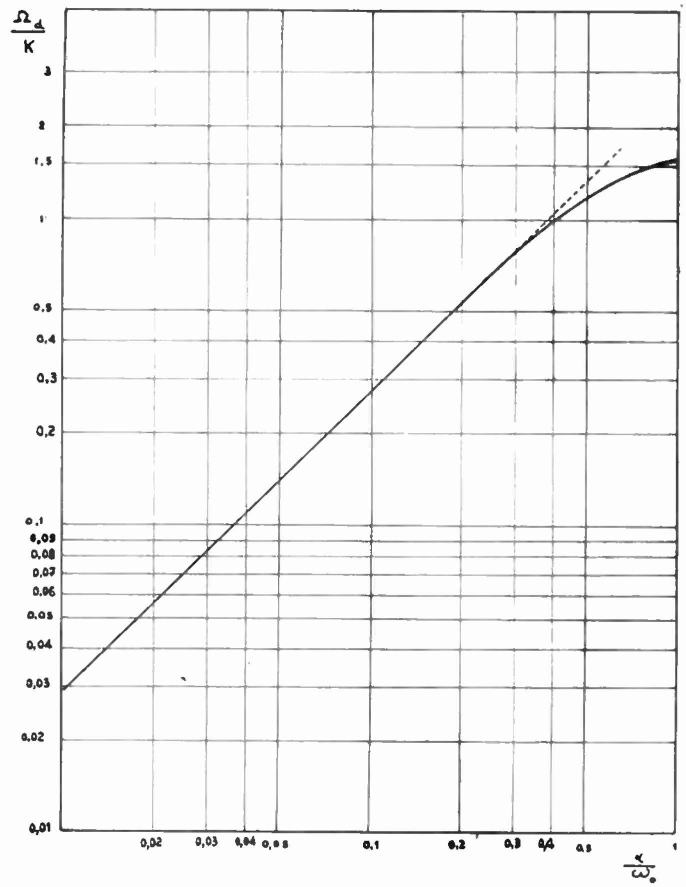

Fig. 14.

$$y = \sim \pi\,\frac{4+\pi}{8x} = \frac{2.81}{x}. \quad (71)$$

Besides the real curve, this line which is an approximate solution, was also drawn.

We see that the value of y given by (71) is still valid for $\alpha/\omega_0 = 0, 6$ at nearly 10 per cent, *i.e.*, for the usual practical values. It should be noted that (71) can also be put in the form:

$$\Omega_d = 1.41\sqrt{\frac{K}{\tau}} = \sim \sqrt{\frac{2K}{\tau}}.$$

A Sideband-Mixing Superheterodyne Receiver*

M. COHN†, ASSOCIATE MEMBER, IRE, AND W. C. KING†, SENIOR MEMBER, IRE

***Summary*—Microwave receivers having bandwidths as much as 22 times greater than the intermediate-frequency amplifier bandwidth have been constructed by generating sidebands on a local oscillator signal and utilizing these sidebands as virtual local oscillators. Both a microwave and a vhf local oscillator signal are injected on a crystal to generate an infinite set of sideband signals separated by the frequency of the vhf oscillator and centered about the microwave oscillator. The low-level received signal mixes with one of these generated virtual local oscillator signals to produce the desired IF signal. The two mixing operations can take place in one crystal or two separate crystals. Measurements have been made of tangential sensitivity and conversion loss and indicate that sensitivities greater than −70 dbm and a continuous bandwidth of 700 mc can be achieved with an intermediate-frequency amplifier having 50 mc bandwidth.**

Introduction

THE DESIRABILITY of having microwave receivers with bandwidths of many hundreds of megacycles and sensitivities approaching those of superheterodyne receivers has long been recognized. One solution to this problem involves the use of cascade-connected traveling-wave tubes in order to obtain the necessary intermediate frequency gain and bandwidth, but with present tubes this system is rather cumbersome and expensive. The moderately high noise figures of the tubes and the wide noise bandwidth of such a system lowers the receiver sensitivity considerably. An alternative to the traveling-wave tube approach is a sideband mixing system which achieves comparable sensitivity and bandwidth by means of an unconventional connection of entirely conventional components.

The system to be described utilizes both a microwave and a vhf local oscillator. These two primary local oscillator signals are injected on a crystal where they cause an infinite set of virtual local oscillator signals to be generated. The generated local oscillator signals are centered about the microwave local oscillator frequency and are separated from each other by the frequency of the vhf local oscillator. The low-level received signal can mix with one of the virtual local oscillator signals to produce the desired IF signal. The mixing of microwave and vhf local oscillators to produce the set of virtual local oscillators and the mixing of virtual local oscillators with signal to produce IF output can be accomplished in the same crystal or in different crystals. These systems are shown in the block diagrams of Fig. 1.

The frequencies produced by the mixing of the two local oscillator outputs are given by:

$$f_n = f_M + nf_{osc}, \quad n = \cdots, -2, -1, 0, 1, 2, \cdots \tag{1}$$

where

f_n = frequency of nth virtual local oscillator,

f_M = frequency of microwave local oscillator,

f_{osc} = frequency of vhf local oscillator.

There will be receiver pass bands located above and below each one of the virtual local oscillators.

Let

f_{IF} = center frequency of IF amplifier.

and

B_{IF} = bandwidth of IF amplifier.

The center frequency of the upper and lower pass bands associated with the nth virtual local oscillator will be denoted by f_n^+ and f_n^- respectively. Then

$$\begin{aligned} f_n^{\pm} &= f_n \pm f_{IF} \\ &= f_M + nf_{osc} \pm f_{IF}. \end{aligned} \tag{3}$$

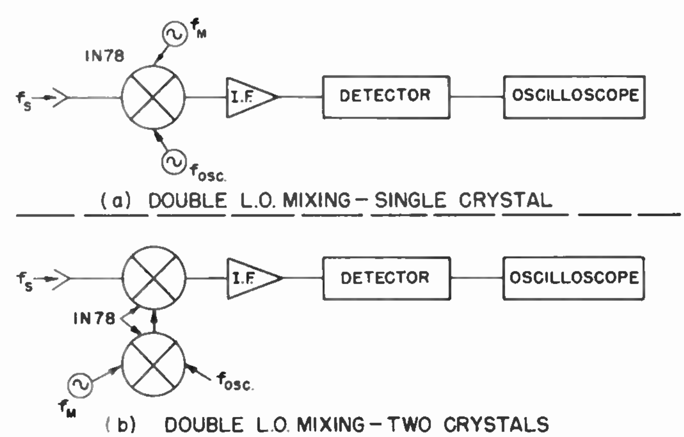

Fig. 1—Sideband mixing systems.

Experimentally it is found that power generated within the mixer at frequency f_n decreases with increasing $|n|$ and as a result conversion loss L_c increases as $|n|$ increases. Thus for a specified receiver sensitivity there is a limiting value N such that $|n| \leq N$. The over-all receiver bandwidth is then given by

$$B = (4N + 2)B_{IF} = mB_{IF},$$

where $m = (4N+2)$ is the multiplicity of the conversion process.

A proper choice of f_{osc}, f_{IF}, and B_{IF} results in a receiver having nearly continuous frequency coverage over bandwidths comparable to those of crystal video detector systems.

* Original manuscript received by the IRE, May 31, 1956; revised manuscript received, August 8, 1956. This research was supported by the United States Air Force through the Office of Scientific Research of the Air Research and Development Command.

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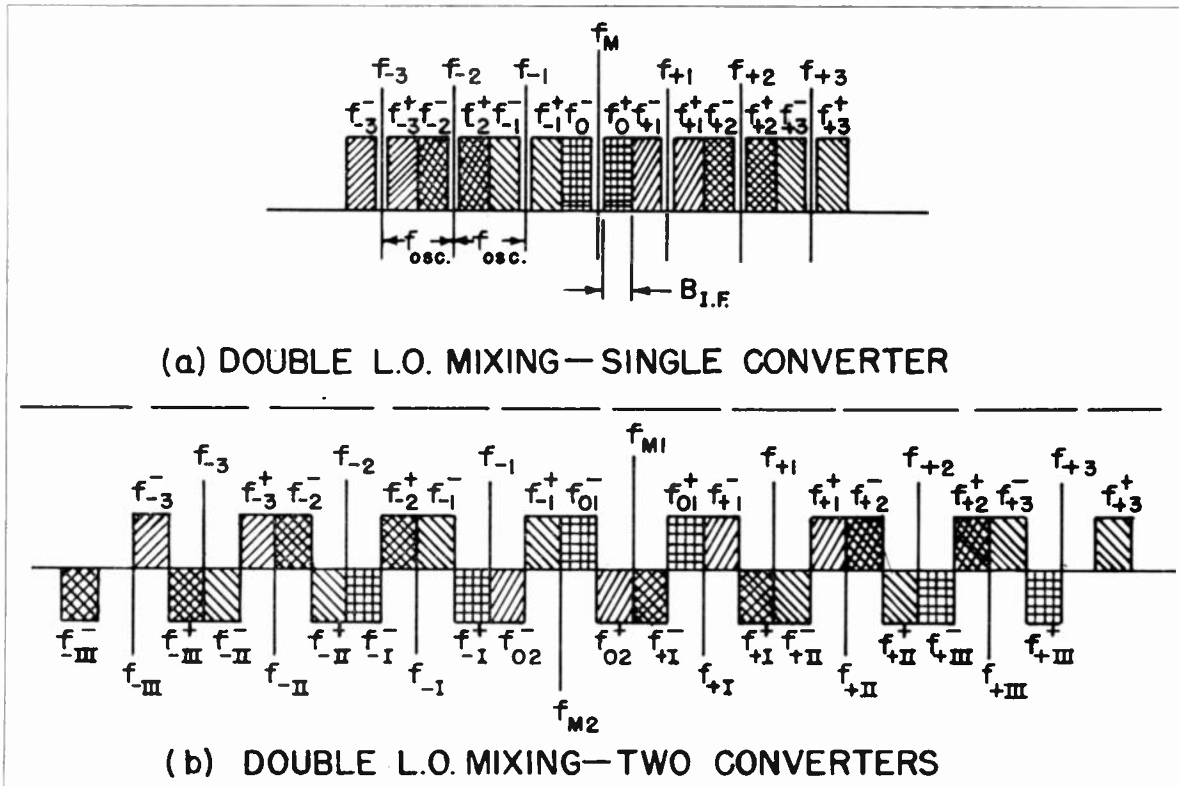

Fig. 2—Possible spectral coverage of sideband-mixing receivers.

Possible Systems

Single Converter

One possible system which illustrates how continuous coverage could be obtained uses an $f_{osc} = 100$ mc and an IF pass band from 5 to 50 mc. See Fig. 2(a). It is apparent that such a system will have holes in its spectrum coverage pattern. These holes can be covered by applying a 10 mc frequency shift to the microwave local oscillator. In this manner a duty ratio of 1.0 can be achieved for 80 per cent of the total bandwidth and 0.5 for the remaining 20 per cent.

Experimental results obtained thus far show that over-all receiver noise figures (referred to the IF bandwidth) of less than 29 db can be obtained in the pass bands associated with the third and all lower order virtual local oscillator sidebands ($n \leq 3$). The above system would, therefore, cover a 700 mc band.

Double Converter

The double converter system consists of two separate multiple mixers of either the single or double crystal type. Each converter has its own microwave local oscillator. The same vhf oscillator is used to feed both multiple mixers, and a single IF amplifier is fed from both converters (Fig. 3). Continuous coverage is obtained with this system if $f_{osc} = 200$ mc and the IF amplifier has a pass band from 50 to 100 mc. The two microwave local oscillators are maintained 100 mc apart by a discriminator control circuit. The double converter system provides interlaced continuous frequency coverage of 1300

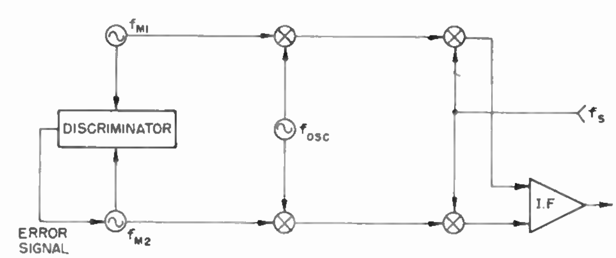

Fig. 3—Double-converter double-crystal system.

mc [Fig. 2(b)]. This system is more complex than th single converter system, but eliminates need to vary mi crowave local oscillator frequency; the one octave I amplifier required by this system is also simpler.

Experimental Results

A number of measurements were made of the tan gential sensitivity and conversion loss of the single an double crystal types of multiple mixers. The measure ments reported here were made using an IF amplifie with a 3 mc pass band; but on the basis of these measurements, the results to be expected from a 50 mc pass band amplifier can be inferred.

A typical result obtained with a single crystal system is shown in Fig. 4. Because of the nearly perfect symmetry of these curves about the frequency of the microwave local oscillator (f_M), only half of the frequency range is plotted. If a horizontal line is drawn midway between the sensitivity and conversion loss curves, it will be seen that the two curves are nearly mirror images about this line. This shows that the deterioration of tangential signal at the outer sidebands is almost

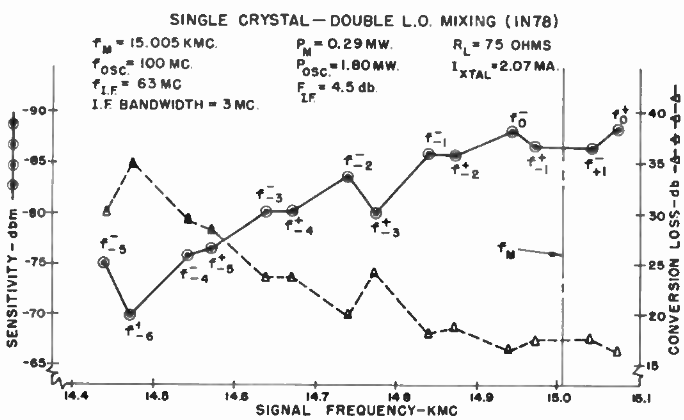

Fig. 4—Measured conversion loss and input signal power required for $s/n=1$ at IF amp output for single crystal system.

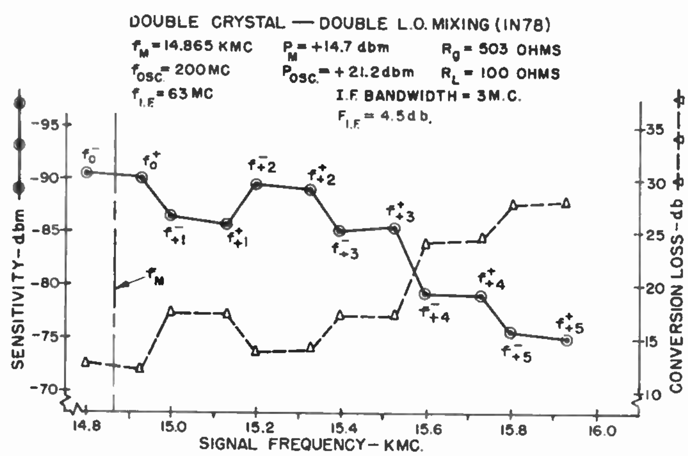

Fig. 5—Measured conversion loss and input signal power required for $s/n=1$ at IF amp output for double crystal system.

entirely accounted for by the increased conversion loss rather than by an increase in noise output.

For signals associated with virtual local oscillators out to the third sideband the tangential signal is less than -77 dbm. Separate measurements made on this system show that, for the particular video system used, tangential video signals represent a signal-to-noise ratio of 3 db at the input to the video detector and therefore the noise level of the IF output is less than -80 dbm out to the third order sidebands. If an IF amplifier with 50 mc bandpass were used to obtain continuous coverage, noise levels of -68 dbm could be expected over an over-all band of 700 mc with a 50 mc video bandwidth. If a video bandwidth of 5 mc is used a noise level of approximately -73 dbm could be expected, a substantial improvement over crystal-video detector systems.

The sensitivities achieved at various sidebands can be substantially altered by varying the power from either of the two primary local oscillators or by varying the dc load resistance on the crystal. The values of these parameters selected for the measurements reported here were those which gave the greatest sensitivities out to the third sideband signals.

A series of similar measurements was made using different values of sideband separation. In all cases the oscillator powers and crystal load resistance were adjusted to give optimum conversion at the third sideband. Results of these measurements are in Table I. The

TABLE I

Tangential Signal and Conversion Loss at Different Sidebands for Various Values of f_{osc}—Single Crystal

Side-bands	$f_{osc}=100$ mc. Tangential Signal (dbm)	$f_{osc}=100$ mc. Conversion Loss (db)	$f_{osc}=200$ mc. Tangential Signal (dbm)	$f_{osc}=200$ mc. Conversion Loss (db)	$f_{osc}=500$ mc. Tangential Signal (dbm)	$f_{osc}=500$ mc. Conversion Loss (db)
f_0	-85.2	16.3	-86.0	16.0	-82.5	18.1
f_1	-83.3	17.6	-79.7	22.5	-77.1	24.2
f_2	-81.8	19.2	-77.1	25.7	-76.7	25.1
f_3	-77.1	23.8	-77.6	24.3	-76.8	24.7
f_4	-74.0	26.5	-61.5	40.6		
f_5	-72.8	28.9	-68.8	32.7		

system sensitivity at the optimized side-band is practically independent of oscillator frequency; hence system design is flexible with respect to this parameter.

Fig. 5 shows the results of measurements of tangential signal and conversion loss for the two-crystal system. In the two-crystal system higher primary local oscillator power levels are impressed on the sideband generating crystal. In this way more powerful virtual local oscillator signals are generated, which cause more efficient mixing to take place in the second crystal. The same symmetry statements as were made for the single crystal mixer also apply to the double crystal case.

The tangential signal is less than -82 dbm in all pass bands out to the third order sidebands. This corresponds to a noise threshold of less than -85 dbm. The conversion loss is less than 18 db at the third sideband. If a 50 mc wide IF amplifier were used with this two-crystal double local oscillator mixer, we could expect a noise level of -73 dbm over a 700 mc band with a video bandwidth of 50 mc.

In the two-crystal system, it is possible to make direct measurements of the virtual local oscillator power developed in the first crystal. The results for two settings of microwave oscillator power P_M, vhf oscillator power P_{osc}, and crystal load resistance R_g are in Table II, on the next page. At each of the settings of R_g, the values of P_M and P_{osc} were adjusted to optimize simultaneously the first three virtual local oscillator sidebands. It was experimentally observed that higher values of R_g increased the amount of power developed in the even order virtual local oscillators at the expense of the odd orders. A value of $R_g=500$ appears to be a good compromise value.

The values in Table II show that it is possible to generate sufficient sideband power to make efficient mixing at the second crystal possible. Even greater nth order virtual local oscillator sideband power P_n could be obtained by using more primary local oscillator power (P_M and P_{osc}). There is no need to do so however, since we have already exceeded the power required for

TABLE II
VIRTUAL LO SIDEBAND POWER DEVELOPED IN CRYSTAL No. 1 OF TWO CRYSTAL DOUBLE LO MIXER

$P_M=+19$ dbm, $P_{osc}=+22$ dbm, $R_g=1000$ ohms		$P_M=+19$ dbm, $P_{osc}=22.6$ dbm, $R_g=500$ ohms	
Sideband No. n	Power Developed in Sidebands P_n (dbm)	Sideband No. n	Power Developed in Sidebands P_n (dbm)
0	+14.0	0	+11.6
1	+ 1.4	1	− 0.6
2	+ 3.6	2	+ 0.1
3	−10.1	3	− 4.4
4	− 6.4	4	− 8.2
5	−18.3	5	−22.9

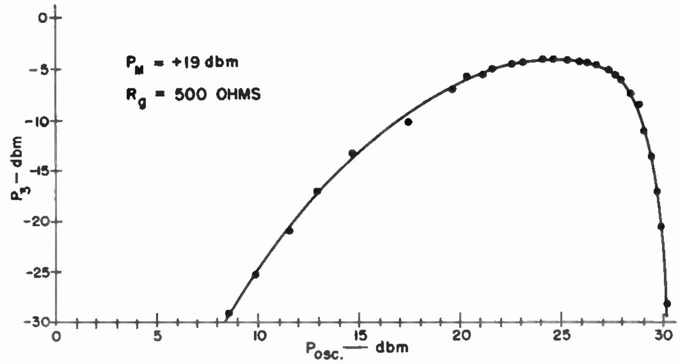

Fig. 6—Third local oscillator sideband power (P_3) developed in crystal No. 1 of two crystal double local oscillator mixer vs P_{osc}.

efficient mixing at the fundamental and low order sidebands. The need is to redistribute the generated virtual local oscillator power so that P_n is more nearly constant for increasing n.

Fig. 6 is a curve of third sideband virtual local oscillator power (P_3) vs P_{osc}. The broad maximum of the curve shows that the setting of P_{osc} is not critical. For larger values of P_M, P_3 would reach a higher value and peak at a greater setting of P_{osc}. A more promising means of obtaining the required P_n's for a range of n is to use a nonsinusoidal waveform for P_{osc}. In this way it may be possible to obtain the required P_n vs n distribution without subjecting the second crystal to excessive total power. If the required distribution of virtual local oscillator power can be obtained, then any excess power makes it possible to inject this generated local oscillator power to the second crystal via the auxiliary arm of a directional coupler. In this way the signal can be injected at the main arm without coupling losses.

A curve of P_3 vs primary microwave local oscillator power P_M is shown in Fig. 7. For lower values of P_{osc} this curve reaches a peak and then decreases in much the same manner as the curve of Fig. 6.

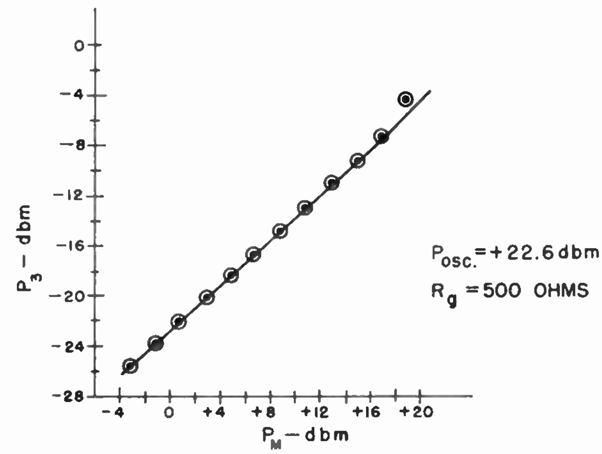

Fig. 7—Third local oscillator sideband power (P_3) developed in crystal No. 1 of two crystal double local oscillator mixer vs P_m.

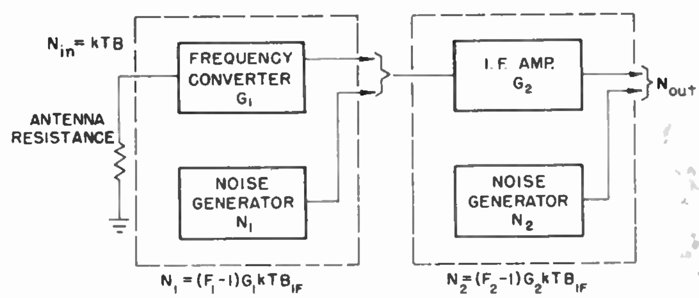

Fig. 8—Equivalent network for superheterodyne receiver.

DISCUSSION

Any theoretical discussion of noise figure and sensitivity of a multiple mixing system is complicated by the fact that bandwidth is not the same in all parts of the system. Consequently the concept of noise figure becomes somewhat ambiguous. The separate roles of rf and IF bandwidths become clear, however, if each of the two noisy linear networks involved is resolved into an equivalent combination consisting of a noiseless linear network plus a noise generator. If this procedure is applied to both the crystal mixer and its associated IF amplifier the system is as shown in Fig. 8.

Note that in a superheterodyne system the output noise power from a frequency converter consists of two contributions: rf noise present at the input to the mixer which is converted into the IF pass band and internally generated noise at intermediate frequency. The latter does not depend upon the former for small signal input and is independent of the bandwidth on the rf side of the converter. Thus the excess noise power output of the converter is calculated using the IF bandwidth, and in the notation of Fig. 8,

$$N_1 = kTB_{IF}(F_1 - 1)G_1. \tag{5}$$

However, the input noise to the converter is a function of both rf and IF bandwidths,

$$N_{in} = kTmB_{IF}.$$

The total noise output and equivalent noise input are

$$N_{out} = N_{in}G_1G_2 + N_1G_2 + N_2$$

$$(N_{in})\text{equiv} = N_{in} + \frac{N_1}{G_1} + \frac{N_2}{G_1G_2}$$

$$= kTB_{IF}\left[m + (F_1 - 1) + \frac{F_2 - 1}{G_1}\right]$$

$$= kTB_{IF}[L_c(t + F_2 - 1) + (m - 1)]$$

where

$$t = F_1G_1 \quad \text{and} \quad L_c = 1/G_1.$$

The value of the conversion loss (L_c) used above is not the crystal manufacturers value but a value to be determined by measurement in a sideband mixer. The conversion loss is a function of the LO sideband number (n) as shown in Fig. 4 and 5. Experimental results show that the crystal noise temperature (t) is essentially constant for all sidebands and insignificantly different from the manufacturer's values.

The over-all noise figure of the receiver when expressed in terms of the IF amplifier bandwidth is

$$F(m) = \frac{(N_{in})\text{equiv}}{kTB_{IF}} = L_c(t + F_2 - 1) + (m - 1). \quad (6)$$

Referred to the rf bandwidth, the noise figure is

$$F'(m) = \frac{(N_{in})\text{equiv}}{kTmB_{IF}} = \frac{L_c}{m}(t + F_2 - 1) + 1 - \frac{1}{m}. \quad (7)$$

Eqs. (6) and (7) reduce to the usual expression for superheterodyne noise figure when the rf and IF bandwidths are equal ($m=1$), as of course they must. Eq. (7) indicates that receivers having noise figures arbitrarily close to unity might be realized by using arbitrarily large m provided that L_c does not increase at as great a rate.

Even when the greatest conversion loss over a band is used in calculating the expected sensitivity the multiple-mixer is comparable to regular mixing. For example, a conventional superheterodyne having $L_c = 6.5$ db, $F_{IF} = 6.5$ db, $t = 2.5$, and a bandwidth of 700 mc would have a sensitivity of -71 dbm; the measured data for multiple-mixing indicate sensitivities of -68 dbm and -73 dbm should be attainable with single- and double-crystal mixing respectively.

The important fact about the multiple-mixing technique is, of course, the bandwidth magnification obtainable. It has been shown experimentally that conversion loss and noise figure are substantially independent of sideband spacing. Hence very wide IF amplifiers (twt's) can be utilized to achieve even wider receivers.

Conclusion

It has been shown both theoretically and experimentally that receiver rf bandwidths of 700 mc and sensitivities in excess of -70 dbm should be achievable with reasonably well-designed 50 mc IF amplifiers. Because of its inherent flexibility the system can be used to magnify the bandwidth of receivers using IF strips of much greater bandwidth. The reduction in receiver sensitivity accompanying such magnification is certainly no greater than that which would result in a conventional superheterodyne system and can probably be made considerably less. Thus the multiple-mixer has sensitivity comparable to a superheterodyne and bandwidth comparable to a crystal-video receiver.

On the basis of these measurements, supported by analysis, it appears that the multiple-mixing technique has a definite role to fulfill whenever rf coverage requirements dictate receiver bandwidths much greater than the spectral width of the expected signals.

Frequency-Temperature-Angle Characteristics of *AT*-Type Resonators Made of Natural and Synthetic Quartz*

RUDOLF BECHMANN, SENIOR MEMBER, IRE†

Summary—**Investigations into the frequency-temperature behavior of *AT*-type quartz resonators have revealed differences between natural and synthetic quartz. The differences refer mainly to a shift of the optimum angle of orientation by a few minutes of arc and to a slight change of the frequency-temperature characteristic itself. To describe the frequency-temperature behavior analytically, the measured change of frequency vs temperature can be developed in a power series, determined by first, second, and third-order temperature coefficients. In the temperature range from −60 to +100°C. higher-order temperature coefficients can be neglected. For a large number of *AT*-type resonators of various angles made from natural and several kinds of synthetic quartz, the temperature coefficients, and their variation with the angle have been determined. It is possible to modify the properties of synthetic quartz by introducing other elements during the growing process. An example is quartz grown in an alkaline solution containing germanium dioxide. Measurements have been made on *AT*-type resonators cut from such synthetic quartz. The third-order temperature coefficient for the *AT*-type resonator is found noticeably reduced; the frequency-temperature curves are flattened over a wider temperature range.**

Introduction

Natural quartz from different sources has displayed a remarkable uniformity as far as all piezoelectric applications are concerned. Regardless of the source of electronic grade natural quartz used, when the orientation of the piezoelectric resonator plates is specified, no significant variations are observed in the performance of the resulting resonator plates. Electronic grade quartz is defined as quartz which contains no defects such as optical and electrical twinning, cracks, solid inclusions, veils, bubbles, needles, and ghosts or phantoms.

Within the last few years, quartz crystals have been grown artificially by a hydrothermal process in the laboratory and pilot plant.[1] Resonator blanks of any usual shape and size can be produced from synthetic quartz.

Since resonators, in particular *AT*-type resonators, made from synthetic quartz have been investigated with respect to the frequency-temperature behavior, it has been observed in various laboratories[2] that differences in this characteristic performance exist. The differences refer mainly to a shift of the optimum angle of orientation by a few minutes of arc, and to a slight change of the frequency-temperature function itself.

It has also been found that synthetic quartz crystals grown from different seed types and grown under different temperature and pressure conditions, show slight changes in the frequency-temperature characteristics. However, quartz grown under the same conditions shows again considerable uniformity and reproducibility of the frequency-temperature characteristics and other physical properties.

The present sources of synthetic quartz are:

1) Pilot Plant, Bedford, Ohio, of the Clevite Research Center, formerly The Brush Laboratories Company, Cleveland. Growth condition:
   a) pressure about 5000 psi, temperature about 350°C., solvent: 2 molar sodium carbonate solution, using *CT* plates as seeds.
   b) pressure 8000 psi, temperature 350°C., solvent: 0.83 molar sodium carbonate solution, using *Y* bars as seeds.
2) The Clevite Research Center, Cleveland, growing quartz under modified conditions: pressure about 1500 psi, temperature under 300°C., using different seed types.
3) Bell Telephone Laboratories, Murray Hill, N. J., growing synthetic quartz under high pressure, about 15,000 to 20,000 psi; temperature about 380°C., solvent: sodium hydroxide solution, using *CT* and *Z* plates as seeds.
4) Research Laboratories, The General Electric Co., Ltd., Wembley, Middlesex, England, growing synthetic quartz under high pressure conditions, using *Z* plates as seeds.

Some details referring to Brush synthetic quartz can be found in Bechmann and Hale.[1] The technology of the growth of synthetic quartz will not be discussed.

I. The Frequency-Temperature Behavior of Quartz Resonators

A. General

The so-called zero temperature coefficient cuts can be divided into two main groups:

1) The zero temperature coefficient in the first approximation depends on the orientation of the specimen only and is independent of the dimensions. Examples: *AT*, *BT*, *CT*, *DT* cuts.

* Original manuscript received by the IRE, December 8, 1955; revised manuscript received, July 30, 1956.

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[1] R. Bechmann and D. R. Hale, "Electronic grade synthetic quartz," *Brush Strokes* (Brush Electronics Company, Cleveland, Ohio), vol. 4, pp. 1–7; September, 1955.

[2] Bell Telephone Labs., U. S. Signal Corps Engrg. Labs. and Industries, Bliley Electric Co., The James Knights Co., Standard Piezo Co., etc.

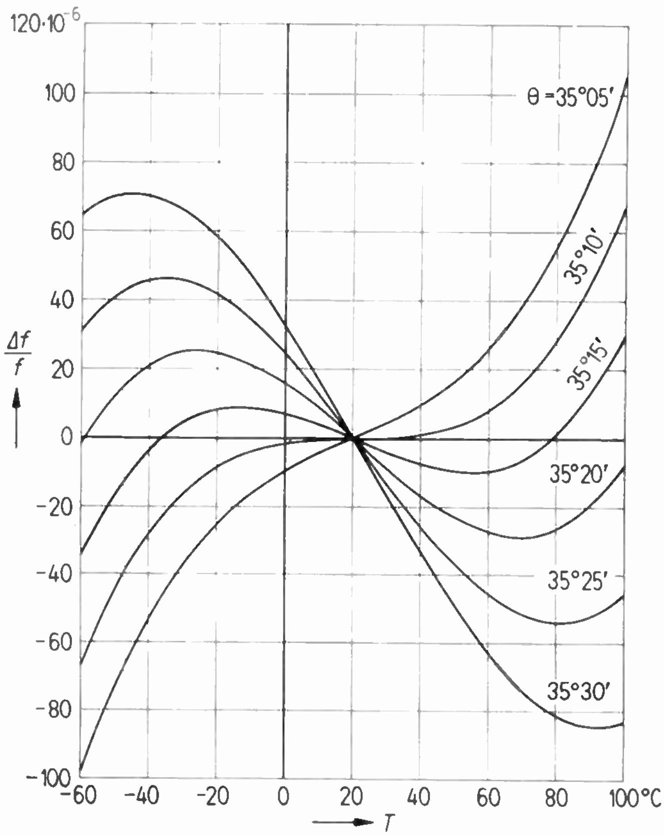

Fig. 1—Frequency-temperature-angle characteristics of *AT*-type quartz resonators, fundamental mode frequency 7 mc, crystal plated.

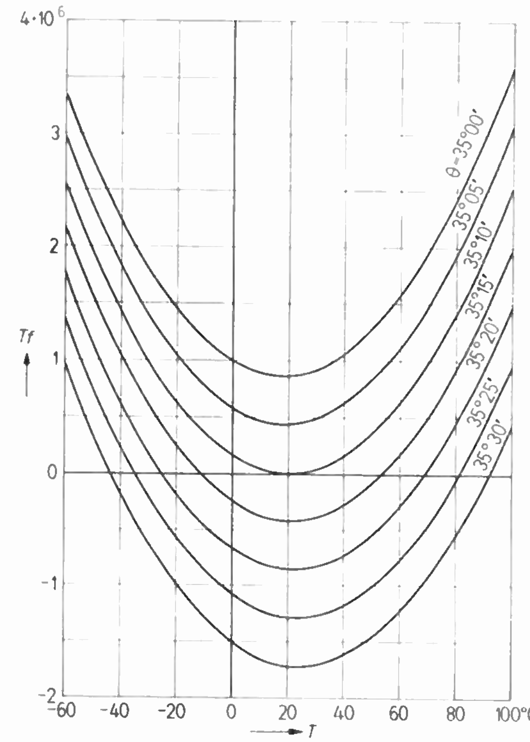

Fig. 2—Temperature coefficient of *AT*-type quartz resonators.

2) The zero temperature coefficient is obtained by choice of dimensions and is caused by coupled modes. Example: *GT* cut.

For more thorough consideration the frequency-temperature behavior of a piezoelectric resonator depends on the following parameters:

1) Orientation—angles of cut.
2) Ratio of dimensions; for example for thickness modes, length or diameter to thickness ratio.
3) Order of overtone.
4) Shape of plate.
5) Type of mounting.

B. Typical Frequency-Temperature-Angle Characteristics of AT- and BT-Type Quartz Resonators

The frequency-temperature-angle characteristics for *AT*-type resonators made from natural quartz, fundamental mode, frequency about 7 mc, in the temperature range −60 to +100°C. are in Fig. 1 above. The value $\Delta f/f$ refers to the relative frequency change as function of the temperature, T. The temperature coefficient of frequency, Tf, as a function of angle of orientation and temperature, follows from the curves in Fig. 1 and is shown in Fig. 2.

The frequency-temperature-angle characteristics of *BT*-type resonators made from natural quartz, fundamental mode, frequency about 2 mc, in the temperature range −60 to +100°C. are in Fig. 3 above. The resonators were measured in a holder with a small air gap. The temperature coefficients of frequency, Tf, as function of angle of orientation and temperature, following

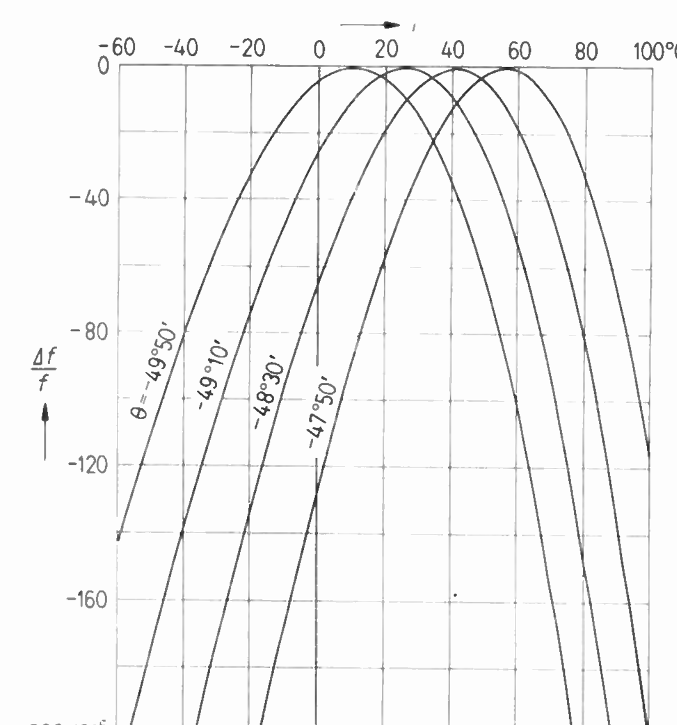

Fig. 3—Frequency-temperature-angle characteristics of *BT*-type quartz resonators, fundamental mode frequency 2 mc., small air gap.

from the curves in Fig. 3, are shown in Fig. 4. The change of the temperature T_{max}, giving the maximum of the frequency-temperature characteristics as function of the angle of orientation, θ, is shown in Fig. 5.

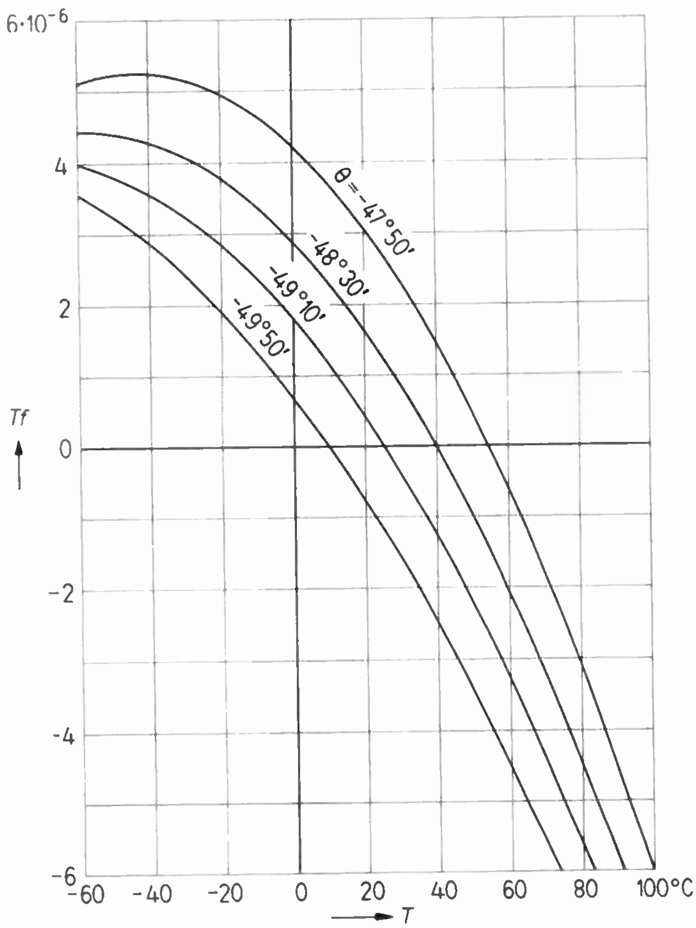

Fig. 4—Temperature coefficient of BT-type quartz resonators.

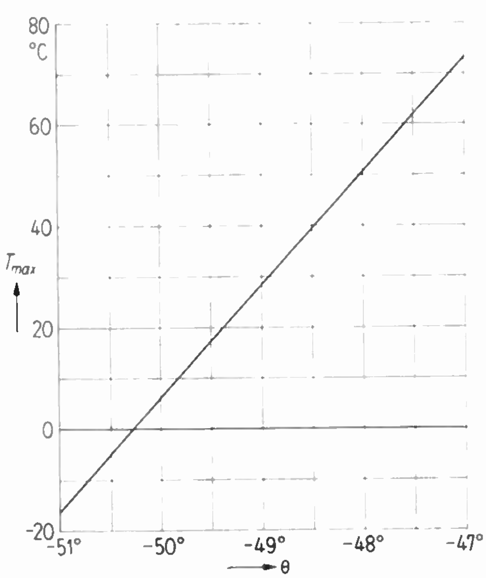

Fig. 5—The maximum temperature of the frequency-temperature-angle characteristics of BT-type quartz resonators as function of the angle of orientation.

C. Analytical Expressions for the Frequency-Temperature-Angle Characteristics

For the practical application as well as for theoretical consideration, it is necessary to define the frequency-temperature-angle characteristics quantitatively, by introducing some temperature coefficients of higher order.[3,4] The measured frequency, f, of a crystal unit as a function of the temperature, T, can be developed in a power series in the vicinity of the frequency, f_0, at the arbitrary temperature T_0

$$\frac{f - f_0}{f_0} = \frac{\Delta f}{f_0} = a_0(\theta)[T - T_0] + b_0(\theta)[T - T_0]^2 + c_0(\theta)[T - T_0]^3 + \cdots \tag{1}$$

where $a_0(\theta)$, $b_0(\theta)$, and $c_0(\theta)$ are the first, second, and third-order temperature coefficients of frequency as defined by

$$a_0(\theta) = \frac{1}{f_0}\left(\frac{\partial f}{\partial T}\right), \qquad b_0(\theta) = \frac{1}{2f_0}\left(\frac{\partial^2 f}{\partial T^2}\right),$$
$$c_0(\theta) = \frac{1}{6f_0}\left(\frac{\partial^3 f}{\partial T^3}\right)_0. \tag{2}$$

These constants are functions of the orientation and the other parameters mentioned in Section I*A*. The temperature coefficient of the frequency is given by

$$Tf = \frac{1}{f_0}\frac{\partial f}{\partial T} = a_0(\theta) + 2b_0(\theta)[T - T_0] + 3c_0(\theta)[T - T_0]^2. \tag{3}$$

For the temperature range -60 to $+100$°C. usually considered, temperature coefficients of higher order than three can be neglected. These three temperature coefficients can be related to the corresponding coefficients of the elastic constants involved and the coefficients of expansion.

The frequency-temperature-angle characteristics are given by the following expressions, assuming the change of the three temperature coefficients with angle of orientation to be linear,

$$\frac{\Delta f}{f} = a_0(\theta_0)[T - T_0] + b_0(\theta_0)[T - T_0]^2 + c_0(\theta_0)[T - T_0]^3$$
$$+ \left\{\frac{\partial a_0(\theta)}{\partial \theta}[T - T_0] + \frac{\partial b_0(\theta)}{\partial \theta}[T - T_0]^2 + \frac{\partial c_0(\theta)}{\partial \theta}[T - T_0]^3\right\}(\theta - \theta_0) \tag{4}$$

where $\partial a_0(\theta)/\partial\theta$, $\partial b_0(\theta)/\partial\theta$, and $\partial c_0(\theta)/\partial\theta$ are the derivatives with respect to the angle of the three temperature

[3] W. P. Mason, "Piezoelectric Crystals and Their Application to Ultrasonics," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.

[4] R. Bechmann, "The frequency-temperature behavior of piezoelectric resonators made of natural and synthetic quartz," 1955 IRE CONVENTION RECORD, vol. 3, part 9, pp. 56–61.

coefficients. For the range considered of about 1°, the linear terms are sufficient; considering a wider range for the orientation, higher terms for the derivatives of the temperature coefficients must be introduced.

In the vicinity of a zero angle of orientation for the frequency, when a_0 is zero or very small, two types of frequency-temperature behavior may be distinguished.

1) In case where b_0 is rather small and c_0 large, the frequency-temperature characteristic has a cubic form—an example is the AT-cut where generally b is smaller than $5\cdot 10^{-9}/(°C)^2$ and c is in the order of $100\cdot 10^{-12}/(°C)^3$. Another example is the GT cut where both the second and third-order temperature coefficients are very small.
2) In most of the other cuts, the second-order temperature coefficient is predominant, giving a parabolic frequency-temperature characteristic.

Considering first the frequency-temperature characteristics of an AT-type crystal, a typical frequency-temperature curve for an angle of orientation, having a small negative value for the first order temperature coefficient of frequency, is shown in Fig. 6. The charac-

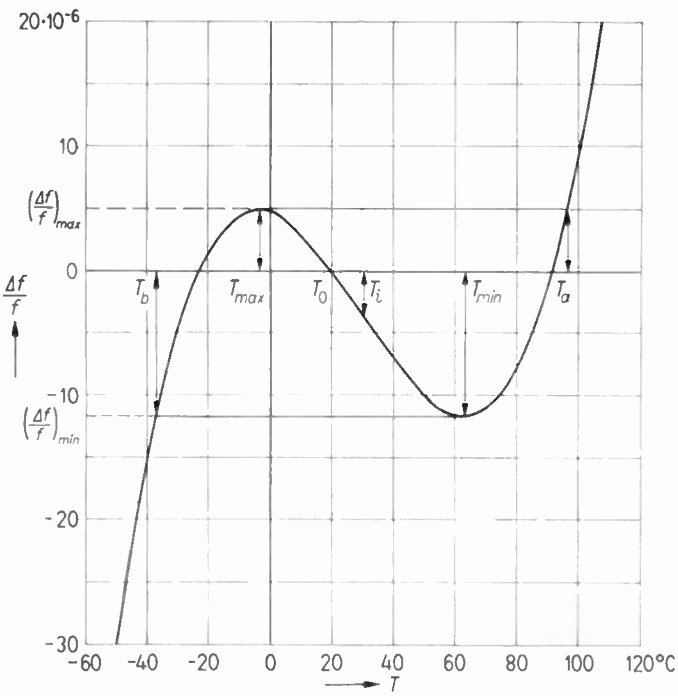

Fig. 6—Typical frequency-temperature characteristic of an AT-type quartz resonator.

teristic quantities determining the frequency-temperature behavior are: Maximum and minimum temperature, ($T_{\max}$, $T_{\min}$), the corresponding maximum and minimum frequency change ($\Delta f/f_{\max}$, $\Delta f/f_{\min}$); the inflection temperature, T_i, that is the temperature for which the derivative of the temperature coefficient of frequency becomes zero; further the temperatures T_a and T_b, where in the case considered

$$T_a - T_b = 2(T_{\min} - T_{\max}).$$

The analytical expressions for $T_{\min}$ and $T_{\max}$ follow from $Tf=0$:

$$T_{\min,\max} - T_0 = \frac{-b_0 \pm \sqrt{b_0^2 + 3a_0c_0}}{3c_0}. \tag{5}$$

The corresponding frequency deviation is given by

$$\left(\frac{\Delta f}{f}\right)_{\max,\min} = \frac{\pm 2\sqrt{b_0^2 - 3a_0c_0}^{\,3} + 2b_0^3 - 9a_0b_0c_0}{27c_0^2}. \tag{6}$$

The inflection temperature T_i is defined by

$$\frac{\partial Tf}{\partial T} = \frac{\partial^2 f}{\partial T^2} = 0,$$

hence

$$T_i - T_0 = -\frac{b_0}{3c_0}. \tag{7}$$

Introducing the inflection temperature T_i as reference temperature instead of T_0, where $T_i - T_0 = -b_0/3c_0$, then $b_i = 0$ and (1) simplifies to

$$\frac{f-f_i}{f_i} = \frac{\Delta f}{f} = a_i(T - T_i) + c_i(T - T_i)^3 \tag{8}$$

where

$$a_i = \frac{1}{p}\,\frac{3a_0c_0 - b_0^2}{3c_0},$$
$$c_i = \frac{1}{p}\,c_0, \tag{9}$$
$$p = 1 + \frac{2b_0^3 - 9a_0b_0c_0}{27c_0^2}.$$

Eqs. (5) and (6) then become

$$T_{\min,\max} - T_i = \pm\sqrt{\frac{-a_i}{3c_i}} \tag{10}$$

and

$$\left(\frac{\Delta f}{f}\right)_{\max,\min} = \pm 2\sqrt{\frac{-a_i^3}{27c_i}}. \tag{11}$$

The above formulas describe the frequency-temperature behavior of an AT cut in the temperature range considered with sufficient accuracy.

In case of the BT-type quartz resonator and similar cuts with a parabolic frequency-temperature characteristic, the relations given above hold. In the temperature range considered, only one maximum of the frequency occurs. Instead of the inflection temperature, T_i, the temperature $T_{\max}$ at which the frequency maximum occurs is the significant temperature.

The frequency-temperature-angle characteristic related to its maximum is given by

$$\frac{\Delta f}{f} = b_m(\theta)[T - T_{\max}]^2 + c_m(\theta)[T - T_{\max}]^3. \tag{12}$$

It follows

$$a_m = \frac{1}{p}\left[a_0 + 2b_0(T_{\max} - T_0) + 3c_0(T_{\max} - T_0)^2\right] = 0 \quad (13)$$

$$b_m = \sqrt{b_0^2 - 3a_0c_0} \qquad c_m = c_0$$

further from

$$T_{\max} - T_0 = \frac{b_m - b_0}{3c_0} = -\frac{a_0}{b_0 + b_m} \approx -\frac{a_0}{2b_0}. \quad (14)$$

$T_{\max}$ as function of the angle of orientation of the plate can be obtained from the derivatives of the temperature coefficients $a_0(\theta)$, $b_0(\theta)$, and $c_0(\theta)$ or from $b_m(\theta)$ and $c_m(\theta)$ with respect to the angle of orientation.

II. The Frequency-Temperature Behavior of *AT*-Type Resonators Made From Natural and Synthetic Quartz—Experimental Studies

It was necessary to make initial studies on natural quartz since no frequency-temperature data for *AT*-type resonators were available which were sufficiently accurate for analyses. Frequency-temperature-angle characteristics found in the literature were analyzed, and serious discrepancies were found. Later reliable, unpublished frequency-temperature characteristics of natural quartz, as measured by Bell Telephone Laboratories, were obtained through the courtesy of the U. S. Signal Corps Laboratories.

A. Technique of Measurements

The investigations at the Brush Laboratories Company have been carried out using *AT*-type quartz resonators made from plates of a diameter of about 17 to 25 mm and a thickness of about 0.7 to 1 mm in the frequency range of about 1600 to 2500 kc for the fundamental mode.

It is well known that thickness shear mode plates of any form with square edges often have series of unwanted frequencies instead of the single frequency response. Frequency and resonance resistance show anomalous behavior and are both very sensitive to slight change of form of the plate and to changes of temperature. These effects usually vanish when spurious modes are absent from the immediate neighborhood of the resonance frequency. By bevelling disks of a certain diameter-thickness ratio, spurious modes can be removed and a single response can be obtained.[5] To detect spurious resonances, an automatic recorder is very useful. When there is one clear response, the series resonance resistance of the crystal is almost independent of temperature changes over a wide range. This property forms another good practical criterion for the quality of the crystal.

[5] R. Bechmann, "Single response thickness-shear mode resonators using circular bevelled plates," *J. Sci. Instr.*, vol. 29, pp. 73–76; March 1952.

The plates were mounted in a three-point ceramic holder with electrodes having a small separation (air gap) from the surfaces of the plate. The plates were excited in an oscillator operating at series resonance frequency, and the frequency was measured with a Berkeley Frequency Counter. Frequency measurements were carried out in the temperature range −60 to +100°C.

While these investigations were in progress, measurements from other laboratories came to our knowledge. These measurements were made usually on *AT*-type crystal cuts prepared according to specifications CR-18/U and CR-23/U. These frequency-temperature-angle characteristics were analyzed and the results are discussed in Part B.

B. Survey of the Results—The Observed Values of the First-, Second-, and Third-Order Temperature Coefficients for *AT*-Type Quartz Resonators

The following quartz material has been investigated with respect to the properties of *AT*-type resonators:

1) Electronic grade natural quartz.
2) Synthetic quartz grown on rectangular seed plates nearly parallel to the minor rhombohedral face (*AT* or *CT* cuts), length extension in the direction of Z' axis, at Brush Pilot Plant.
3) Synthetic quartz grown on small seed bars with their length extension in the direction of the Y-axis, also at Brush Pilot Plant.
4) Synthetic quartz grown on rectangular seed plates parallel to the minor rhombohedral face, length extension in the direction of Z' axis, under special conditions (low pressure), at Brush Laboratories Company.
5) Synthetic quartz grown on rectangular seed plates parallel to the minor rhombohedral face, at Bell Telephone Laboratories.
6) Synthetic quartz grown on seed plates orientated perpendicular to the Z axis, length extension parallel to the Y axis, at General Electric Company, Wembley, England.

Further

7) Synthetic quartz grown on rectangular seed plates nearly parallel to the minor rhombohedral face (*AT* or *CT* cuts) with germanium addition in NaOH solution, at Brush Laboratories Company.
8) Synthetic quartz grown on rectangular seed plates parallel to the minor rhombohedral face with germanium addition in Na_2CO_3 solution, also at Brush Laboratories Company.

Table I gives a survey of the constants determining the frequency-temperature-angle characteristics of *AT*-type resonators made from the various quartz materials mentioned above. This table lists the various types of

TABLE I

FIRST-, SECOND-, AND THIRD-ORDER TEMPERATURE COEFFICIENTS OF FREQUENCY AND THEIR DERIVATIVES WITH THE ANGLE OF ORIENTATION OF AT-TYPE QUARTZ RESONATORS MADE FROM NATURAL AND SYNTHETIC QUARTZ

Quartz Type	Electrode Arrangement	Diameter Thickness Ratio	Order of Mode n	$a_0(\theta_x)=0$ θ_x	$b_0(\theta_x)$ $10^{-9}/(°C)^2$	$c_0(\theta_x)$ $10^{-12}/(°C)^3$	$\frac{\partial a_0}{\partial\theta}$ $10^{-6}/°\theta$	$\frac{\partial b_0}{\partial\theta}$ $10^{-9}/°\theta$	$\frac{\partial c_0}{\partial\theta}$ $10^{-12}/°\theta$	$b_0(\theta_y)=0$ θ_y	T_0 °C	(T_i-T_0) for θ_x	References for Measurements
Natural	Air Gap	20–25	1	35°10′	−1.3	110	−5.15	−4.5	−10	34°53′	25	4.0	B.L.
			3	35°18′	−1.4	100	−5.15	−4.5	−10	34°59′		4.5	
			5	35°20′	−1.1	95	−5.15	−4.5	−10	35°05′		4.0	
	Air Gap	25–40	1	35°11′	−1.4	110	−5.15	−4.5	−20	34°56′	25	4.0	B.L.
			3	35°18′	−1.7	105	−5.15	−4.5	0	34°55′		5.5	
	Plated	~50	1	35°10′	−0.1	130	−5.15	−4.5	−20	35°09′	20	0.5	St.P.
			3	35°18′	−1.7	105	−5.15	−4.5	0	34°55′		5.5	
	Plated		5	35°22′	−1.2	105	−5.5	−4.5	0	35°06′	20	4.0	B.T.L.
	Air Gap	5.8	1	34°28′	0.1	75	−5.15	−4.5	0	34°29′	25	−0.5	B.L.
Synthetic Minor Rhombohedral Seed Type Brush	Air Gap	20–25	1	35°15′	−2.3	115	−5.15	−4.5	−10	34°44′	25	7.0	B.L.
			3	35°24′	−4.0	105	−5.15	−4.5	−10	34°31′		12.5	
			5	35°25′	−4.3	110	−5.15	−4.5	−10	34°30′		13.0	
	Plated	~50	1	35°16′	−3.0	120	−5.15		−10	34°36′	20	9.0	St.P.
			3	35°24′	−3.3	115	−5.15	−4.5	−10	34°40′		9.5	
	Plated	~40	5	35°26′	1.7	105	−5.25	−4.5	0	36°00′	45	−7.5	S.C.E.L.
Synthetic Y Bar Seed Type Brush	Air Gap	~20	1	35°13′							25		B.L.
			3	35°20′		80							
			5	35°20′		75							
	Plated	~40	5	35°22′	1.4	100	−5.05	−5.3	0	35°38′	30	−5.0	S.C.E.L.
Synthetic Minor Rhombohedral Seed Type Low Pressure—Brush	Air Gap	~25	1	35°18′	−4.0	120	−5.15	−4.5	0	34°25′	25	11.0	B.L.
			3	35°29′	−4.9	125	−5.15	−4.5	0	34°24′		13.0	
Synthetic Minor Rhombohedral Seed Type High Pressure Bell	Air Gap	18–25	1	35°17′	−3.2	105	−5.15	−4.5	−10	34°34′	25	10.0	B.L.
		19–26	3	35°25′	−3.6	110	−5.15	−4.5	−10	34°36′		11.0	
	Plated	~40	5	35°29′	2.6	125	−5.15	−4.5	0	36°04′	45	−7.0	S.C.E.L.
Synthetic Z Seed Type General Electric Co., Wembley, England	Plated		1	35°12′	−1.5	118	−5.15	−4.5	0	34°52′	20	4.0	G.P.O.
			3	35°21′	−1.4	72				35°02′		6.5	
			5	35°21′	−1.7	69				34°58′		8.0	
			7	35°21′	−2.6	84				34°46′		10.0	
Synthetic Ge Addition NaOH Solution Brush	Air Gap	20–25	1	35°29′	−2.2	65	−5.15	−4.5	0	35°00′	25	11.5	B.L.
			3	35°36′	−2.3	105	−5.15	−4.5	−10	35°01′		7.5	
			5	35°36′	−2.3	95	−5.15	−4.5	−10	35°02′		8.0	
Synthetic Ge Addition Na_2CO_3 Solution Brush	Air Gap	19–23	1	35°13′	−3.1	100	−5.15	−4.5	0	34°32′	25	10.0	B.L.
			3	35°22′	−3.5	100	−5.15	−4.5	0	34°35′		11.5	

quartz material investigated; the electrode arrangement used for the excitation of the plates, that is, air gap holder or plated crystal surfaces; the approximate diameter-thickness ratio of the plate; the order of mode, where $n=1$ is the fundamental mode, $n=3$, 5, 7 are the third, fifth, and seventh overtones respectively. The column headed $a_0(\theta_x)=0$ gives the angle of orientation, θ_x, for which the first order temperature coefficient of frequency, $a_0(\theta_x)$, becomes zero. The columns headed $b_0(\theta_x)$ and $c_0(\theta_x)$ give the values for the second and third order temperature coefficient of frequency for the angle of orientation θ_x. The columns headed $\partial a_0/\partial\theta$, $\partial b_0/\partial\theta$, and $\partial c_0/\partial\theta$ give the values for the derivatives of the first, second, and third-order temperature coefficient of frequency with respect to the angle of orientation, where θ is taken in degrees of arc. The column headed $b_0(\theta_y)=0$ gives the angle θ_y for which the second order temperature coefficient of frequency is zero. The column T_0 is the reference temperature at which the coefficients, a_0, b_0, c_0, are determined. For the measurements at Brush Laboratories $T_0=25$°C. was chosen; in all other cases the original reference temperatures of the evaluated graphs were used. In the column (T_i-T_0) are the differences between the inflection temperature T_i and the reference temperature T_0. The column "Reference for Measurement" lists the origin of the measurements: Brush Laboratories Company (B.L.); Bell Telephone Laboratories, Murray Hill, N. J. (B.T.L.); U. S. Signal Corps Engineering Laboratories, Fort Monmouth, N. J. (S.C.E.L.); British Post Office Engineering Department, Radio, Experimental, and Development Laboratory, Dollis Hill, London, England (G.P.O.), and Standard Piezo Company, Carlisle, Pa. (St.P.). The values from the Brush Laboratories are original measurements made with the technique described in Section II*A*. Values referring to sources other than Brush Laboratories are evaluated from frequency-temperature-angle characteristics in form of graphs.

Natural quartz and the ordinary types of synthetic quartz are discussed in this section; the properties of modified synthetic quartz with additions will be discussed in Section IV. Considering the zero angle, θ_x, for the first order temperature coefficient of frequency, a_0 all measurements show a difference of about 7 to 9′ of arc between the values for the fundamental and the third overtone. The difference for the zero angle between the third overtone and the higher modes is very small, about 1′. An explanation for this effect is given in

TABLE II

First-, Second-, and Third-Order Temperature Coefficients of Frequency and Their Derivatives with the Angle of Orientation of BT-Type Quartz Resonators Made from Natural Quartz

Quartz Type	Electrode Arrangement	Diameter Thickness Ratio	Order of Mode n	$a_0(\theta_x)=0$ θ_x	$b_0(\theta_x)$ $10^{-9}/(°C)^2$	$c_0(\theta_x)$ $10^{-12}/(°C)^3$	$\frac{\partial a_0}{\partial\theta}$ $10^{-6}/°\theta$	$\frac{\partial b_0}{\partial\theta}$ $10^{-9}/°\theta$	$\frac{\partial c_0}{\partial\theta}$ $10^{-12}/°\theta$	$\frac{\partial T_m}{\partial\theta}$	$T_0=T_{max}(\theta_x)$°C	References for Measurements
Natural	Air Gap	~20	1	−49°12′	−40	−132	−1.8	−2.0	38.0	−22.5	25	B.L.

Bechmann.[6] The dependence of the diameter-thickness ratio on the zero angle, θ_x, is well known.[7,8] Significant differences for the zero angle, θ_x, exist between natural quartz and the various types of synthetic quartz. Referring to the shift of the zero angle of the fundamental mode, the British quartz shows a difference of about 2′ from natural quartz, Brush Y-bar quartz 3′, Brush quartz grown on minor rhombohedral seed plates 5′, and quartz grown by Bell Telephone Laboratories has still a higher shift of the zero angle. Similar behavior of the zero angle is found for the higher order modes. These differences may be due to different growing conditions as well as to differences in the seed types used.

The second-order temperature coefficient of frequency, b_0, usually shows higher values for synthetic quartz compared with natural quartz, although the values for the British material are close to those of natural quartz.

The behavior of the third-order temperature coefficient of frequency, c, is very significant. Generally, the overtones show smaller values for c than the fundamental mode. The values for c are smaller for air gap-type resonators than for plated crystals. Much smaller values for c are evaluated for the overtones of crystals made from the British material. A small value for c is found for the resonators made from natural quartz having a diameter-thickness ratio of about 6. No dependence on the third order temperature coefficient as function of the diameter-thickness ratio is known yet.

The derivatives with respect to the angle of orientation for the first-order temperature coefficient of frequency, $\partial a_0/\partial\theta$, was found to be very constant for most types of quartz and order of modes. Only the evaluation of the Signal Corps measurements of Brush quartz gave slightly different values in the order of ± 2 per cent. The evaluation of the measurements carried out by Bell Telephone Laboratories for natural quartz lead to a substantially higher value, $-5.5\cdot 10^{-6}/°\theta$ compared with $-5.15\cdot 10^{-6}/°\theta$. The values for the derivatives with respect to the angle of the second order temperature coefficient of frequency, $\partial b_0/\partial\theta$, was found to be remarkably constant, equal to $-4.5\cdot 10^{-9}/°\theta$. The derivative of the third-order temperature coefficient of frequency, $\partial c_0/\partial\theta$, is in the order of 0 to $-20\cdot 10^{-12}/°\theta$, but an accuracy of about 5 per cent for the values of c do not allow for a higher accuracy for this derivative.

III. BT-Type Resonators

The frequency-temperature-angle characteristic of BT-type quartz resonators has a parabolic form. The values for the temperature coefficients of frequency obtained from the fundamental mode of BT-type resonators are in Table II above. The arrangement of this table is similar to that of Table I, describing the properties of AT-type resonators. The reference temperature, T_0, is identical to T_m, the temperature for the maximum of the frequency-temperature curve, since $a_0(\theta_x)=0$. No complete values for the constants of BT-type resonators made from synthetic quartz are available at present. The third order temperature coefficient of frequency is in the same order of magnitude as that for AT-type quartz resonators, but with opposite sign. However, the second-order temperature coefficient has a rather large value and outweighs the influence of the third order temperature coefficient.

The temperature at which the maximum of the frequency-temperature curve occurs, is in first approximation a linear function of the angle of orientation.

IV. Modified Quartz

Recently, a new development has had the aim of changing the properties of synthetic quartz, particularly of the frequency-temperature-angle characteristics of piezoelectric resonators. Considering AT-type reso-

[6] R. Bechmann, "Influence of the order of overtone on the temperature coefficient of frequency of AT-type quartz resonators," Proc. IRE, vol. 43, pp. 1667–1668; November, 1955.
[7] R. Bechmann, "Properties of quartz oscillators and resonators in the frequency range 300–5000 kc/s," *Hochfreq. und Elektroak.*, vol. 59; pp. 97–105; April, 1942.
[8] E. A. Gerber, "Temperature coefficient of AT cut quartz crystal vibrators," Proc. IRE, vol. 43, p. 1529; October, 1955.

nators, two effects are of particular interest: A reduction of the values of the third-order temperature coefficient of frequency compared with natural quartz, giving a smaller frequency-temperature change in the normal temperature range and a shift of the inflection temperature to higher values, giving a smaller frequency-temperature change at elevated temperature.

Numerous specimens of electronic grade quartz have been subjected to spectrographic analyses to determine the incidence of other elements in the quartz. The total amount of impurities ordinarily present, computed as oxides, has been found to be less than 0.04 per cent by weight.[9] The oxides frequently present are: Aluminum, lithium, boron, calcium, magnesium, manganese, sodium, and titanium.

The presence of detectable quantities of such elements as aluminum in varying proportions in natural quartz may have some effect upon the properties of natural crystals. Thus different samples of natural quartz have been found to have sufficient differences in the lattice parameters of the crystalline substance to make it impractical to use clear crystalline quartz in the calibration of X-ray diffraction cameras without independent determination of the parameters of the quartz used.[10] There is doubt as to whether or not these small variations in lattice parameters are due entirely to variations in amounts of impurities or are due in part to variations in the physical conditions prevailing during the formation of the quartz by geological processes. Nevertheless, in the applications of natural quartz crystals in the radio industry, natural quartz from different sources has displayed remarkable uniformity as far as all piezoelectric applications are concerned. Regardless of the source of the electronic grade quartz used, if the crystallographic orientation of a piezoelectric resonator plate is specified with respect to the orientation, no significant variations are encountered in the performance of the resulting resonator plates.

It has been found possible to modify the composition of quartz single crystals, in order to obtain different properties, in particular the frequency-temperature behavior of the *AT*-type resonators, by introducing some other elements during the growing process in a much greater amount than found in natural quartz. There is a possibility of substituting to some extent the silicon ion by another ion with an ionic radius fairly close to that of silicon ion of the valence charge of 4^+. Ions of similar radii but valences different from silicon still may be substituted in the lattice, provided a carrier ion is added to maintain the electrical balance. The carrier ion itself need not possess an ionic radius close to that of silicon since it could substitute as interstitial ion.

In general, synthetic quartz has considerably lower concentrations of impurities than are found in natural quartz, although the impurities in electronic-grade natural quartz ordinarily are very low. Furthermore, it is quite possible that the conditions under which quartz is produced synthetically differ very considerably from the geological conditions which gave rise to the formation of natural quartz crystals. Both the physical and chemical environments may be varied in the synthesis of quartz.

The Brush Laboratories Company has grown some quartz crystals with the addition of germanium dioxide[11] in NaOH solution and in Na_2CO_3 solution, the first resulting in quartz material containing in solid solution about 0.25 per cent by weight of germanium dioxide. From these materials, *AT*-type resonators have been prepared and the frequency-temperature characteristics have been measured. The results of these investigations are given in Table I. Considering synthetic quartz with germanium addition grown in NaOH solution, there is a shift of about 19′ for the zero angle of the first order temperature coefficient of frequency. The third-order temperature coefficient of frequency for the fundamental mode shows a considerably smaller value than that for natural quartz. This is a very desirable effect. The third-order coefficients of frequency for the third and fifth overtones, however, do not show these reduced values and are in the same order as natural quartz. More investigation is necessary to explain this effect. A disadvantage for practical purposes is the large shift of the zero angle which may be a function of the germanium concentration in quartz.

At the Brush Laboratories quartz crystals have been grown in the presence of manganese, silicon, lithium, boron, and aluminum, in addition to germanium already mentioned. Since the crystals were not large enough to provide resonators, no information regarding the piezoelectric behavior can be reported. Further work on modified quartz has been suspended at the Brush Laboratories.

Acknowledgment

The author wishes to thank H. W. Dodds and W. H. Charbonnet for the provision of the Pilot Plant quartz material and Dr. D. R. Hale for the special quartz types. Most of the cutting and orienting of the blanks were carried out by Dr. T. J. Turobinski, and the preparation of the resonators and measurements by H. G. Chase and C. Krausz. Most of the evaluations were carried out by Miss V. Durana. Some precision X-ray orientation tests were performed by W. R. Cook, Jr.

[9] C. S. Hurlbut, Jr., "Influence of twinning on the usability of quartz from various localities," *Amer. Mineralogist*, vol. 31, pp. 443–455 September–October, 1946.

[10] H. D. Keith, "The lattice-parameters of clear crystalline quartz," *Proc. Phys. Soc.* (London), vol. B63, pp. 208–214; March, 1950.

[11] F. Augustine and A. D. Schwope, "Quartz Crystal (Germanium in Quartz)," patent application U. S. Ser. No. 492,006; March 3, 1955.

Distortion in Frequency-Modulation Systems Due to Small Sinusoidal Variations of Transmission Characteristics*

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Summary—**It is shown that the distortion generated in fm systems by small sinusoidal ripples on either group delay or amplitude characteristics can be evaluated in terms of the distortion due to a single echo. Using results already established for echo distortion, curves are plotted relating intermodulation distortion of a frequency-division-multiplex signal to the amplitude, periodicity, and location of the small sinusoidal ripple for a 600 channel system. These curves are of value in estimating permissible limits of variation of transmission characteristics over the significant rf band.**

Introduction

An important design consideration governing frequency-division-multiplex radio telephony systems concerns the minimization of intermodulation distortion (usually appearing as unintelligible noise) due to the passage of the modulated carrier through nonlinear circuits. When the system employs frequency-modulation the problem of the evaluation of the distortion due to nonlinearity is of particular difficulty, owing to the complicated nature of the rf spectrum.

It is customary for analytical and test purposes to simulate the multiplex signal by a band of random noise of constant spectral density, introduced at a suitable level. This choice of model is suggested by the statistical properties of frequency-division-multiplex signals containing a large number of channels.[1]

With such a modulating signal, it is possible to evaluate the intermodulation distortion when the departures of transmission characteristics from their ideal forms can be represented by the first few terms of power series (with the departure from carrier frequency as variable).[2,3] In practice, however, adequate representation of characteristics by power series tends to require a substantial number of high order terms, leading to excessively elaborate distortion formulas.

Thus, it seems necessary to look for alternative approaches which may yield information about characteristics more complicated than the comparatively tractable low-order ones. One such approach involves echo distortion, which has been investigated quite extensively.[4–6] It is well known[7] that a single small echo is equivalent in its distorting effect to simultaneous small sinusoidal ripples superimposed on flat phase and amplitude characteristics. Alternatively, a small sinusoidal ripple on either characteristic alone is equivalent to a pair of equal-amplitude echoes, one advanced and one delayed by equal times. Since there is considerable information on the distorting effect of a single echo, a profitable next step would seem to be to investigate whether the distorting effect of a sinusoidal ripple associated with either characteristic alone bears any simple relationship to the distortion due to a single echo.

It is shown in the present paper that the distortion due to a small ripple on either phase or amplitude characteristic can be expressed as the product of the distortion due to a single echo (of appropriate amplitude, delay, and phase) and a trigonometrical factor involving the ripple wavelength (measured in units of frequency) and the baseband modulating frequency. From this relationship, curves have been constructed showing the distorting effects of ripples associated with either characteristic for a 600 channel system. These provide more information than has hitherto been available on permissible limits of variation of transmission characteristics (and hence on the limiting accuracy required in measuring equipment).

The harmonic distortion of single-tone frequency modulation due to small sinusoidal ripples on the transmission characteristics was evaluated by Assadourian.[8] It was shown in footnote reference 2 that the approach used by Assadourian could be extended to cover quite arbitrary shapes of characteristic, provided that the

* Original manuscript received by the IRE, June 5, 1956.

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[1] B. D. Holbrook and J. T. Dixon, "Load rating theory for multi-channel amplifiers," *Bell Sys. Tech. J.*, vol. 18, pp. 624–644; October, 1939.

[2] R. G. Medhurst, "Harmonic distortion of frequency-modulated waves by linear networks," *Proc. IEE*, vol. 101, pt. III, pp. 171–181; May, 1954.

[3] R. G. Medhurst and H. D. Hyamson, "Discriminator distortion in frequency-modulation systems" (submitted for publication to *Proc. IEE*).

[4] W. J. Albersheim and J. P. Schafer, "Echo distortion in the fm transmission of frequency-division-multiplex," Proc. IRE, vol. 40, pp. 316–328; March, 1952.

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[6] W. R. Bennett, H. E. Curtis, and S. O. Rice, "Inter-channel interference in fm and pm systems," *Bell Sys. Tech. J.*, vol. 34, pp. pp. 601–636; May, 1955.

[7] A. Bloch, "Modulation theory," *Jour. IEE*, vol. 91, pt. III, pp. 31–42; March, 1944.

[8] F. Assadourian, "Distortion of a frequency-modulated signal by small loss and phase variations," Proc. IRE, vol. 40, pp. 172–176; February, 1952.

resultant distortion was not too large. It is not yet known whether a similar extension is possible for the more complicated modulating signal considered in the present paper.

Analysis

Single Echo

As an introduction to the present work it will be useful to review briefly the case of a single echo of small amplitude.[4,5]

Consider one component of the spectrum of the modulated wave, having angular frequency ω and, for convenience, unit amplitude. After addition of the echo, this component becomes

$$\cos \omega t + r \cos \omega(t - \tau)$$

(where r is the relative echo amplitude and τ the delay time)

$$= (1 + r \cos \omega\tau) \cos \omega t + r \sin \omega\tau \sin \omega t$$
$$= A \cos (\omega t + \phi), \text{ say,}$$

where

$$A = \sqrt{1 + 2r \cos \omega\tau + r^2}$$

and

$$\phi = \tan^{-1}\left[\frac{-r \sin \omega\tau}{1 + r \cos \omega\tau}\right].$$

When r is sufficiently small, we have approximately

$$A = 1 + r \cos \omega\tau$$
$$\phi = - r \sin \omega\tau$$

so that the addition of the echo produces, to first order, the same distorting effect as passage through a network whose phase and amplitude characteristics each consist of sinusoidal ripples, of suitable phasing and amplitude.

It will in general be possible to represent the phase modulation as the sum of a number of tones. Thus, calling the phase modulation μ_t, we shall have

$$\mu_t = \sum_p f(p) \sin (pt + \phi_p). \tag{1}$$

The sum of signal and echo will be of the form

$$\cos (\omega_c t + \mu_t) + r \cos \left[\omega_c(t - \tau) + \mu_{t-\tau}\right]$$

where ω_c is the angular carrier frequency

$$= B \cos (\omega_c t + \mu_t + \Psi), \text{ say.}$$

Then, for sufficiently small r,

$$\Psi \simeq - r \sin \left[\omega_c\tau + \mu_t - \mu_{t-\tau}\right]$$
$$= - r \sin (\omega_c\tau) \cos \left[\mu_t - \mu_{t-\tau}\right] - r \cos \omega_c\tau \sin \left[\mu_t - \mu_{t-\tau}\right].$$

From (1),

$$\mu_t - \mu_{t-\tau}$$
$$= \sum_p f(p) \sin (pt + \phi_p) - \sum_p f(p) \sin (pt - p\tau + \phi_p)$$
$$= 2 \sum_p f(p) \sin (\tfrac{1}{2}p\tau) \cos (pt - \tfrac{1}{2}p\tau + \phi_p)$$
$$= 2 \sum_p f(p) \sin (\tfrac{1}{2}p\tau) \cos \left[p(t - \tfrac{1}{2}\tau) + \phi_p\right]. \tag{2}$$

Suppose that

$$- \cos \left[\mu_t - \mu_{t-\tau}\right] = \sum_l \frac{1}{r} D_S(l) \cos \left[l(t - \tfrac{1}{2}\tau) + \Psi_l\right] \tag{3}$$

$$- \sin \left[\mu_t - \mu_{t-\tau}\right] = \sum_m \frac{1}{r} D_C(m) \cos \left[m(t - \tfrac{1}{2}\tau) + \xi_m\right]. \tag{4}$$

Then, from (2), the phase modulation distortion becomes

$$\sin (\omega_c\tau) \sum_l D_S(l) \cos \left[l(t - \tfrac{1}{2}\tau) + \Psi_l\right]$$
$$+ \cos (\omega_c\tau) \sum_m D_C(m) \cos \left[m(t - \tfrac{1}{2}\tau) + \xi_m\right]. \tag{5}$$

When the frequency modulation is a flat noise band, of the form

$$M_t = \alpha \sum_{p=p_0}^{p_m} \cos (pt + \phi_p)$$

(where p increases in unit steps and ϕ_p is a random phase angle), we have

$$\mu_t = \alpha \sum_{p=p_0}^{p_m} \frac{1}{p} \sin (pt + \phi_p).$$

Then, (3) and (4) become

$$- \cos \left[\mu_t - \mu_{t-\tau}\right] = \sum_{p=0}^{\infty} \frac{1}{r} D_S(p) \cos \left[p(t - \tfrac{1}{2}\tau) + \Psi_p\right] \tag{6}$$

$$- \sin \left[\mu_t - \mu_{t-\tau}\right] = \sum_{p=0}^{\infty} \frac{1}{r} D_C(p) \cos \left[p(t - \tfrac{1}{2}\tau) + \xi_p\right] \tag{7}$$

where p again increases in unit steps, and the phase modulation distortion, given generally by (5), becomes

$$\sin (\omega_c\tau) \sum_{p=0}^{\infty} D_S(p) \cos \left[p(t - \tfrac{1}{2}\tau) + \Psi_p\right]$$
$$+ \cos (\omega_c\tau) \sum_{p=0}^{\infty} D_C(p) \cos \left[p(t - \tfrac{1}{2}\tau) + \xi_p\right]. \tag{8}$$

D_S and D_C depend on the delay time, the modulation conditions and the baseband frequencies. They are known over a wide range of conditions.

Sinusoidal Ripple on Group Delay Characteristic

Let the phase characteristic be of the form

$$\phi = - r \sin \omega\tau,$$

so as to preserve the same notation as in the single echo case. The corresponding group delay characteristic is

$$\frac{d\phi}{d\omega} = -r\tau \cos \omega\tau.$$

Since

$$\cos(\omega t - r \sin \omega\tau) \simeq \cos \omega t - \tfrac{1}{2} r \cos[\omega(t+\tau)] + \tfrac{1}{2} r \cos[\omega(t-\tau)],$$

when r is sufficiently small, the assumed phase characteristic is equivalent to two echoes, one advanced and one retarded, provided that the amplitude characteristic is flat.

Following the same procedure as in the case of the single echo, it is found that the phase modulation distortion is given approximately by

$$-\tfrac{1}{2} r \sin(\omega_c\tau) \cos[\mu_t - \mu_{t+\tau}] + \tfrac{1}{2} r \cos(\omega_c\tau) \sin[\mu_t - \mu_{t+\tau}]$$
$$-\tfrac{1}{2} r \sin(\omega_c\tau) \cos[\mu_t - \mu_{t-\tau}] - \tfrac{1}{2} r \cos(\omega_c\tau) \sin[\mu_t - \mu_{t-\tau}]. \quad (9)$$

Assuming that μ_t can be written as in (1), we have shown that

$$\mu_t - \mu_{t-\tau} = 2 \sum_p f(p) \sin(\tfrac{1}{2} p\tau) \cos[p(t - \tfrac{1}{2}\tau) + \phi_p]. \quad (2)$$

Also,

$$\mu_t - \mu_{t+\tau} = -2 \sum_p f(p) \sin(\tfrac{1}{2} p\tau) \cos[p(t + \tfrac{1}{2}\tau) + \phi_p]. \quad (10)$$

It was assumed further in the previous section that

$$\cos[\mu_t - \mu_{t-\tau}] = -\sum_e \frac{1}{r} D_S(l) \cos[l(t - \tfrac{1}{2}\tau) + \Psi_l] \quad (3)$$

and

$$\sin[\mu_t - \mu_{t-\tau}] = -\sum_m \frac{1}{r} D_C(m) \cos[m(t - \tfrac{1}{2}\tau) + \xi_m]. \quad (4)$$

From these, the corresponding functions for the advanced echo can be immediately written down, since

$$\begin{aligned}
\cos[\mu_t - \mu_{t+\tau}] &= \cos\left\{-2 \sum_p f(p) \sin(\tfrac{1}{2} p\tau) \cos[p(t + \tfrac{1}{2}\tau) + \phi_p]\right\} \\
&= \cos\left\{2 \sum_p f(p) \sin(\tfrac{1}{2} p\tau) \cos[p(t + \tfrac{1}{2}\tau) + \phi_p]\right\} \\
&= -\sum_l \frac{1}{r} D_S(l) \cos[l(t + \tfrac{1}{2}\tau) + \Psi_l], \qquad (11)
\end{aligned}$$

and

$$\begin{aligned}
\sin[\mu_t - \mu_{t+\tau}] &= \sin\left\{-2 \sum_p f(p) \sin(\tfrac{1}{2} p\tau) \cos[p(t + \tfrac{1}{2}\tau) + \phi_p]\right\} \\
&= -\sin\left\{2 \sum_p f(p) \sin(\tfrac{1}{2} p\tau) \cos[p(t + \tfrac{1}{2}\tau) + \phi_p]\right\} \\
&= +\sum_m \frac{1}{r} D_C(m) \cos[m(t + \tfrac{1}{2}\tau) + \xi_m].
\end{aligned}$$

Then, the phase modulation distortion becomes

$$\begin{aligned}
&+ \tfrac{1}{2} \sin(\omega_c\tau) \sum_l D_S(l) \cos[l(t + \tfrac{1}{2}\tau) + \Psi_l] \\
&+ \tfrac{1}{2} \cos(\omega_c\tau) \sum_m D_C(m) \cos[m(t + \tfrac{1}{2}\tau) + \xi_m] \\
&+ \tfrac{1}{2} \sin(\omega_c\tau) \sum_l D_S(l) \cos[l(t - \tfrac{1}{2}\tau) + \Psi_l] \\
&+ \tfrac{1}{2} \cos(\omega_c\tau) \sum_m D_C(m) \cos[m(t - \tfrac{1}{2}\tau) + \xi_m] \\
&= \sin(\omega_c\tau) \sum_l D_S(l) \cos(\tfrac{1}{2} l\tau) \cos(lt + \Psi_l) \\
&\quad + \cos(\omega_c\tau) \sum_m D_C(m) \cos(\tfrac{1}{2} m\tau) \cos(mt + \xi_m).
\end{aligned}$$

For modulation by a flat noise band, using (6) and (7), this becomes

$$\begin{aligned}
&\sin(\omega_c\tau) \sum_{p=0}^{\infty} D_S(p) \cos(\tfrac{1}{2} p\tau) \cos(pt + \Psi_p) \\
&+ \cos(\omega_c\tau) \sum_{p=0}^{\infty} D_C(p) \cos(\tfrac{1}{2} p\tau) \cos(pt + \xi_p).
\end{aligned} \qquad (12)$$

Since $D_S(p)$ and $D_C(p)$ express the distortion due to a single echo, as in (8), we have now arrived at an expression for the distortion due to a group delay sinusoidal ripple in terms of the distortion generated by a single echo, together with a trigonometrical factor involving the repetition rate of the ripple and the position in the baseband at which the distortion is measured.

Sinusoidal Ripple on Amplitude Characteristic

The amplitude characteristic is taken as

$$A = 1 + r \cos \omega\tau \quad \text{where} \quad r \ll 1.$$

The analysis is closely similar to that of the previous section. For noise band modulation, the final result, corresponding to (12) in the previous section, is: phase modulation distortion =

$$\begin{aligned}
&\sin(\omega_c\tau) \sum_{p=0}^{\infty} D_S(p) \sin(\tfrac{1}{2} p\tau) \sin(pt + \Psi_p) \\
&+ \cos(\omega_c\tau) \sum_{p=0}^{\infty} D_C(p) \sin(\tfrac{1}{2} p\tau) \sin(pt + \xi_p).
\end{aligned} \qquad (13)$$

Numerical Results for a 600 Channel System

The modulation conditions are those used for the numerical example of footnote reference 9. The baseband extends from 60 kc to 2540 kc, and the peak deviation (taken arbitrarily as 11 db above the rms multichannel deviation which is exceeded for not more than 1 per cent of the busy hour) is 4.0 mc. To a good approximation,[5] D_S and D_C are given by formulas of the form

$$\frac{D_S}{S} = \frac{D}{S}\left[1 - \exp\left(-\frac{D_2}{S}\bigg/\frac{D}{S}\right)\right] \tag{14}$$

and

$$\frac{D_C}{S} = \frac{D}{S}\left[1 - \exp\left(-\frac{D_3}{S}\bigg/\frac{D}{S}\right)\right]. \tag{15}$$

Here, D_S and D_C have the same meaning as in the previous section above, S is the undistorted signal level (phase modulation) in the same 4 kc channel, D_2 and D_3 are respectively second and third order distortions and D/S is the distortion/signal ratio due to a long-delayed echo. For small r, D_2/S and D_3/S are given[4] by

$$\begin{aligned}\frac{D_2}{S} &= 0.20 r\tau^2 s p\sqrt{1 - \tfrac{1}{2}(p/p_m)} \\ &= 0.14 r\tau^2 s p_m \text{ in the top channel}\end{aligned} \tag{16}$$

and

$$\begin{aligned}\frac{D_3}{S} &= 0.028 r\tau^3 s^2 p\sqrt{1 - \tfrac{1}{3}(p/p_m)^2} \\ &= 0.023 r\tau^3 s^2 p_m \text{ in the top channel.}\end{aligned} \tag{17}$$

where s is the peak deviation in the sense defined above (radians/sec.), p is a baseband frequency (radians/sec.) and p_m is the maximum baseband frequency (radians/sec.).

In the top channel, D/S is given, for small r, by

$$\frac{D}{S} = rK$$

where K is a function of s/p_m, shown graphically in Fig. 7 of footnote reference 5. In the present case, K is about 1.05.

Numerical values based on (12), (13), (14), and (15) are shown in Figs. 1 to 4 (p. 1612). Two sets of curves have been plotted for each type of characteristic, one set relating to characteristics disposed symmetrically about carrier frequency, and the other to characteristics disposed skew-symmetrically. These dispositions of characteristics have the convenient analytical feature that only the second and first terms, respectively, of (12) and (13) are required. The plotted variation of characteristic is half the total variation over a band 12 mc wide centered on the carrier frequency, this being the frequency band outside which equalization need not be maintained.[9]

The curves give distortion in the top channel. A striking feature of these curves is that the permissible variation periodically rises sharply, as the ripple wavelength varies. This phenomenon is associated with the factors cos $(1/2p\tau)$ and sin $(1/2p\tau)$ in (12) and (13). These zeros of distortion are probably of no particular design value, since under such conditions substantial distortion will occur elsewhere in the baseband.

Curves are plotted for three distortion levels, -70, -80, and -90 dbmo (*i.e.*, db referred to a milliwatt at zero relative level, the distortions being measured in a 4 kc channel: to convert these levels to distortion/signal ratio, add 13 db[9]). The value required in a particular case will depend on the system design and the over-all performance envisaged. According to a system analysis[9] based on CCIR over-all specifications for a system 175 miles (280 km) long, containing five repeater stations with an average spacing of 29 miles, the distortion level permitted for a single amplifier is -82 dbmo.

Limiting Forms of Transmission Characteristics

It will be noticed in the figures that the curves of constant distortion associated with the group delay and amplitude characteristics behave somewhat differently in the region of large ripple spacing. Some discussion seems required to clarify this. In the case of group delay characteristics, the symmetrical characteristic tends to the form $a_0 + a_1(\omega - \omega_c)$, and the skew-symmetrical characteristic to $a_0 + a_2(\omega - \omega_c)^2$ when the separation of adjacent maxima and minima becomes large.

The corresponding phase characteristic is obtained from the group delay characteristic by integration with respect to frequency. The phase characteristic for the symmetrical case is therefore $a_0(\omega - \omega_c) + \frac{1}{2}a_1(\omega - \omega_c)^2$, and for the skew-symmetrical case $a_0(\omega - \omega_c) + \frac{1}{3}a_2(\omega - \omega_c)^3$. The distortion generated by these phase characteristics may be evaluated.[3] The results give the limiting values appearing at the right hand edges of Figs. 1 and 2.

In the case of the amplitude characteristics, the curves of constant distortion do not tend to limits, though in the figures these curves have been terminated in order to exclude amplitude characteristics having excessive excursions. The reason for this failure to tend to limits is to be found in the limiting forms of the characteristics which, as in the phase characteristic case, become $b_0 + b_1(\omega - \omega_c)$ and $b_0 + b_2(\omega - \omega_c)^2$. Neither of these terms, to the order considered here, generate distortion,[2,3] so that for fixed distortion no limit to the excursion in a given band exists, beyond that imposed by the condition that the equivalent echo amplitude must not be too large.

As a matter of interest, distortion values due to the lowest order term giving the appropriate type of symmetry and generating distortion are shown at the right hand sides of Figs. 3 and 4.

[9] R. G. Medhurst, "RF bandwidth of frequency-division multiplex systems using frequency modulation," Proc. IRE, vol. 44, pp. 189–199; February, 1956.

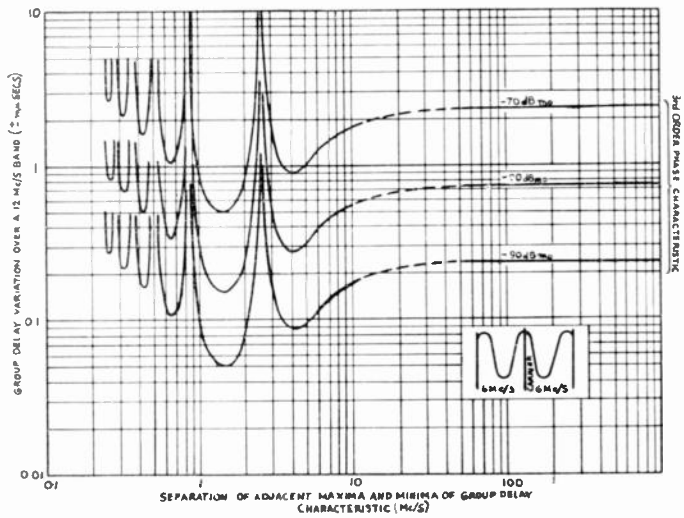

Fig. 1—FM distortion due to group delay characteristic consisting of a sinusoidal ripple arranged skew-symmetrically with respect to carrier frequency. Top 4 kc channel; number of channels = 600; maximum modulating frequency = 2.540 mc; peak deviation (*i.e.*, 11 db above rms) = 4.0 mc.

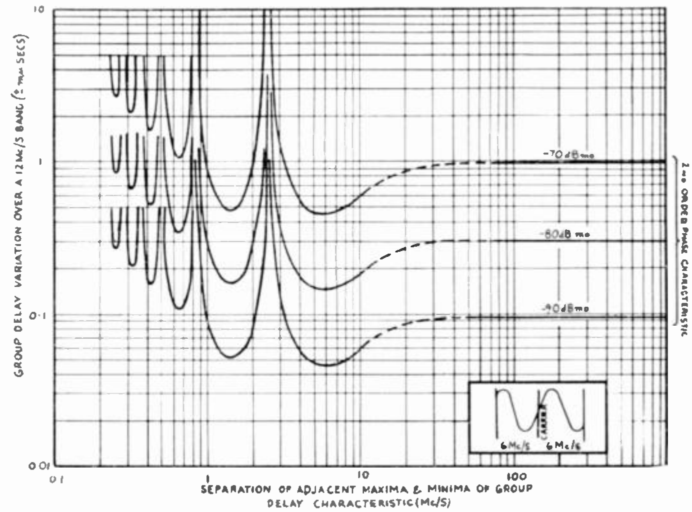

Fig. 2—FM distortion due to group delay characteristic consisting of a sinusoidal ripple arranged symmetrically with respect to carrier frequency. Number of channels = 600; top 4 kc channel; maximum modulating frequency = 2.540 mc; peak deviation (*i.e.*, 11db above rms) 4.0 mc.

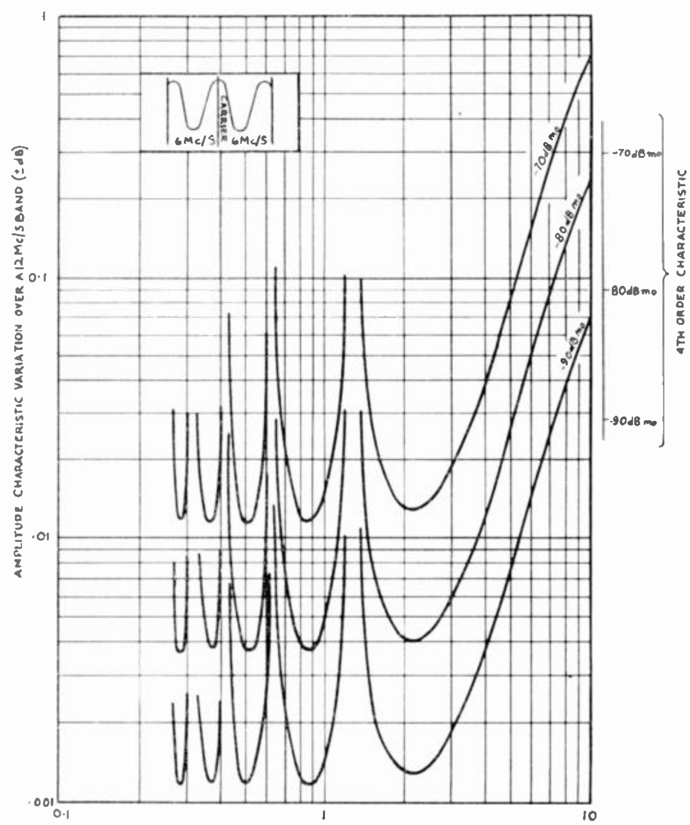

Fig. 3—FM distortion due to amplitude characteristic consisting of a sinusoidal ripple arranged skew-symmetrically with respect to carrier frequency. Top 4 kc channel; number of channels = 600; maximum modulating frequency = 2.540 mc; peak deviation (*i.e.*, 11 db above rms) = 4.0 mc.

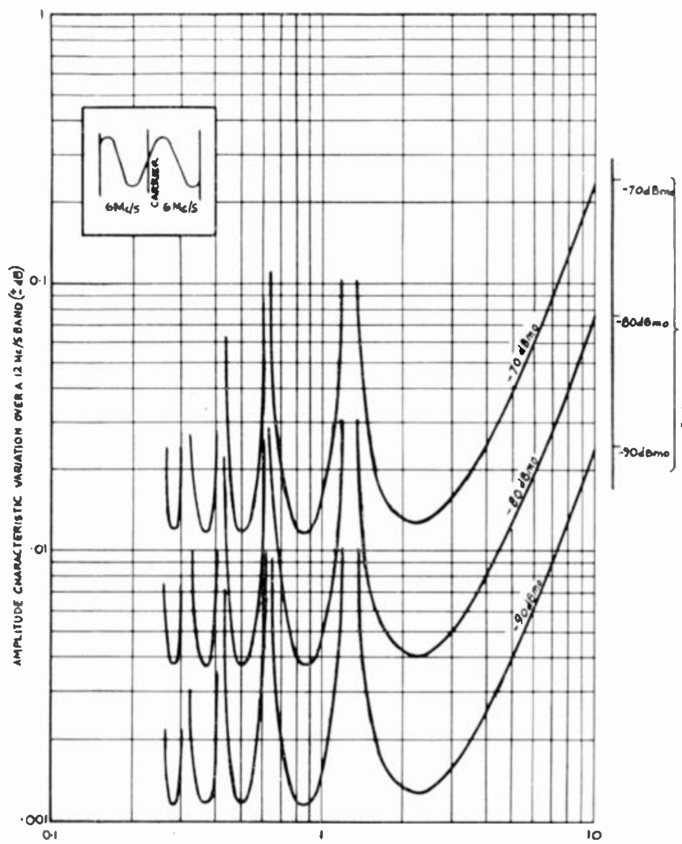

Fig. 4—FM distortion due to amplitude characteristic consisting of a sinusoidal ripple arranged symmetrically with respect to carrier frequency. Top 4 kc channel; number of channels 600; maximum modulating frequency 2.540 mc; peak deviation (*i.e.*, 11 db above rms) 4.0 mc.

Precision Electronic Switching with Feedback Amplifiers*

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Summary—**Electronic-switching signal transmission devices have been, until recently, used principally for the simultaneous display of more than one waveform on a cathode ray oscillograph. During the past few years, however, in the field of electronic analog computers, several applications have required the development of electronic switches which have transmission characteristics consistent with the performance of precision computing elements. These precision switches utilize highly stabilized feedback amplifiers with the switching elements included in the forward gain portion of the loop to minimize the nonlinear characteristic effects. Multiple output switches, which provide either current or voltage output signal transmission from a single input voltage source, are available. Precision electronic time-division multipliers have been developed which utilize both current and voltage type switching circuits. Modulators and demodulators, utilizing voltage type switched-feedback amplifiers, have been developed with a linearity of 0.1 per cent. A multiple input switch has also been developed which involves an unusual circuit design. This design incorporates two separate input stages and a common output stage with separate feedback paths provided between the common output and the two inputs. With a square wave keying voltage alternately activating the two input stages, signals applied to the two input stages can be alternately connected to the output. The transmission stability and precision are largely dependent on the feedback loop gain of the amplifier and the switching speed on the amplifier bandwidth.**

INTRODUCTION

ELECTRONIC SWITCHES have been widely used as on-off and signal transmission control devices. The principal use for on-off switches has been in the field of electronic digital computers where electronic switching techniques have been highly developed for the control of current and voltage levels. In most of these applications, diodes, multi-electrode flip-flops, and magnetic cores are used to cause a current or voltage to be switched on or off. The change in level usually is a large percentage of the current or voltage value which represents one state of operation of the device. The precision of level, therefore, is usually insignificant. In digital computer applications, switching time is the most important switching-circuit operating characteristic.

Until recently, the electronic switches which have been developed to control signal transmission have been used for the simultaneous display of more than one waveform on a cathode ray oscillograph[1] where the excellence of transmission is measured in terms of transient response. Although gain and phase relations need not be held to close absolute tolerances, the channels of the switch are normally very similar in operating characteristics. The switching circuit must not generate unwanted signals which would cause an erroneous display.

With the development of precision electronic analog computers, many programs were initiated to provide all electronic analog devices which would perform the mathematical operations of multiplication and function generation with a precision comparable to the linear operations readily accomplished by high gain operational amplifiers. Since multiplication is a basic computer operation which can be used for function generation, the need for an all-electronic function generator accelerated the search for a completely electronic multiplying circuit.

A review of analog multiplier development[2] indicates that precision switching has been the basis for several successful multiplying schemes. The use of precision electronic switching, however, is not restricted to analog computer multipliers. These switches are currently being used in modulation and demodulation circuits and have been considered for multiplexing applications and comparison transmission measuring schemes. Undoubtedly, there are many other applications where such devices can be used to advantage.

All of the switches described utilize a high gain feedback amplifier to minimize the differences and nonlinearities in the electronic elements which are used to switch the transmission paths. The switches are separated into classes with multiple inputs and those with multiple outputs. The two classes are divided into current-switching circuits and voltage-switching circuits.

Although most of the material presented in this paper represents original development work carried on at the Bendix Research Laboratories and the Massachusetts Institute of Technology, a significant amount of review information has been included to provide a complete picture of the state of the art on electronic switching techniques as they apply to analog computers.

MULTIPLE OUTPUT SWITCHES

The multiple output class switches receive one input signal which can be channeled to one of two or more

* Original manuscript received by the IRE, August 8, 1956. Presented at the 1955 Wescon Session on "Computers II—Analogue Computers," San Francisco, Calif., August 26, 1955.

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[1] H. J. Reich, "An electronic switch for the simultaneous observation of two waves with the cathode ray oscillograph," *Rev. Sci. Instr.*, vol. 12, p. 191; 1941.

[2] C. M. Edwards, "Survey of analog multiplication schemes," *J. Assoc. Computing Machinery*, vol. 1, pp. 27–35; January, 1954.

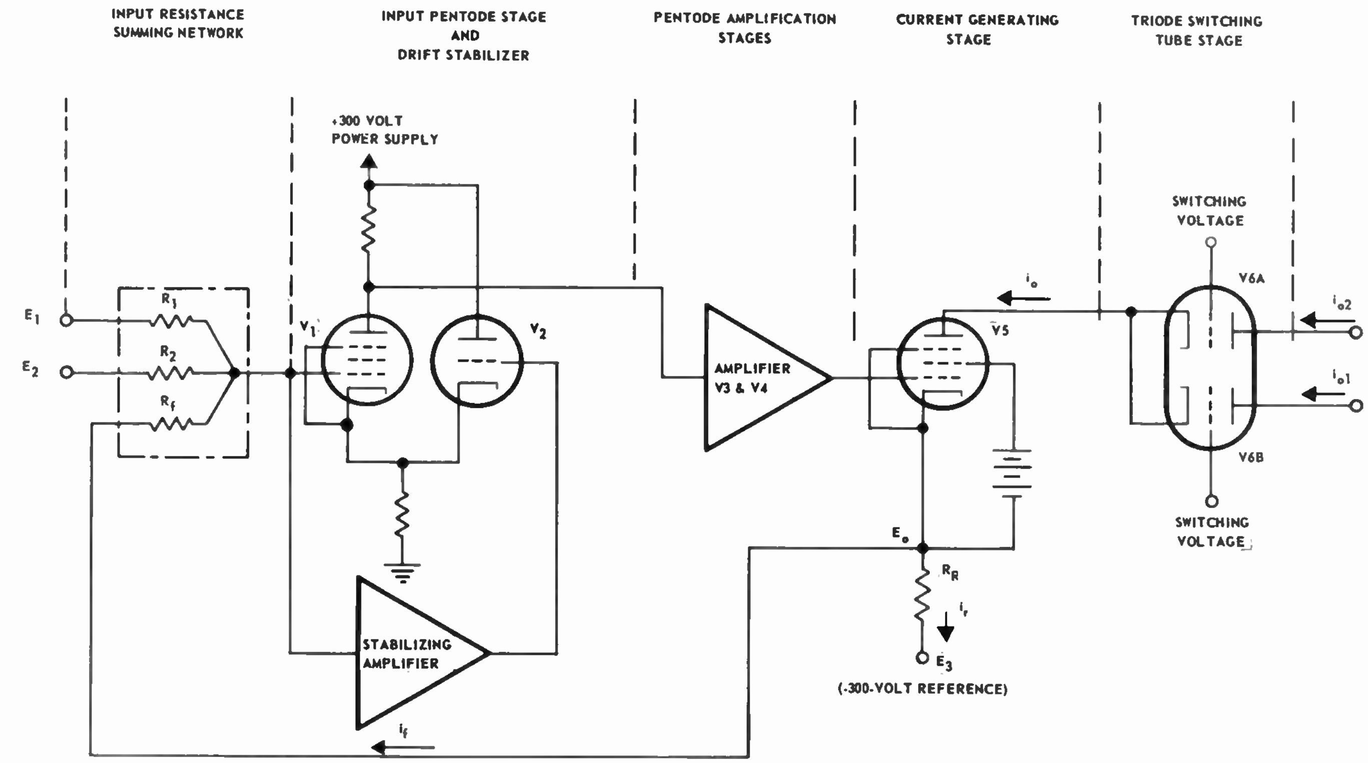

Fig. 1—Precision current switch—functional schematic.

outputs. Such switches have been used extensively in precision time-division multiplier circuits developed within the last few years. In 1952, E. A. Goldberg of the Radio Corporation of America announced the development of a precision multiplier utilizing a feedback current switch as the principal element.[3] C. D. Morrill and R. V. Baum of the Goodyear Aircraft Corporation, also in 1952, described a high-accuracy multiplier utilizing a feedback voltage switch as the primary element.[4] In the latter part of 1951, the Research Laboratories Division of the Bendix Aviation Corporation initiated the development of a large scale high-performance one-to-one time scale flight simulator for the Navy. Since this simulator is required to accomplish high speed multiplications, Bendix undertook the development of an electronic multiplier based on the time-division principle. The current switch which was developed for use in the multiplier is described in the following paragraphs.

A functional diagram of the current switch which is used in the master (or time-division channel) and slaves (or multiplying channels) of the unit is given in Fig. 1. A current i_o is established by tube V5 and the direct coupled amplifier composed of tubes V1, V2, V3, and V4. Because of the high gain of the dc amplifier, the current i_o is accurately determined as follows:

$$i_o = i_f + i_R$$

$$-i_f = -\frac{E_o}{R_f} = \frac{E_1}{R_1} + \frac{E_2}{R_2}$$

$$-E_o = \frac{R_f}{R_1} E_1 + \frac{R_f}{R_2} E_2$$

$$i_R = \frac{E_o - E_3}{R_R}.$$

Therefore,

$$-i_o = \frac{\frac{R_f E_1}{R_1} + \frac{R_f E_2}{R_2} + E_3}{R_R} + \frac{E_1}{R_1} + \frac{E_2}{R_2}$$

$$-i_o = \frac{R_f E_1}{R_1 R_R} + \frac{R_f E_2}{R_2 R_R} + \frac{E_3}{R_R} + \frac{E_1}{R_1} + \frac{E_2}{R_2}$$

$$-i_o = E_1\left[\frac{R_f}{R_R R_1} + \frac{1}{R_1}\right] + E_3 \frac{1}{R_R} + E_2\left[\frac{R_f}{R_R R_2} + \frac{1}{R_2}\right].$$

[3] E. A. Goldberg, "A high-accuracy time-division multiplier," *RCA Rev.*, vol. 13, pp. 265–274; (September, 1952), and "Project Cyclone Symposium II on Simulation and Computing Techniques," Part 2, Reeves Instr. Corp., New York, N. Y., pp. 215–223; April 28–May 2, 1952.

[4] C. D. Morrill and R. V. Baum, "A stabilized electronic multiplier," IRE Trans., vol. PGEC-1, pp. 52–59; December, 1952.

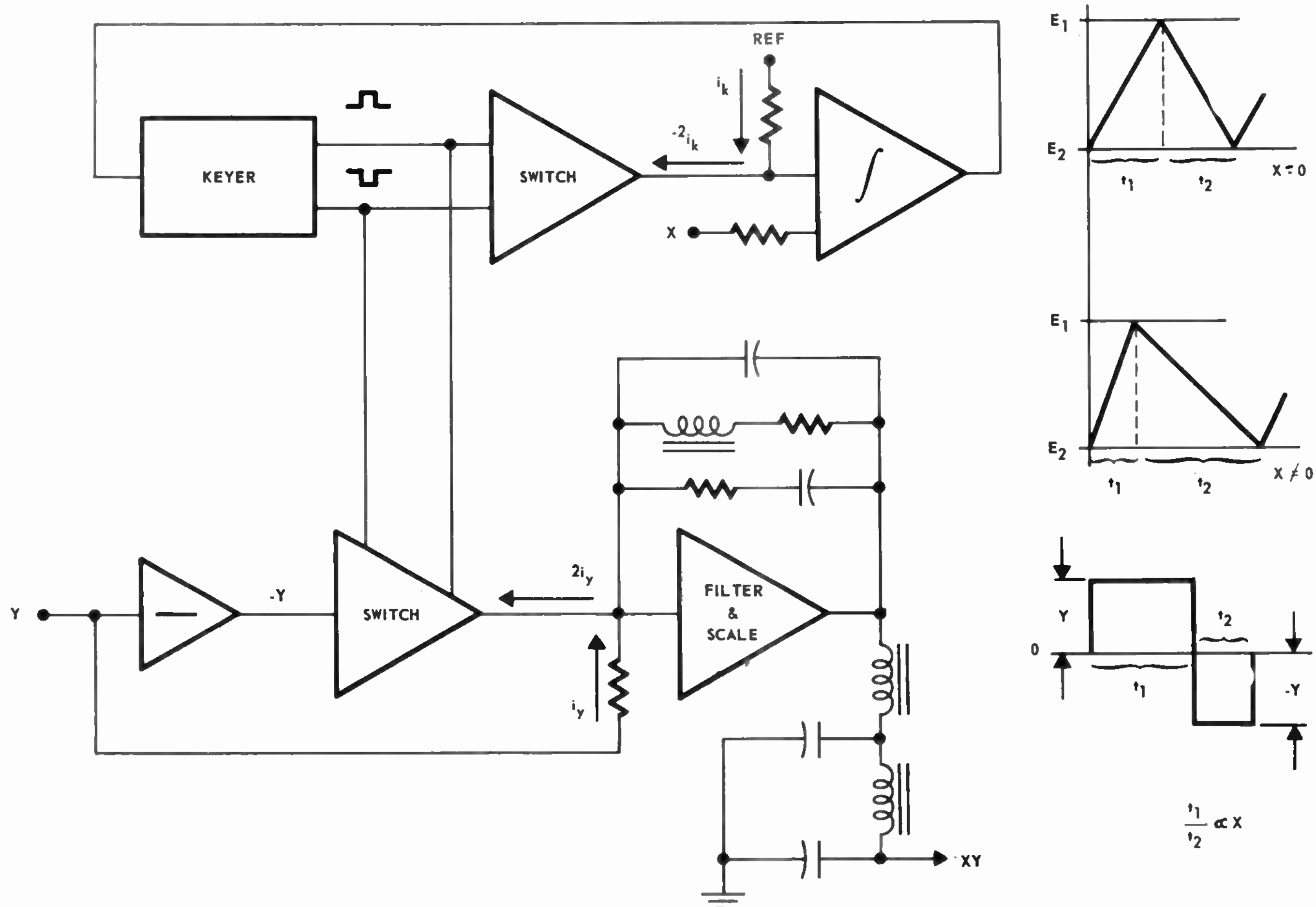

Fig. 2—Time division multiplier—functional schematic.

In actual operation, E_2 and E_3 are fixed reference voltages while E_1 is a signal voltage. Thus, the current i_o varies in proportion to E_1 with respect to a quiescent current determined by E_2 and E_3. Switching tubes V6A and V6B are alternately made conducting by a control circuit, such that either i_{o1} equals i_o and i_{o2} equals zero, or i_{o2} equals i_o and i_{o1} equals zero. Because of the current generating action of tube V5 in combination with the high gain dc amplifier, the impedance variations of tubes V6A and V6B have a negligible effect on i_{o2} and i_{o1}. Therefore, the signal represented by voltage E_1 can be switched to either output with little modification. For example, at a switching rate of 25 kilocycles, currents i_{o1} and i_{o2} are alternately proportional to voltage E_1 to within 0.1 per cent over the full range of E_1.

A functional diagram of the electronic multiplier utilizing the switch shown in Fig. 1 is given in Fig. 2. As shown in Fig. 2, the upper switch is considered part of the master or X channel, while the lower switch is considered part of the slave or Y channel. A number of Y channels may be contained in any given circuit arrangement. In the X channel the purpose of the switch is to alternately supply a current to the integrator exactly equal to $2i_k$ and zero. The integrator output changes at a rate in the positive direction that, depending on the value of X, is different from the rate in the negative direction. The time difference between time t_1 and time t_2 is proportional to X, since the switching occurs at the same output level for all periods of operation. Therefore, the accuracy of the time difference (t_1-t_2) is directly dependent on the accuracy of the switch current i_{o1} for the two conditions $i_{o1}=2i_k$ and $i_{o1}=0$. (The period of the operation is equal to time t_1 plus time t_2 and changes with X. For the multiplier under discussion, the period corresponds to a repetition rate of 25 kilocycles which decreases to approximately 12 kilocycles for a full scale X.) Similarly, the Y switch supplies an accurate current to the filter amplifier in the slave channel such that, alternately, current i_{o1} equals current $2i_Y$ and current i_{o1} equals zero. The value of $2i_Y$, however, is not constant but is proportional to Y. Consequently the input to the filter amplifier is alternately proportional to $+Y$ and $-Y$. The area represented by (Yt_1-Yt_2) is, therefore, proportional to the product XY, and the filter amplifier obtains the average of the area to generate the XY product and reduce the repetition frequency components to an acceptable level. This filtering action introduces the bandwidth limitation in this type of multiplier circuit. Multiplier linearity characteristics with each input separately varied are given in Fig. 3. It should be noted that for these tests the switch in the Y channel provided a variable output when Y was changed and a constant output when X was changed. The results indicate an over-all multiplier performance of 0.1 per cent or better, and thus verify a highly precise switching operation.

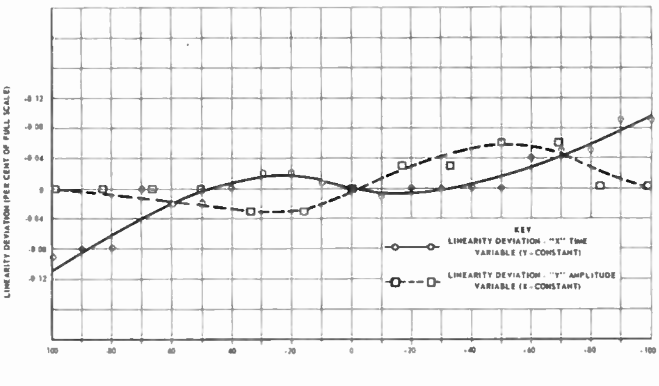

Fig. 3—Time division multiplier linearity.

The circuit for a multiple output voltage switch differs from a current switch in one important aspect. For the current switch shown in Fig. 1, the feedback circuit is essentially the same for both output conditions since tube V5 effectively isolates the feedback circuit from tube V6. In contrast, for a voltage switch the feedback circuit is provided by a different set of components for the respective output conditions as shown in Fig. 4. It is thus possible for the output voltage E_A of the dc amplifier to be quite different for the two switching conditions depending on the characteristics of the switching elements. Therefore, the bandwidth-voltage output characteristics are much more severe at a given switching rate for a voltage switch as compared to a current switch of the type shown in Fig. 1. Similar to the current switch, one of the switching elements of the voltage switch (Fig. 4) must be conducting in order to provide a closed loop around the amplifier at all times. Unlike the current switch, the output impedance of the voltage switch can be very low since it is an inverse function of the loop gain of the dc amplifier.

In the time-division multiplier developed at the Goodyear Aircraft Corporation,[5] the switch elements are triodes which are grid-controlled by a voltage obtained from a bistable multivibrator. In applications where highly idealized breakpoints are required for function generation, the switching elements can be diodes that are properly biased to conduct according to the desired

[5] *Ibid.*

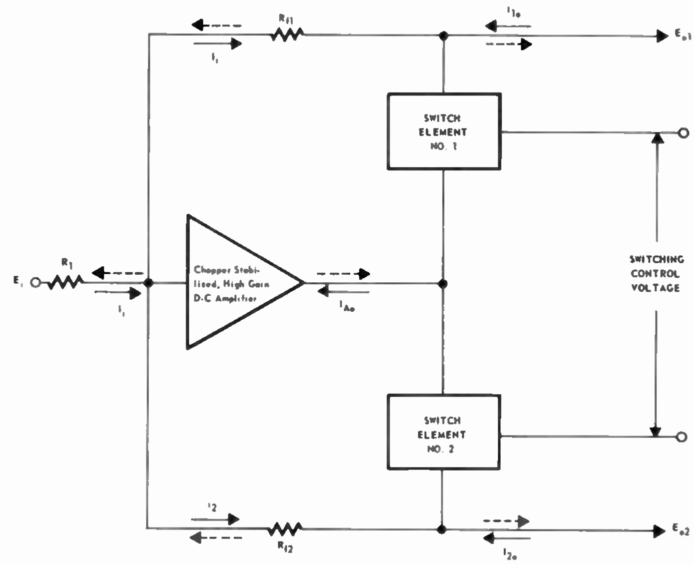

Fig. 4—Voltage feedback switch block diagram.

breakpoint. A review of the applicable literature indicates that many electronic analog users have devised circuits utilizing diodes in the feedback network of dc operational amplifiers, but only two sources[6,7] are cited.

Utilizing the voltage feedback switch principle, the Bendix Research Laboratories developed a highly linear and stable modulator and demodulator for use with ac computing resolvers that are part of the Navy Flight Simulator. The modulator and demodulator circuit is shown in Fig. 5. In an application, there are

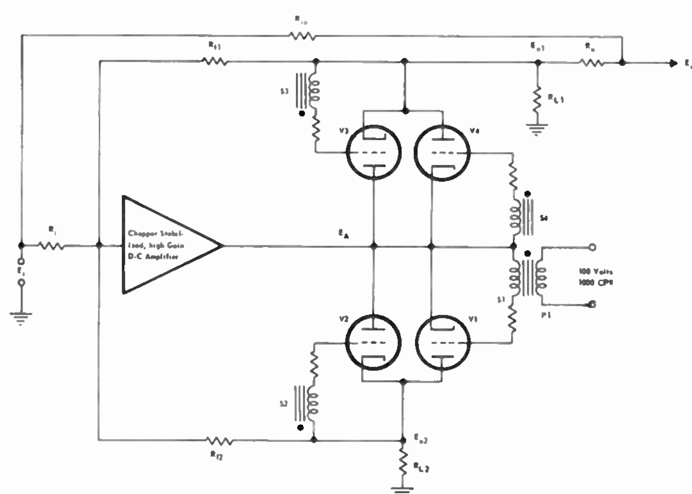

Fig. 5—Feedback switch modulator (or demodulator) functional schematic.

[6] G. D. McCann, C. H. Wilts, and B. N. Locanthi, "Application of the California Institute of Technology electric analog computers to non-linear mechanics and servomechanisms," *Trans. AIEE*, pp. 652–660; 1949.

[7] C. D. Morrill and R. V. Baum, "The role of diodes in an electronic differential analyzer," in "Project Cyclone Symposium II on Simulation and Computing Techniques," Part 2, Reeves Instr. Corp., New York, N. Y., pp. 201–213; April 28–May 2, 1952.

slight differences between the two units to provide the different gain factors required and to provide a simple filter network at the demodulator output. The principle components of the switch modulator and demodulator are the high gain dc amplifier and the switching tubes V1, V2, V3, and V4. The conduction time of the switching tubes is controlled by the voltage applied between the grid and the cathode by the four secondaries of the switching transformer. Since the switching voltage magnitude is much greater than the grid cutoff voltage magnitude, current limiting resistors are required to protect the switching tube grids. The secondaries S3 and S4 are phased to provide the proper output phasing for the generated suppressed carrier signal while the secondaries S1 and S2 are phased to maintain a feedback loop around the amplifier through tubes V1 and V2 during the nonconducting periods of tubes V3 and V4. To minimize the transient effects in the amplifier, the two feedback paths are essentially identical. It has been found, however, that the feedback resistance elements in the unused output can be much less precise than those used in the active output. The waveforms encountered at significant points in the circuit are shown in Fig. 6. The

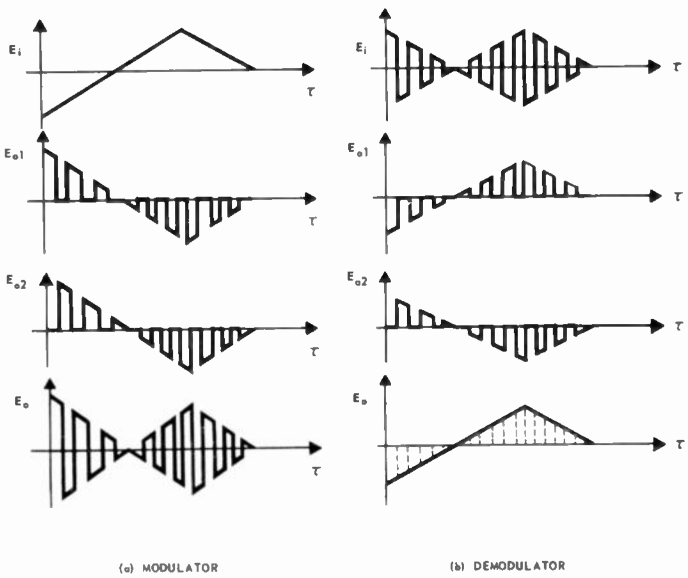

Fig. 6—Feedback switch modulator and demodulator waveforms.

output at E_{o1} contains a component of the input signal E_i. The path through R_{1o} is necessary, therefore, to subtract this component and generate at E_o a balanced suppressed carrier wave for the modulator or a full wave-rectified output signal for the demodulator. Also, E_{o2} is properly phased to maintain a closed loop around the dc amplifier when E_{o1} is zero.

For a modulator developed as shown in Fig. 5, the significant performance characteristics are as follows:

Keying frequency	1000 cps
Full scale output (E_0)	50 volts
Residual output (Principal component—1000 cps)	10 millivolts
Linearity of full scale	0.1 per cent
Equivalent dc drift at the input	1 millivolt
Stability for ± 5 per cent plate supply fluctuation	0.1 per cent
Stability for ± 15 per cent filament supply fluctuation	0.1 per cent

Similar characteristics are obtained for a demodulator which exhibits a quadrature rejection factor of 400 or greater. A linearity curve for a complete resolver chain as used in the Navy Flight Simulator is shown in Fig. 7. The linearity is within a 0.1 per cent of full scale for the complete chain.

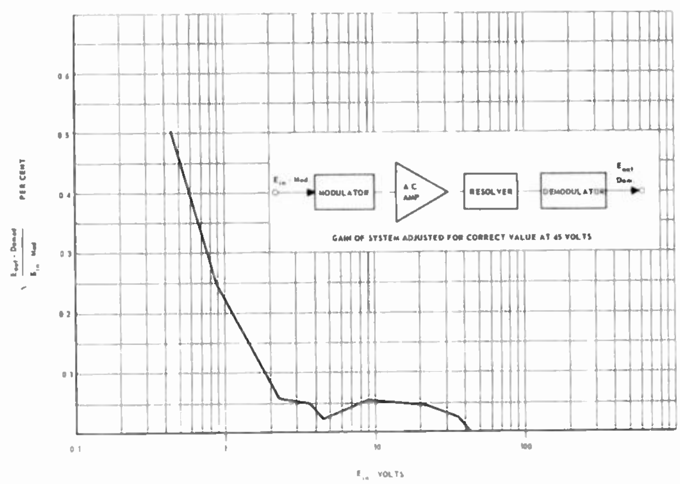

Fig. 7—Linearity of resolver chain.

In the Navy Flight Simulator application, the resolver chains which utilize these modulators and demodulators work efficiently with the square wave sensitivity function[8] developed by overdriving the switching tube grids with a sinusoidal keying voltage. Reference to the waveforms given in Fig. 6 indicates that, for the ideal case, the square wave carrier at the output of the modulator is rectified by the demodulator with very little ripple in the dc output signal. In actual application, the square wave carrier is slightly modified by the ac

[8] M. A. Goldstein, Jr., "Sensitivity-Function Analysis of Modulation Systems with Statistical Inputs," submitted as an S.M. Thesis, Dept. Elect. Engrg., Mass. Inst. Tech., Cambridge, Mass.

amplifier and resolver that are located between the modulator and demodulator in a resolver chain. Thus, a spike occurs at each crossover of the carrier and filtering is required to reduce ripple at the demodulator output to an acceptable level. Using this filter, the resolver chain has a residual ripple of 0.02 volt with 50 volts full scale and a phase shift of two degrees at 150 cycles per second.

A sine wave carrier modulator is required in many applications for satisfactory system operation. Utilizing a circuit similar to the circuit shown in Fig. 5, C. G. Blanyer of the Massachusetts Institute of Technology developed a modulator[9] which essentially has a sine wave carrier output since the total harmonic output at the 400 cps carrier frequency is less than 0.25 per cent. This small harmonic content is obtained with very little phase shift for modulation frequencies up to 100 cps by utilizing two feedback modulators in parallel. One modulator operates at the fundamental carrier frequency while the other operates at three times the fundamental carrier frequency. Thus, a step wave carrier is obtained that has the proper fundamental without any third harmonic content. Conventional filtering techniques are used to remove most of the remaining odd harmonic components. Since the filter is required for components above the third harmonic, the cutoff frequency is higher than the frequency normally used with a square wave modulator and, consequently, the low-frequency phase response is correspondingly improved. Also, a pre-emphasis technique is used to improve the phase response in the pass band of the modulator.

Multiple Input Switches

The multiple input class of switch connects one of two or more input signals to a single output channel as shown by the block diagram in Fig. 8. (This class of switch includes the nonfeedback type normally used with oscilloscopes.) The switch described in the following paragraphs was originally developed as an analog computer relay element[10] for applications that required switching rates much higher than could be accomplished by electromechanical relays.

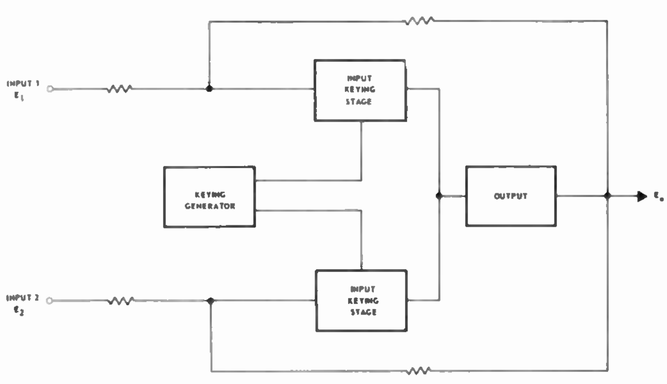

Fig. 8—Multiple input switch block diagram.

A circuit diagram of the original switch, consisting of two switching tubes, V1 and V2, and a common output stage, V3B, is shown in Fig. 9. Triode V3A provides a low impedance bias supply for the screens of tubes V1 and V2. Direct coupling is utilized to minimize switching transient effects.

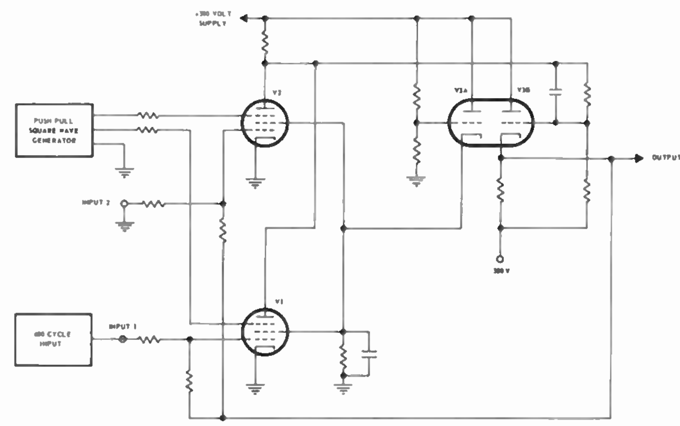

Fig. 9—Original multiple input feedback switch circuit diagram.

With a 20-volt, 400-cps signal applied to tube V1, static measurements were made on the circuit shown in Fig. 9. With tube V2 cutoff, the loop gain is 133, the output 20 volts, the gain accuracy one per cent and the dc output three volts. With tube V1 cutoff, the loop gain is approximately 133, the output 0.09 volt, and the dc output 4.5 volts.

Under dynamic operation, the dc output shift observed in the static test resulted in an output ripple at the switching frequency. The switching characteristics, however, were very satisfactory up to three kilocyles (the limit for the switching source assembled for testing). For the circuit shown in Fig. 9, the approximate dynamic switching characteristics can be predicted by considering the output vs input function for a change in forward loop gain.

The gain with either tube V1 or V2 can be expressed as:

$$A = \frac{KG(s)}{1 + KG(s)\beta}.$$

Considering K as the independent variable:

[9] C. G. Blanyer, "Precision Modulators and Demodulators," presented at the June 23–25, 1954 meeting of the Association for Computing Machinery, based on an S.M. Thesis submitted to the Dept. Elect. Engrg., M.I.T.

[10] Patent applied for by the Navy Dept., Office of Naval Res., M.I.T. Contract NOrd-9661.

$$A = \frac{1}{\dfrac{1}{KG(s)} + \beta}$$

and,

$$\frac{1}{A} = \frac{1}{KG(s)} + \beta.$$

Let,

$$G(s) = \frac{1}{(T_1 s + 1)} \qquad \begin{aligned} K &= 0 \text{ at } 0 > t \\ &= K' \text{ at } t > 0, \end{aligned}$$

then,

$$KG(t) = K'[1 - e^{-\frac{t}{T_1}}].$$

If the equivalent time constant T_e of the switched amplifier is considered to be the time at which $A = 0.622/\beta$, $KG(t)$ can be solved for $A = 0.622/\beta$. Thus,

$$\frac{1}{A} = \frac{1}{KG(t)} + \beta$$

$$\frac{1}{KG(t)} = \beta \frac{1}{0.622} - 1$$

$$KG(t) = \frac{1.65}{\beta}.$$

Solving for t when $KG(t) = 1.65/\beta$

$$KG(t) = \frac{K'}{T'} t.$$

Since $K' \gg 1$, only the initial slope of $KG(t)$ need be considered. Hence,

$$\frac{1.65}{\beta} = \frac{K'}{T_1} T_e$$

$$T_e = \frac{1.65 T_1}{\beta K'}.$$

For the amplifier shown in Fig. 9,

$$K' = 100$$

$$T_1 = 5 \ \mu \text{ sec (20-kilocycle approximate cutoff)}$$

$$\beta = 1.$$

Then, $T_e = 0.0825$ μsec.

Thus, with proper design, it should be possible to provide a unit which is capable of a switching rate in excess of 500 kilocyles. The double input arrangement is essential even when only one input is active because the shift in the quiescent operating point is minimized by the second input and can be made negligible by using high loop gain and balanced input tubes. Also, by operating the tube which is cutoff into the internal loop of a feedback amplifier, the discrimination against the switched-off signal is greatly increased as compared with the discrimination of a single sided circuit.

Based on the principles contained in the circuit shown in Fig. 9, a demodulator using a multiple input switched-feedback amplifier was developed by Bendix personnel. A block diagram of the demodulator, involving the summation of the outputs of two feedback amplifiers that are alternately keyed on and off by the carrier signal, is shown in Fig. 10. The amplifier in channel A

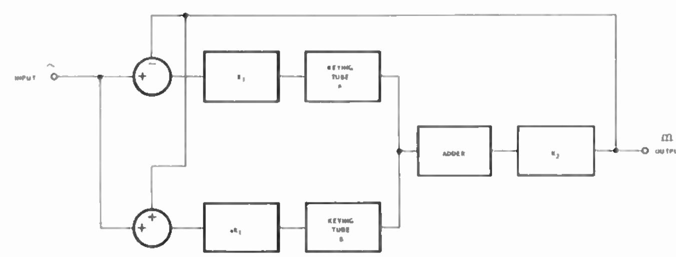

Fig. 10—Dual-input switched-feedback demodulator block diagram.

has a gain of plus one while the amplifier in channel B has a gain of minus one. Hence, if channel A is switched on when the input signal is plus, a positive signal appears at the output. If channel B is switched on as the input signal goes negative, a positive signal continues to appear at the output since channel B has a gain of minus one. The unit therefore operates as a rectifier which is phase sensitive since a change in the relation between the input polarity and the switching sequence results in a corresponding change in the output polarity.

A schematic diagram of the demodulator based on the block diagram of Fig. 10 is given in Fig. 11 on the following page. Tubes V1 and V3 comprise the input keying stage for the A channel which yields a positive output with a positive input. Tubes V2 and V4 comprise the input keying stage for the B channel which yields a negative output with a positive input. The two channels are summed at the plates of tubes V3 and V4. Tube V6 serves as the output stage. Both channels are designed to be stable high loop gain feedback amplifiers.

Using a 400 cps carrier with a 40-volt dc full scale output, the demodulator shown in Fig. 11 has the following performance characteristics.

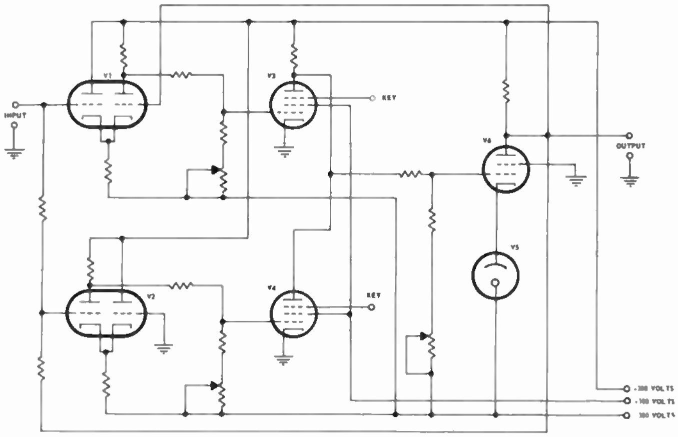

Fig. 11—Dual-input switched-feedback demodulator schematic diagram.

Linearity (25-volt full scale output)	±0.1 per cent of full scale
Noise rms output (zero input)	0.25 per cent of full scale
Zero output stability (±3 per cent plate supply variation)	±0.005 per cent of full scale
Zero output stability (±5 per cent filament supply variation)	0.05 per cent of full scale
Gain stability (±5 per cent supply variation)	0.002 per cent of full scale
Quadrature rejection (1 volt output) Minimum 2:1	
Quadrature rejection (30 volts output) Maximum 30:1	

Essentially the same operating characteristics are obtained for this demodulator at carrier frequencies as high as 10,000 cps. At frequencies above 1000 cps, however, the maximum available output is reduced because of saturation in the output stage. Reference should be made to Fig. 12 for complete dual-input demodulator linearity data.

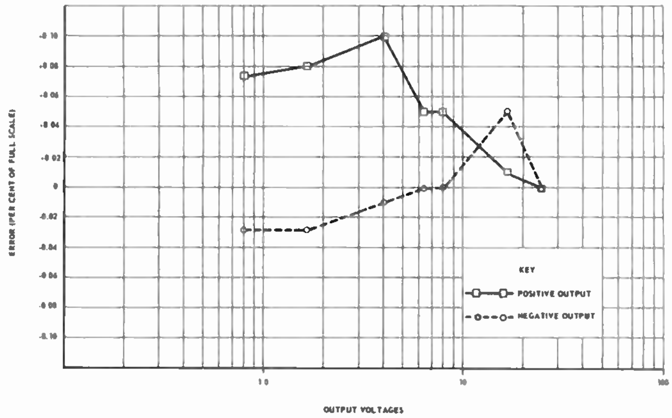

Fig. 12—Dual-input demodulator linearity.

Conclusion

The circuits described show the practicability of developing electronic switches which have characteristics suitable for precision signal transmission switching. The undesirable characteristics of switching tubes can be minimized by the use of high loop gain feedback amplifiers. Thus, the nonlinear computing functions of modulation, demodulation, and multiplication can be performed with approximately the same accuracy as obtained with linear operations such as summation and integration. It is also evident that these circuits can be used in digital-to-analog conversion applications, signal comparison test devices, and communication switching schemes.

Acknowledgment

A number of staff members contributed to the development of the circuits described. Particular mention should be made of P. F. Fischer, R. A. Wilson, and W. J. Chalmers.

Correspondence

Special Case of a Bridge Equivalent of Brune Networks*

The general equivalence of a bridge network to a Brune network terminated in a resistor has been given by Reza.[1] For a biquadratic admittance function, Reza's bridge network contains eight elements, although he suggests that in special cases five elements can represent such a function. Two five-element bridge networks which are an interesting special case in representing a biquadratic admittance function are shown in Fig. 1(a) and 1(b). These networks are

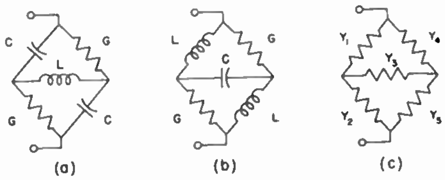

Fig. 1

two of a number recently studied by Kim;[2] this letter points out additional features of these networks.

For the biquadratic admittance function

$$Y(s) = H\frac{s^2 + a_1 s + a_0}{s^2 + b_1 s + b_0} \tag{1}$$

to be positive real and minimum, that is with $\operatorname{Re} Y(j\omega_1) = 0 \pm jB(\omega_1)$ at one ω_1 with $B(\omega_1) \neq 0$, it is necessary and sufficient that

$$a_1 b_1 = (\sqrt{a_0} - \sqrt{b_0})^2. \tag{2}$$

The admittance functions for the networks of Fig. 1(a) and 1(b) have the form of a quotient of a third-order to third-order polynomial. However, with the particular arrangement of elements with equal R's and C's in Fig. 1(a) and equal R's and L's in Fig. 1(b), a root of the numerator is always equal to a root of the denominator. This being the case, the admittance functions for the two networks are biquadratic functions having the following forms.

$$Y_a(s) = 2G\frac{s^2 + \dfrac{k}{2LG}s + \dfrac{1}{2LC}}{s^2 + \dfrac{G}{kC}s + \dfrac{2}{LC}} \tag{3}$$

$$Y_b(s) = \frac{G}{2}\,\frac{s^2 + \dfrac{k}{LG}s + \dfrac{2}{LC}}{s^2 + \dfrac{G}{2kC}s + \dfrac{1}{2LC}} \tag{4}$$

The functions Y_a and Y_b satisfy (2) for all R, L, and C and for both networks the real part of $Y(j\omega)$ vanishes at the frequency

$$\omega_1 = \frac{1}{\sqrt{LC}}. \tag{5}$$

The susceptance functions have the following values at the frequency ω_1.

$$B_a(\omega_1) = +k\sqrt{\frac{C}{L}} \quad \text{and}$$

$$B_b(\omega_1) = -k\sqrt{\frac{C}{L}}. \tag{6}$$

This result suggests that the two bridge networks are canonical forms of networks

$$Y = \frac{y_1y_2y_3 + y_1y_2y_4 + y_1y_2y_5 + y_1y_3y_5 + y_1y_4y_5 + y_2y_3y_4 + y_2y_4y_5 + y_3y_4y_5}{y_1y_3 + y_1y_4 + y_1y_5 + y_2y_3 + y_2y_4 + y_2y_5 + y_3y_4 + y_3y_5}. \tag{10}$$

and that network a may be used when specifications require that $B(\omega_1)$ be positive and network b when negative. The admittances of the two networks at $\omega = 0$ and $\omega = \infty$ are finite and nonzero as required of minimum functions and have the following values.

$$Y_a(0) = Y_b(\infty) = \frac{G}{2} \quad \text{and}$$

$$Y_a(\infty) = Y_b(0) = 2G. \tag{7}$$

For a biquadratic function of the form of (1) to be realizable in the form of one of the networks of Figs. 1(a) and 1(b), it is necessary that the coefficients be related as follows.

$$\frac{b_0}{a_0} = 4 \text{ (network } a\text{)},$$

$$\frac{b_0}{a_0} = \frac{1}{4} \text{ (network } b\text{)}. \tag{8}$$

For both networks, it is necessary that

$$2a_1 b_1 = \frac{1}{LC} = \omega_1^2. \tag{9}$$

When these conditions are fulfilled, then the network elements are simply related to the coefficients of (1).

Network a	Network b
$G = \dfrac{H}{2}$	$G = 2H$
$\dfrac{L}{k} = \dfrac{1}{a_1 H}$	$\dfrac{L}{k} = \dfrac{1}{2a_1 H}$
$kC = \dfrac{a_1 H}{2a_0}$	$kC = \dfrac{4a_1 H}{a_0}$

The limitations placed on the coefficients of $Y(s)$ by (8) and (9) mean that only a special class of biquadratic functions can be realized by the five-element bridge networks of Fig. 1. These limitations are also evident in the restrictions in pole and zero locations. If the zeros of $Y(s)$ are complex and frequency is scaled such that the zeros lie on a unit circle in the s plane, then the poles, if complex, must lie on a circle of radius 2 for network a and on a circle of radius $\frac{1}{2}$ for network b. Similar restrictions apply to the other possibilities for real or complex poles and zeros.

It is interesting to point out that the operation of the bridge network in fulfilling the minimum admittance function requirements cannot be visualized in terms of series or parallel resonance of elements as in the methods of Brune, Bott and Duffin, and Pantell. The driving-point admittance for the general bridge network of Fig. 1(c) is

The manner in which the real parts of the eight triple-product terms add to zero at ω_1 is difficult to picture.

Eq. (10) also points out the difficulty in finding other five-element bridge networks suitable for the synthesis of biquadratic minimum functions. It is necessary that three reactive elements be used for $Y(s)$ to be a potentially minimum function.[2] The common root in the numerator and denominator of $Y(s)$ must be found. Finally, it is necessary that $Y(s)$ satisfy (2). Similar analysis for higher ordered functions is even more difficult since more than five elements must be used in the bridge network except in special cases.

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* Received by the IRE, July 16, 1956. This work was supported by the Office of Ordnance Research, U.S. Army.

[1] F. M. Reza, "A bridge equivalent for a Brune cycle terminated in a resistor," Proc. IRE, vol. 42, p. 1321; August, 1954.

[2] W. H. Kim, "A new method of driving-point function synthesis," Rept. No. 1, Contract DA-11-022-ORD-1983, Elect. Engrg. Res. Lab., Univ. of Ill., April 1, 1956.

Useful Bandwidth in Scatter Transmission*

I have read with interest the numerous papers published recently in the Proceedings of the IRE and Transactions of the IRE on propagation by scattering, and I take the liberty of drawing your attention to some of my own publications related to subjects discussed in these papers.

1) In 1953, I proposed to define a useful bandwidth, for transmission by tropospheric scattering, on the basis of the correlation existing between the fluctuations of the received field strength on adjacent frequencies.[1] Some time later, in 1955, Gordon[2] and

* Received by the IRE, June 25, 1956.

[1] J. P. Voge, "Note relative à la bande de fréquences utilisable pour des transmissions en ondes ultra-courtes," *Ann. Télécommun.*, vol. 8, pp. 308–311; August–September, 1953.

[2] William Gordon, "Radio scattering in the troposphere," Proc. IRE, vol. 43, pp. 23–28; January, 1955.

Booker and de Bettencourt[3] proposed another definition of this bandwidth, based on propagation time differences, due to the existence of multiple propagation modes. However, quite recently, Staras,[4] Norton,[5] and others,[6] again proposed to define this useful bandwidth on a correlation basis.

Having used the results of a calculation made by Rice,[7] I have been led[1] to a bandwidth independent of the frequency, and inversely proportional to the size of the scattering volume and to the sine of the half of the scattering angle. For a transmission distance of 315 km (from Wrotham to Bagneux), I have found a bandwidth of 100 kc (with a correlation coefficient greater than 0.9) and of 300 kc (with a correlation coefficient greater than 0.4), which is in good agreement with the results reported by Staras and by Norton (bandwidth equal to 100 kc for a transmission path of 226 miles and a correlation coefficient of 0.9).

Staras has stated that "no one has as yet analyzed the detailed implication of a 0.5 correlation on different types of modulation systems." May I remark that the case of fm transmissions has been studied by P. Clavier,[8] who has been able to determine the distortion and diaphony corresponding to a bandwidth of a given correlation magnitude. Thus, for instance, for the above mentioned transmission path of 315 km, the diaphonies of the second and third orders would reach respectively 36 and 50 db for a multiplex link of 12 channels extending from 12 to 60 kc, with a frequency shift of ± 150 kc. In the case of a multiplex link of 12 channels extending from 60 to 108 kc, the frequency shift being ± 250 kc, the diaphony would be practically negligible on the second order and reach 36 db on the third.

On the other hand, in my above mentioned paper,[1] I have come to definite conclusions which have been adopted and developed by Booker and de Bettencourt[3]; *e.g.*, the use of highly directive antennas, with rather narrow beams capable of intercepting a fraction of the scattering volume, permits the widening of the useful bandwidth, but huge antennas give quite a noticeable loss of gain.

Finally, the size of the scattering volume which I found for the 315 km path (3.6 km) is in good agreement with the results of the theory proposed by Gordon (height of the volume: 1.4 km; width: 3.6 km, for a 315 km path and a standard earth radius of 8500 km).

The same method, when applied to ionospheric scattering, leads[9] to bandwidths of a few kilocycles. Experimental results reported by Bailey[10] correspond fairly well to these theoretical predictions.

2) In a comparative study of the theories of Booker and Gordon, Megaw, and Villars and Weisskopf, which I published,[11] I was led to the following formula for the derivation of the scattered power from the turbulence spectrum:

$$\sigma = -\frac{2\pi^4}{\lambda^4}\left|\frac{\Delta\epsilon}{\epsilon_0}\right|^2 \frac{1}{k}\frac{dF(k)}{dk}\sin^2\chi \qquad (1)$$

where, with the usual symbols: θ is the scattering angle, and ϵ, the dielectric constant:

$$k = \frac{4\pi}{\lambda}\sin\frac{\theta}{2};$$

σ is the scattering cross section, $F(k)$ is the spectral density and χ is the angle between the electric field vector at the scattering point and the direction of the receiver.

For a spectral region where $F(k) \propto k^{-n}$, σ is proportional to

$$\lambda^{n-2}\left(\sin\frac{\theta}{2}\right)^{-(n+2)}.$$

If l_s is the smallest blob size of the turbulence spectrum, we have: $n=2$ after Booker and Gordon; $n=5/3$ and 7 (if $k \gg \pi/l_s$) after Megaw; $n=7/3$ and 9 (if $k \gg \pi/l_s$) after Villars and Weisskopf[12] (1954); $n=3$ after Villars and Weisskopf (1955).[12] Thus (1) permits a simple study of the phenomenon, whatever the turbulence spectrum one considers, and Norton started with this same formula in his latest theory.[5]

This formula has further enabled me[11] to appreciate the influence on the expression of the scattered power of the upper part, (corresponding to $k \gg \pi/l_s$) the less known one, of the turbulence spectrum. This influence appears to be negligible in the case of tropospheric scattering, but is very important in that of ionospheric scattering. Wheelon has published similar conclusions.[13] It is also easy to show that the various spectra proposed (or at least the first three mentioned above) lead to almost the same results, in the case of tropospheric scattering.

In his paper,[13] Wheelon suggests a modification to the exponential correlation law (for the fluctuations of the dielectric constant) of Booker and Gordon, which should include effects of the smallest blobs in the turbulent spectrum. By the way, I have to draw attention to the fact that this spectrum corresponds, if $kk \gg (\pi/L_s)$, to $F(k) \propto k^{-4}$, $n=4$.

I have proposed a similar modification myself,[14] with a correlation function which would read, with Wheelon's notation

$$C(R) = e - \frac{R^2}{l_0\sqrt{R^2 + l_s^2}}. \qquad (2)$$

For $R \ll l_s$, as for $R \gg l_s$, both correlation functions thus modified are equivalent and differ only in the transition zone (R of the order of l_s). I have been led to (2) after having found by calculation the formulas given without demonstration by Megaw[15] and expressing the fluctuations of the received field strength, on paths in the line of sight. For radio waves, I could use the correlation law and the method of Booker and Gordon, but for shorter wavelengths (waves of light), I had to use geometrical optics and to modify the exponential correlation law (or to use a different spectrum, such as the Kolmogoroff—Megaw spectrum).

3) Gordon has calculated[2] the scattered power and the useful bandwidth by an approximate method, with the hypothesis of an exponential correlation law and of distribution of the refractive index fluctuations inversely proportional to the altitude.

In order to evaluate the error thus introduced, I made the calculation anew, with a exact integration and have obtained the following results[11]: the factor 2.45 has to be replaced by 3.4 in (8) in Gordon's work, for the scattered power; the factor 30 has to be replaced by 18 in (19) giving the bandwidth as a function of the distance (this, for 300 km gives 2.75 mc instead of 4.6 mc). Booker and de Bettencourt, after a first correction,[3] had obtained 3.3 mc for 300 km.

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Laboratoire National
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Paris, France

[3] H. Booker, and J. T. deBettencourt, "Theory of radio transmission by tropospheric scattering using very narrow beams," Proc. IRE, vol. 43, pp. 281–290; March, 1955.

[4] Harold Staras, "Forward scattering of radio waves by anisotropic turbulence," Proc. IRE, vol. 43, pp. 1374–1380; October, 1955.

[5] K. A. Norton, "Point-to-point radio relaying via the scatter mode of tropospheric propagation," Trans. IRE, vol. CS-4; pp. 39–49; March, 1956. See pp. 42, 47.

[6] A. P. Barsis, et al., "The Cheyenne Mountain tropospheric propagation experiments," NBS Circular 554; January 3, 1955.

[7] S. O. Rice, "Statistical fluctuations of radio field strength far beyond the horizon," Proc. IRE, vol. 41, pp. 274–281; February, 1953.

[8] P. Clavier, note technique. (private publication.) Compagnie Francaise Thomson-Houston; March 29, 1954; "Calcul de la diaphonie dans une transmission multiplex en modulation de frequence en propagation par diffusion tropospherique," *C. R. Groupe d'etude de la Propagation*, February–June, 1955.

[9] J. P. Voge, "Problèmes d'actualité dans l'étude de la transmission des ondes ultra-courtes," *Onde élect.*, vol. 34, p. 488; June, 1954.

[10] D. K. Bailey, R. Bateman, and R. C. Kirby, "Radio transmission at vhf by scattering and other processes in the lower ionosphere," Proc. IRE, vol. 43, pp. 1181–1230; October, 1955. See p. 1225.

[11] J. P. Voge, "Radioelectricité et troposphere," *Onde élect.*, vol. 35, pp. 565–575; June, 1955.

[12] F. Villars and V. F. Weisskopf, "The scattering of electromagnetic waves by turbulent atmospheric fluctuations," *Phys. Rev.*, vol. 94, p. 232; April, 1954. "On the scattering of radio waves by turbulent fluctuations of the atmosphere," Proc. IRE, vol. 43, pp. 1232–1239; October, 1955.

[13] A. H. Wheelon, "Note on scatter propagation with a modified exponential correlation," Proc. IRE, vol. 43, pp. 1381–1383; October, 1955.

[14] J. P. Voge, "Fluctuation du champ électromagnetrique dues à la turbulence à l'extremité d'un trajet de propagation en visibilité directe," *C. R. Acad. Sci. Paris*, vol. 237, pp. 351–353; July 27, 1953.

[15] E. C. S. Megaw, "Waves and fluctuations," *Proc. IEE*, vol. 100, pt. III, pp. 1–8; January, 1953. See p. 5.

Russian Resistance and Resistor Terminology*

The Russian word for both resistance and resistor is сопротивление (literally, *opposition*); in translation, some interpretation is generally needed to decide which is meant. Although резистор and резистер are given in some technical dictionaries, they are very seldom used in the literature.

The terms полное or кажущееся сопротивление (impedance) and реактивное сопротивление (reactance), so confusing to the English reader, follow German terminology (*cf. Scheinwiderstand, Blindwiderstand, Widerstand*). Again, the English-French cognates of импеданс, реактанс and резистанс are only occasionally encountered in the literature.

* Received by the IRE, May 25, 1956.

RESISTANCE—Сопротивление

acoustic r	акустическое с.
alternating-current r	с. переменному току
antenna r	с. антенны
apparent r (impedance)	кажущееся с.
blocked r	блокированное с.
capacitive reactance	емкостное с.
circuit r	с. цепи (контура, схемы)
combined r	комбинированное с.
contact r	1 с. контакта
	2 контактное с.
	3 переходное с.
coupling r	с. связи
critical r	критическое с.
direct-current r	с. постоянному току
dynamic r	динамическое с.
effective (active) r	активное с.
effective (watt) r	ваттное с.
effective r	действующее с.
electrical r	электрическое с.
electrode r	электродное с.
electrolytic r	электролитическое с.
equivalent r	эквивалентное с.
external r	внешнее с.
filament r	с. нити накала
fixed r	с. постоянной величины
full r (impedance)	полное с.
ground r	с. заземленной цепи
high r	большое с.
high-frequency r	с. токам в. ч.
inductive reactance	индтктивное с.
internal r	внутреннее с.
joint r	сложное с.
load r	нагрузочное с.
low r	малое с.
magnetic r (reluctance)	магнитное с.
mechanical r	механическое с.
negative r	отрицательное с.
ohmic r	омическое с.
parallel r	параллельное с.
potentiometer r	с. потенциометра
radiation r	с. излучения
radio-frequency r	радиочастотное с.
reactance	реактивное с.
reflected r	отраженное с.
regulating r	регулировочное с.
relatively high r	относительно большое с.
resistance box	магазин сопротивлений
resistivity	сопротивляемость
shunt r	с. шунта
specific r	удельное с.
stabilizing r	стабилизующее с.
total r	общее с.
total (complex) r	комплексное с.
useful r	полезное с.
winding r	с. намотки

RESISTOR—Сопротивление

adjustable r	регулируемое с
ballast r	балластное с
ballast tube	балластная лампа
barretter	барреттер
bias r	с. смещения
biasing r	смещающее с
carbon r	угольное с
cathode r	с. катода
center-tapped r	с. с отводом посредине
ceramic r	керамическое с
composition (chemical) r	химическое с
dropping r	падающее с
experimental r	экспериментальное с
filament-coated r	с. с проводящим слоем нанесенным на стеклянную нить
fixed r	постоянное с
flexible r	гибкое с
insulated r	изолированное с
low-power r	1) маломощное с
	2) с. малой мощности
medium power r	1) среднемощное с
	2) с. средней мощности
metallized r	металлированное с
metallized-filament r	с. с металлизированной стеклянной нитью
molded r	1) лепное с
	2) прессованное с
noninductive r	безындукционное с
plug-in r	сменное с
power r	мощное с
precision r	точное с
protective r	предохранительное с
radial-lead r	с. с радиальными проводами
solid-body r	массивное с
standard r	эталонное с
suppressor	1) заглушающее с
	2) уничтожитель
tape-wound r	с. в виде тесьмы
tapped r	с. с отводами
tapped (sectionalized) r	секционированное с
thermal r	термическое с
thermistor	термосопротивление
type r	типовое с
typical r	типичное с
variable r	переменное с
wire-wound r	проволочное с

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Contributors

R. Bechmann

Rudolf Bechmann (SM'54) was born in Nuremberg, Germany, on July 22, 1902. He received the Ph.D. degree in theoretical physics in 1927 from the University of Munich.

From 1927 to 1945 Dr. Bechmann was employed by Telefunken Company for Wireless Telegraphy, Ltd., Berlin. He was at first concerned with antenna problems, especially with questions of radiation resistance and radiation characteristics of composite antennas. In 1931 he developed the so-called emf method. Later Dr. Bechmann turned his full attention to piezoelectric quartz crystals. In 1933 he discovered, independently, several quartz cuts having zero frequency temperature coefficients—the *AT*-, *BT*-, *CT*-, and *DT*-type resonators. He has made many contributions to the field of elasticity and piezoelectricity and its application to quartz. By joining the production of oscillators and resonators to his scientific laboratory activities, Dr. Bechmann became involved in questions related to quartz crystals. During World War II he directed, in addition, to his specialized activities with Telefunken, several agencies covering the quartz industry as a whole.

After the war he joined the Oberspree Company in Berlin and directed the company from 1946 to 1948.

Moving to England in 1948, he was Principal Scientific Officer at the British Post Office Research Station, Dollis Hill, London. Here he studied the properties of several water-soluble piezoelectric materials, and developed methods for determining the elastic and piezoelectric constants, using the resonance method applied to various modes of plates.

In 1953 he came to the Clevite Research Center, Cleveland, Ohio, at that time the Brush Laboratories Company, as head of the Dielectric Phenomena Section of the Electrophysical Research Department. He extended his studies on methods of determining these constants into the field of ferroelectric ceramics. His chief activity, however, was the investigation of properties of synthetic quartz resonators.

In 1956 he joined the Signal Corps Engineering Laboratories, Fort Monmouth, N. J. as consultant physicist.

Dr. Bechmann is a member of the IRE Piezoelectric Crystals Committee, and a Fellow of the American Physical Society.

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G. Bruun

Georg Bruun (A'46–SM'56) was born in Næstved, Denmark, on October 13, 1916. He received the M.S. degree in telecommunications engineering from The Royal Technical University of Denmark in 1941. After graduation he was employed as an engineer in the radio development division of the Royal Danish Navy. In 1943–44 he was research assistant in telecommunications at the Royal Technical University of Denmark. Since 1944 he has been director of the Radio Receiver Research Laboratory, the Academy of Technical Sciences, Copenhagen. This laboratory is mainly engaged in research and development in the field of AM and fm receivers and television. During the period 1949–51, and since 1954 he has taught telecommunications at the Royal Technical University. From September, 1954 to August, 1955 he worked at Electronics Research Laboratory, Stanford University and Stanford Research Institute, on a research fellowship granted by the National Academy of Sciences in Washington. He was engaged in research and development work concerned with transistor circuitry.

He is author of several technical papers and co-author of a textbook on radio measurements.

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M. Cohn

Marvin Cohn (S'49–A'51) was born in Chicago, Ill., on September 25, 1928. He received the B.S.E.E. degree in 1950 and the M.S.E.E. degree in 1953, both from the Illinois Institute of Technology.

From 1951 to 1952, Mr. Cohn was employed by the Glenn L. Martin Company, Baltimore, Md.; he was with the Radiation Laboratory from 1952 until he entered the U. S. Army Signal Corps in 1953. In 1955 he returned to the Radiation Laboratory where he is doing research and development work on broadband superheterodyne receivers for the microwave bands.

Mr. Cohn is a member of Eta Kappa Nu and Tau Beta Pi.

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C. M. Edwards

Charles M. Edwards (S'41–A'43–M'45–SM'53) was born October 18, 1917, in Centralia, Ill. In 1941, he received the B.S. degree in electrical engineering from Massachusetts Institute of Technology, Cambridge, Mass., and the M.S. degree in electrical engineering at the same time. From 1939 to 1946, Mr. Edwards was associated with the Bell Telephone Laboratories, New York, N. Y., the American Telephone and Telegraph Company, Princeton, N. J., and the Western Electric Company, Kearny, N. J. He was employed as a research engineer at M.I.T. from 1946 to 1951. Most of his work there concerned the development of a large scale analog computer known as the Dynamic Analysis and Control Laboratory Flight Simulator. In 1951, he joined the Research Laboratories Division of the Bendix Aviation Corp., Detroit, Mich., where he is head of the computer department.

Mr. Edwards is a member of the Engineering Society of Detroit, Eta Kappa Nu, and Sigma Xi.

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D. W. Fraser

Donald W. Fraser (M'53–SM'55) was born on May 22, 1910. He attended the United States Naval Academy from which he received the B.S. degree in 1934. He did advanced work at the naval preradar and radar schools at Harvard and M.I.T. In 1948, he received the M.S. degree in electrical engineering from Georgia Tech and in 1955, he received the Ph.D. degree in electrical engineering.

During World War II, Mr. Fraser served with the Electronic Field Service Group of the U. S. Navy from June, 1942 until September, 1946. In the Korean War he became an electronics officer on Staff Commander Operational Development Force in the Navy, serving with this group from September, 1950 until January, 1953.

From 1946 until 1950 Mr. Fraser was assistant professor of electrical engineering and research associate at the Georgia Institute of Technology in Atlanta, where he did research in high-frequency oscillators. He was also research engineer at the engineering experiment station of Georgia Tech from 1953 to 1955. Mr. Fraser was director of projects on frequency control. At the present time he is the head of the department of electrical engineering at the University of Rhode Island at Kingston.

Mr. Fraser is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

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Edward G. Holmes (M'49) was born on February 19, 1923. In 1944, he received the B.E. degree in electrical engineering from

Tulane University in New Orleans, La. He also attended Georgia Tech where he received the M.S. degree in electrical engineering in 1953.

E. G. Holmes

Mr. Holmes served with the navy during World War II from 1944 to 1946, in the capacity of radar material officer. After his return to civilian life he became chief engineer of radio station WTPS in New Orleans, staying with them until 1948. From this position he went to Earl Lipscomb Associates where he worked as applications engineer until 1951. From 1951 until 1955 he was research engineer at the Engineering Experiment Station of Georgia Tech. He also did part-time teaching in the electrical engineering department. His experience has been in short-pulse modulation, microwave, antennas, uhf techniques and pulse transformers. At the present time Mr. Holmes is manager of Southeastern Industrial Instruments—an engineering representative and consulting firm.

Mr. Holmes is a Registered Engineer in the state of Texas and a member of Eta Kappa Nu.

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William Connor King (SM'56) was born March 10, 1927, in Granville, Ohio. He received the B.A. degree in physics from Denison University in 1949 and the Ph.D. degree in physics from Duke University in 1953. From 1944 to 1948 he served in the Armed Forces.

W. C. King

In 1953, Dr. King became a research associate with the Radiation Laboratory, where he was a project leader in charge of design and development of a microwave crystal-video receiver, including antennas, filters, detectors, amplifiers, and display equipment.

In 1956, he joined the staff of the Aerosciences Laboratory, Special Defense Projects Department of General Electric Company, as a propagation specialist.

He is a member of the American Physical Society, American Association of Physics Teachers, and Sigma Xi.

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Kam Li was born in 1927, in Canton, China. He was educated at the Chiao-tung University in Shanghai, where he received the B.S. degree in 1949. In 1951 and in 1955 respectively he received the M.S. degree and the Ph.D. degree in electrical engineering from the University of Pennsylvania.

Since 1951 Dr. Li has been associated with the electromedical group of the Moore School of Electrical Engineering in Philadelphia, and the Department of Physical Medicine, Graduate School of Medicine, University of Pennsylvania. His work has been concerned with electrical properties and absorption of electromagnetic energy of biological substances.

K. Li

Dr. Li is a member of Sigma Xi.

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For a biography and photograph of R. G. Medhurst, see page 265 of the February, 1956 issue of Proceedings of the IRE.

❖

Gaston Salmet was born on September 23, 1921 in Paris, France. Since 1941, he has been employed at the Société des Télécommunications Radioélectriques et Téléphoniques in Paris. His position there has been as a research engineer. Since 1954, Mr. Salmet has been chief of the Mobile Telecommunication Sets Laboratory. In his research work he has been concerned with multichannel transceivers, master oscillator units, fm broadcasting transmitters, and electronic controlled tuning.

G. Salmet

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Herman P. Schwan (M'53–SM'55) was born in 1915 in Germany. He studied physics, electrical engineering and biophysics in Goettingen and Frankfurt and spent two years in industry as an electrical engineer (Siemens Telefunken). He received the Ph.D. degree in physics and biophysics in 1940 and 1946 respectively, from the University of Frankfurt and was engaged in biophysical research and ultrahigh frequency development work from 1938 to 1947 at the Kaiser-Wilhelm-Institute at Frankfurt. From 1946 to 1947, he held positions as assistant director and assistant professor at the same institute.

H. P. Schwan

Dr. Schwan came to this country in 1947 and worked for the United States Navy's Aero-Medical Equipment Laboratory as a research specialist. Since 1950, he has been with the University of Pennsylvania and holds appointments as Associate Professor of Physical Medicine and Physics in Medicine in the Graduate School and School of Medicine, and as Associate Professor of Electrical Engineering in the Moore School of Electrical Engineering. He heads the electromedical research team which has been organized at the University of Pennsylvania by the Electrical Engineering and Medical Schools, and is conducting research in the fields of biophysics and medical electronics.

He is a member of the American Association for the Advancement of Science, the Physical Society, the New York Academy of Science, the AIEE, and Sigma Xi.

❖

G. F. Small was born on November 23, 1923 in London, England. In 1944 he received the B.Sc. degree in engineering from London University. From 1944 to 1946 he worked for Standard Telephones and Cables Ltd. on the design of coaxial cables for telephony.

G. F. Small

Since 1946 Mr. Small has been on the staff of the Research Laboratories of the General Electric Company, Ltd. of England in Wembley. He is concerned with the development of microwave components and the general design of radio relay systems for television and multichannel telephony.

Mr. Small is an associate member of the Institute of Electrical Engineers.

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Mary N. Torrey was born on February 2, 1910, in Worcester, Mass. She received the B.A. in mathematics and physics from Wellesley College in 1930 and the M.A. in mathematical statistics from Columbia University in 1946. Since July, 1930, she has been a member of the Quality Assurance Department of Bell Telephone Laboratories. During that time she has done mathematical and statistical work for H. F. Dodge on quality assurance, statistical quality control, sampling inspection, and quality rating problems. She is a joint author with Mr. Dodge of two papers on continuous sampling, and check inspection plans. She also assisted in preparation of "Sampling Inspection Tables," by H. F. Dodge and H. G. Romig, the *ASTM Manual on Quality Control of Materials* and American War Standards Z1.1, Z1.2, and Z1.3 on Quality Control published by American Standards Association.

M. N. Torrey

She is a member of the Institute of Mathematical Statistics, American Statistical Association, Biometric Society, and a Fellow of the American Society for Quality Control.

IRE News and Radio Notes

VLF Symposium Releases List of its Chairmen and Papers

The National Bureau of Standards and the IRE Professional Group on Antennas and Propagation will jointly sponsor a symposium on very-low-frequency propagation at Boulder, Colorado, January 23–25, 1957. Persons wishing to attend this symposium should notify Mrs. M. Halter, National Bureau of Standards, Boulder, Colorado, as soon as possible.

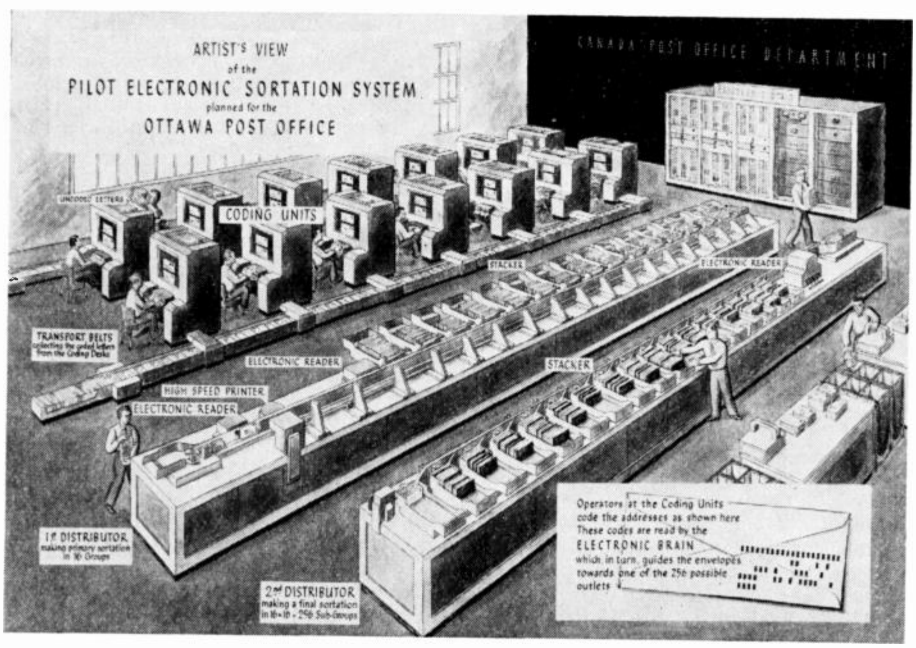
Committee chairmen for this symposium are: J. R. Wait, Steering Committee; R. Silberstein, Local Arrangements; J. R. Johler, Finance; C. H. Bragaw, Publicity; T. N. Gautier and R. A. Helliwell, Panel Discussions; Technical Papeis, J. M. Watts and J. R. Wait. F. W. Brown, K. A. Norton, and R. J. Slutz are on the advisory staff.

Contributions will still be accepted if they are considered suitable. The contributed papers will be reproduced for a symposium record before the meeting. It is therefore requested that authors of accepted papers submit a typed copy (single spacing) of their manuscripts on $8\frac{1}{2}'' \times 11''$ bond suitable for photographic reproduction. The length should not be more than ten pages, including diagrams which should be inserted and mounted appropriately on the typed page. The page numbers should be indicated in pencil. The absolute deadline for submission of this material, with no exceptions, is November 30, 1956.

The following papers will be presented at the symposium: *Some Physical Problems in the Generation and Propagation of VLF Radiation*, E. L. Hill, Dept. of Physics, University of Minnesota; *Studies of High Power VLF Antennas*, W. Gustafson and E. Devaney, U. S. Navy Electronics Laboratory, San Diego; *Some Properties and Applications of the Magneto-Ionic Theory at VLF*, R. A. Helliwell, Radio Propagation Laboratory, Stanford University; *The Relation Between Group Delay of a Whistler and the Distribution of Ionization Along the Ray Path*, R. L. Smith, Radio Propagation Laboratory, Stanford University; *Measurement and Interpretation of the Polarization and Angle of Arrival of Whistlers*, J. H. Crary, Radio Propagation Laboratory, Stanford University; *The Effect of the Earth's Magnetic Field on the Transmission and Reflection of VLF Waves at the Lower Edge of the Ionosphere*, Irving Yabroff, Radio Propagation Laboratory, Stanford University; *Records of VLF Hiss at Boulder, Colorado During 1956*, J. M. Watts, National Bureau of Standards, Boulder; *Extra-Terrestrial Origins of VLF Signals*, Roger Gallet, National Bureau of Standards, Boulder; *Extensions to the Geometrical Optics of Sky Wave Propagation at VLF*, J. R. Wait and Anabeth Murphy, National Bureau of Standards, Boulder; and *Wave Guide Mode Calculations for VLF Ionospheric Propagation Including the Influence of Ground Conductivity*, by J. R. Wait and H. H. Howe, National Bureau of Standards, Boulder.

Also *A Study of Signal-Versus-Distance Data at VLF*, J. L. Heritage and S. Weisbrod of Smyth Research Associates and J. E. Bickel of U. S. Navy Electronics Laboratory; *Basic Experimental Studies of the Magnetic Field of Electromagnetic Sources Immersed in a Semi-Infinite Conducting Medium*, M. B. Kraichman, U. S. Naval Ordnance Laboratory, White Oaks, Silver Spring, Maryland; *A Technique for the Rapid Analysis of Whistlers*, J. K. Grierson and L. R. O. Storey, Radio Physics Laboratory, Ottawa, Canada; *A Method to Interpret the Dispersion Curves of Whistlers*, L. R. O. Storey, Radio Physics Laboratory, Ottawa, Canada; *Relations Between the Character of Atmospherics and Their Place of Origin*, J. Chapman and E. T. Pierce, Cavendish Laboratory, Cambridge, England; *Survey of Investigations of VLF Propagation at Cambridge*, K. G. Budden, Cavendish Laboratory, Cambridge, England; *A Study of VLF Ground Wave Propagation in Alaska*, G. M. Stanley, Geophysical Institute, College, Alaska; *The Phase and Group Velocity of the VLF Ground Wave*, J. R. Johler, National Bureau of Standards, Boulder; *Polarization of the Ground Wave of a Radio Atmospheric*, A. W. Sullivan, University of Florida; *Noise Investigation at VLF by the National Bureau of Standards*, W. Q. Crichlow, National Bureau of Standards, Boulder; *Spectrum Analysis of Spherics*, W. Taylor, National Bureau of Standards, Boulder; *Statistical Descriptions of Atmospheric Radio Noise*, A. D. Watt, National Bureau of Standards, Boulder; *On the Polarization of Spherics*, A. G. Jean, National Bureau of Standards, Boulder; *The Effect of Receiver Bandwidth on the Amplitude Distribution of VLF Atmospheric Noise*, F. F. Fulton, Jr., National Bureau of Standards, Boulder; *Our Present State of Knowledge of the Lower Ionosphere*, A. H. Waynick, Ionospheric Research Laboratory, State College, Pa.; *Heavy Ion Effects in Audio-Frequency Propagation*, C. O. Hines, Radio Physics Laboratory, Ottawa; *Some Recent Measurements of Atmospheric Noise in Canada*, C. A. McKerrow, Radio Physics Laboratory, Ottawa; and *Performance and Design Criteria for High Power VLF Antennas*, W. W. Brown, Bureau of Ships, Washington, D. C.

In addition to technical papers, round table discussions will be included in the program. Among participants will be Owen Storey, Ottawa, Canada; K. G. Budden, Cavendish Laboratory, England; R. A. Helliwell, Stanford University; M. M. Newman, Lighting and Transients Institute, Minneapolis, Minnesota; W. Q. Crichlow, J. M. Watts and others of NBS Boulder Laboratories; staff members of the Navy Electronics Laboratory, Stanford University.

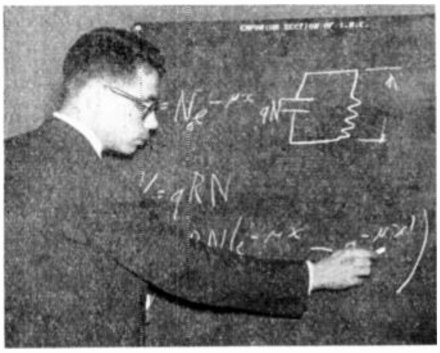


One of the highlights of the recent Canadian IRE Convention, marking the thirtieth anniversary of Canadian IRE activity, was the presentation of papers on the electronic sortation system for mail by W. J. Turnbull, Canadian Deputy Postmaster General, M. Levy of the Ottawa Post Office Department, and C. G. Helwig, H. B. Brown and L. R. Wood of Ferranti Electric Ltd., Toronto. This system, serviced by one technician, can be operated at a speed of ten letters per second. Over 130 exhibits and 132 papers in 26 sessions were presented during the three-day convention.

Rio de Janeiro Section Formed

On August 22, the IRE Board of Directors approved the establishment of the Rio de Janeiro (Brazil) Section. This is the second Section to be established in South America, and the sixteenth to be established outside the territorial limits of the United States.

The Board of Directors had also, on the previous day, approved the formation of the Las Cruces-White Sands Proving Ground Subsection of the El Paso Section.



Fred London, Curtiss-Wright Corp., is shown presenting his paper on the principles and applications of radioisotopes to noncontact measurements for continuous processes to the Emporium IRE Section during its recent summer seminar. The seventeenth annual seminar featured the presentation of four papers, a tour of the Sylvania radio tube plant, a picnic, and a golf tournament.

N. W. Florida Section Hears W. G. Shepherd on Cathodes

At their recent meeting in Panama City, Florida, the Northwest Florida IRE Section heard W. G. Shepherd, head of the electrical engineering department of the University of Minnesota, delivered a talk on the influence of the cathode base on properties of oxide cathodes. Among the fifty attendees at the meeting were L. W. McKeehan, Yale University; H. E. Hartig, University of Minnesota, W. M. Whyburn, University of North Carolina; and R. W. Stewart of the Bureau of Ships, Navy Department.

Biophysics Conference Is Set for Columbus, Ohio, Mar. 4–6

A steering committee of over fifty scientists, representing various aspects of biophysical research in this country, has organized a national biophysics conference to be held in Columbus, Ohio, March 4–6, 1957. The conference will encompass studies which employ the approach of physics in biological measurement and theory, at levels of organization from molecules and cells to complex systems and psychophysics.

The program is expected to include twelve invited papers related to different biophysical fields and a large number of contributed papers. Scientists with biophysical interests may write to H. P. Schwan, School of Medicine, University of Pennsylvania, Philadelphia 4, Pa., for further details and information on presenting contributed papers.

National Science Foundation Announces Colloquia Speakers

A. T. Waterman, Director of the National Science Foundation, recently released a schedule of speakers and their topics for the 1956–1957 Colloquia Series of meetings. Speakers and their topics are set for the following dates: December 5—Ralph Cleland on *Genetics in Japan;* January 9—W. L. Duren on *High School and College Mathematics;* February 6—Clarence Zener on *Industrial Laboratory Research;* and March 6—Percy Priest on *A Congressman Views Science.*

All meetings will be held in the Board Room of the National Science Foundation, Washington, D. C., from 10:30 A.M. to noon.

Reliability Symposium at USC

The RETMA Symposium on Applied Reliability will take place on December 19–20, 1956, at Bovard Hall, University of Southern California, Los Angeles, California. Sessions on mechanical reliability, information feedback, and component evaluation usage will be presented. A highlight of the meeting will be an evening panel session on "Failure Feedback—Is It Effective?"

Advance registrations at $3.00 each will be handled by the RETMA Engineering Office, Room 650, 11 W. 42nd St., New York, 36, N. Y.

M.I.T. and IBM Cooperate on Computation Center Project

More than one hundred scientists and engineers from New England colleges took the first step toward using the facilities of the new M.I.T. Computation Center at a special two-week program at the Massachusetts Institute of Technology recently.

They learned the principles of preparing problems to be solved by a modern high-speed computing machine. They were trained to use the IBM type 704 Data Processing Machine, a large electronic computer, which will be installed at M.I.T. in early 1957.

The two-week program was given by nine members of the M.I.T. staff, two representatives of International Business Machines Corporation, and one faculty member from a participating college, Professor John McCarthy of Dartmouth College. In charge of the program, in addition to Professor Morse, were F. M. Verzuh, Assistant Director of the Computation Center and Dean N. Arden, Assistant Professor of Electrical Engineering, both of M.I.T.

During the two-week period of the course its members visited International Business Machines Corporation operations in Poughkeepsie, New York, to see a type 704 computer in operation and to inspect production facilities there.

The M.I.T. Center will be one of the largest and most versatile data processing facilities yet made available primarily for education and basic research. I.B.M. will install the type 704 computer in M.I.T.'s new Karl Taylor Compton Laboratory and contribute toward the cost of maintaining and operating it. Under special arrangements with IBM, the machine will be operated at M.I.T. to solve problems which require high-speed computation facilities.

The program marks the opening of a cooperative venture between the International Business Machines Corporation, M.I.T., and at least 23 other New England colleges to increase the numbers of scientists and engineers qualified to use modern computing machines, and to learn more about their application to research problems in many fields. The center also will be used for instruction and research in management science.

Calendar of Coming Events

Conference on Electrical Techniques in Medicine and Biology, McAlpin Hotel, N. Y., Nov. 7–9

Kansas City IRE Technical Conference, Town House Hotel, Kansas City, Kan., Nov. 8–9

Symposium on Applications of Optical Principles to Microwaves, Washington, D. C., Nov. 14–16

New England Radio Engineering Meeting, Bradford Hotel, Boston, Mass., Nov. 15–16

Office Automation & Human Engineering Conferences of the International Automation Exposition, Trade Show Bldg., N. Y., N. Y., Nov. 26–30

PGVC Eighth National Meeting, Fort Shelby Hotel, Detroit, Mich., Nov. 29–30

Midwest Symposium on Circuit Theory, Michigan State University, E. Lansing, Mich., Dec. 3–4

Second Instrumentation Conference & Exhibit, Biltmore Hotel, Atlanta, Ga., Dec. 5–7

IRE-AIEE-ACM Eastern Joint Computer Conference, Hotel New Yorker, New York City, Dec. 10–12

Winter Meeting of Amer. Nuclear Society, Sheraton-Park Hotel, Washington, D. C., Dec. 10–12

RETMA Symposium on Applied Reliability, Bovard Hall, Univ. of So. Calif., Los Angeles, Calif., Dec. 19–20

Symposium on Communication Theory and Antenna Design, Hillel House, Boston Univ., Boston, Mass., Jan. 9–11

Symposium on Reliability & Quality Control in Elec., Statler Hotel, Wash., D. C., Jan. 14–15, 1957

Symposium on VLF Waves, Boulder Labs., Boulder, Colo., Jan. 23–25

Electronics in Aviation Day, New York City, Jan. 30

PGME Symposium on Recording of Heart Sounds, Univ. of Buffalo Medical School, Buffalo, N. Y., Feb. 14

Conference on Transistor and Solid-State Circuits, Philadelphia, Pa., Feb. 14–15

Western Joint Computer Conference, Statler Hotel, Los Angeles, Calif., Feb. 26–28

National Biophysics Conference, Columbus, Ohio, March 4–6

EJC Second Annual Nuclear Science and Engineering Congress; Fifth Atomic Energy for Industry Conference; International Atomic Exposition, Philadelphia, Pa., March 11–15

IRE National Convention, Waldorf-Astoria and New York Coliseum, New York City, March 18–21

Industrial Electronics Educational Conference, Ill. Inst. of Tech., Chicago, Ill., April 9–10

Ninth Southwestern Regional Conference & Show, Shamrock-Hilton Hotel, Houston, Tex., April 11–13

National Simulation Conference, Shamrock-Hilton Hotel, Houston, Tex., April 11–13

Region Seven Technical Conference & Trade Show, San Diego, Calif., April 24–26

Eleventh Annual Spring Television Conference, Engrg. Society Bldg., Cincinnati, Ohio, April 26–27

Professional Group News

PGIE Sets Education Meeting

The first annual Industrial Electronics Educational Conference, to be jointly sponsored by the IRE Professional Group on Industrial Electronics and the Armour Research Foundation, is scheduled for April 9–10, 1957, at the Illinois Institute of Technology, Chicago, Ill.

Dr. Eugene Mittelmann is general chairman and E. A. Roberts is in charge of the program. James Deterting and Joseph Koval will represent the Armour Research Foundation.

PGRQC Appoints Advisory Board for January Symposium

The Third National Symposium on Reliability and Quality Control in Electronics, jointly sponsored by IRE, RETMA, AIEE, and ASQC, will be held at the Hotel Statler, Washington, D. C., January 14–15, 1957.

The following people have been appointed to the Advisory Board of the symposium by the IRE Professional Group on Reliability and Quality Control: Max Batsel, RCA; W. H. Martin, Office of the Secretary of the Army; L. A. Hyland, Hughes Aircraft Company; R. D. Huntoon, National Bureau of Standards; J. W. McRae, Sandia Corporation; J. K. Sprague, Sprague Electric Company; Capt. H. E. Bernstein (USN, retired); J. E. Keto, Wright Air Development Center; and L. M. Clement, Crosley Division of Avco Manufacturing Company.

The program will consist of sixty-five speakers in twelve technical sessions, a movie, three tours, and a banquet. Symposium transactions will be made available.

San Francisco Chapter of PGEM Develops New Program

The San Francisco Chapter of the Professional Group on Engineering Management has developed a program designed to appeal to their members who at times hold widely differing interests.

Meetings will be held at several electronic firms located in the San Francisco Bay area, where local engineers and managers can talk over specific management and organization techniques and thus broaden their knowledge of management. In this way, it is hoped that members will see several different ways in which successful engineering management is carried out in firms other than their own.

The first meeting, held October 11, took place at the Lenkurt Electric Company, San Carlos, California, one of the world's largest manufacturers of telephone carrier equipment.

Four Chapters Are Annexed

The IRE Executive Committee, at its meeting on August 21, approved the formation of the following Professional Groups: PG on Instrumentation, Washington, D. C. Section; PG on Military Electronics and PG on Telemetry & Remote Control, Philadelphia Section; PG on Vehicular Communications, Baltimore Section.

Fink Receives SMPTE Award

Donald G. Fink, Editor of the IRE and Director of Research for the Philco Corporation, has been awarded the Journal Award by the Society of Motion Picture and Television Engineers for his paper "Color Television vs Color Motion Pictures," published in the June, 1955 *Journal of the SMPTE.* The award was presented on October 9, 1956, during the SMPTE Convention at the Ambassador Hotel, Los Angeles, California.

IRE Names Two Award Winners

R. A. Heising, radio pioneer and consulting engineer, has been named recipient of the Founders Award. The award, which is given only on special occasions to an outstanding leader in the radio industry, was bestowed on Dr. Heising "for his leadership in IRE affairs, for his contributions to the establishment of the permanent IRE Headquarters, and for originating the Professional Group system." Presentation of this award will be made at the annual IRE banquet to be held at the Waldorf-Astoria Hotel, New York, N. Y. on March 20, 1957 during the IRE National Convention.

R. A. Heising

Dr. Heising was associated with the Western Electric Company and Bell Telephone Laboratories from 1914 until his retirement in 1953. He played a major role in the original development of transoceanic and ship-to-shore radio telephone systems for the Bell System and contributed many firsts in this field. He conducted and supervised much research work on ultra-short waves, electronics, and piezoelectric crystal devices that underlie modern radio.

The creator of many important inventions, he is best known for developing several widely-used modulation systems, in particular, the constant-current or Heising modulation system. He has over one hundred U. S. patents, including the patent on the class *C* amplifier, and has published numerous technical papers in engineering journals.

Since 1953 Dr. Heising has been engaged in independent consulting and patent work.

He is a Fellow and Life Member of the IRE and a Fellow of the American Institute of Electrical Engineers and American Physical Society. He received the IRE Morris Liebmann Memorial Prize in 1921 and the Modern Pioneer Award from the National Association of Manufacturers in 1940.

Dr. Heising served as President of the IRE in 1939, Treasurer from 1943 to 1945, and member of its Board of Directors for seventeen years. His chairmanship of numerous IRE committees, especially those on Sections, Professional Groups, and Office Quarters, played an important role in the development of the IRE into the largest engineering society in the world.

J. A. Stratton, Chancellor of the Massachusetts Institute of Technology, has been named to receive the IRE 1957 Medal of Honor, the highest technical award in the radio and electronics field. The award is to be given "for his inspiring leadership and outstanding contributions to the development of radio engineering as a teacher, physicist, engineer, author, and administrator."

J. A. Stratton

The formal presentation of the award will be made at the annual IRE banquet, to be held at the Waldorf-Astoria Hotel, New York City, on March 20, 1957 during the 1957 IRE National Convention.

Dr. Stratton joined MIT in 1925 and served on the staff of the electrical engineering and physics departments for twenty years. In 1945 he was appointed head of the Research Laboratory of Electronics. He became Vice-President and Provost of MIT in 1949, and this year was appointed to the specially created position of Chancellor.

During World War II he served as expert consultant in the Office of the Secretary of War, for which he received the Medal for Merit.

He is the author of a number of important technical papers and books on theoretical physics, especially in the field of electromagnetic theory, and is well known as an authority on college administration and on science and engineering education.

Dr. Stratton is a Fellow of the IRE, American Institute of Physics and the American Academy of Science, and a member of the National Academy of Sciences, Tau Beta Pi and Sigma Xi.

This year Dr. Stratton was appointed a trustee of the Ford Foundation and a member of the nine-member National Science Board of the National Science Foundation.

His many activities include membership on the Naval Research Advisory Committee, Army Scientific Advisory Panel, New York University Self-Study Project, American Institute of Physics's Hutchisson Committee to Evaluate Physics in Engineering Colleges, and National Science Foundation Advisory Committee on Government-University Relations.

NEREM Becomes Fall Meeting

The New England Radio-Electronics Meeting, the annual activity of the Boston and Connecticut Valley Chapters of IRE, is being changed this year from a spring to a fall event. This change has become necessary, with the growth of NEREM, in order to give New England engineers opportunities to learn latest developments since spring national conventions in New York.

This year's meeting will be held on Thursday and Friday, November 15–16, 1956, at the Hotel Bradford, Boston. Besides the technical sessions and exhibits the eleventh NEREM will include discussions on the engineering evaluation of materials. The social part of the program will comprise a cocktail party and a banquet with a speaker.

The committee handling this year's NEREM consists of: R. M. Purinton, Richard Purinton, Inc., General Chairman; F. J. Finnegan, Raytheon Mfg. Co., Vice-Chairman; S. B. Fishbein, American Machine and Foundry Co., Treasurer; Richard Purinton and Francis Finnegan, Exhibits; T. P. Cheatham, Jr., Melpar, Inc. and David Van Meter, Melpar, Inc., Program (Technical Sessions); Paul Wilson, Raytheon Mfg. Co., Program (Value Analysis Sessions); R. P. Axten, Raytheon Mfg. Co., Publicity; Dale Pollack, Consulting Engineer, Arrangements; Leo Rosen, Anderson-Nichols and Co., Registration; Beverly Dudley, Massachusetts Institute of Technology, Past Chairman; B. R. Kamens, Robert A. Waters, Inc., Connecticut Valley Chairman; and R. L. McFarlan, Consulting Engineer, Boston Section Chairman.

Papers Solicited For Solid-State Circuits Symposium

In April, 1957 a symposium entitled "The Role of Solid-State Phenomena in Electrical Circuits" will be held covering the more recent developments in the application to electrical circuits or systems of the more unusual or unexploited physical effects in solids. This symposium is being given because of the ever-increasing importance of solid-state effects in the simplification of circuit functions and electronic apparatus leading to decreased size and increased reliability.

The major area of interest will be in effects which provide for new or improved electronic devices functioning as generators, amplifiers, detectors, measuring instruments, components, etc. There would also be, of course, interest in new circuit responses not heretofore readily obtainable such as the nonreciprocity provided at microwave frequencies by ferrites, or the possibility of a solid-state negative resistance diode.

One aim is to provide an opportunity for electrical engineers to become better informed on the physical effects available for use in electrical circuits and to better understand their operation and basic limitations. The other aim is to provide an opportunity for the physicists and chemists interested in this area to become better acquainted with the relationship of their work to the basic needs in electrical circuit and equipment design. Papers will emphasize the phenomenological description of new or unexploited effects which may be useful in electrical circuits and the consideration of the basic limitations of these effects, as well as the application of these phenomena to electric circuits. Invited papers are planned covering a review of developments in those categories of materials and effects useful in the electrical engineering field as well as an evaluation of the state of the art from an engineering viewpoint.

Abstracts of about 100 words as well as additional material, if available, should be submitted before November 30, 1956 to:
John W. E. Griemsmann, Chairman,
Solid-State Circuits Symposium Committee,
Microwave Research Institute,
55 Johnson Street,
Brooklyn 1, N. Y.

Technical Committee Notes

P. A. Redhead presided at a meeting of the **Electron Tubes** Committee on September 14 at IRE Headquarters. S. E. Webber gave a report on the 1956 Conference on Electron Tube Research which was held at the University of Colorado, Boulder, Colorado, June 26 through June 29, 1956. The committee gave Mr. Webber a vote of thanks for the excellent job which he had done as the 1956 Conference Chairman.

The Proposed Standard on Electron Tubes: Noise Definitions was discussed, amended and unanimously approved on motion by G. D. O'Neill and seconded by G. A. Espersen. This proposed standard will now be forwarded to the Definitions Coordinator for review and comment.

The **Industrial Electronics** Committee met at IRE Headquarters on September 12 with Chairman J. E. Eiselein presiding. It was reported that E. A. Keller and R. D. Chipp have been appointed to the committee. Mr. Eiselein announced that there will be an education conference on industrial electronics at the Illinois Institute of Technology. Program and dates will be reported later.

R. J. Roman, Chairman of Subcommittee 10.1 on Definitions, submitted a list of 34 definitions which had been prepared by his subcommittee.

Eugene Mittelmann, Chairman of Subcommittee 10.3 on Industrial Electronics Instrumentation and Control, gave a report on the present activities of his subcommittee.

Chairman M. W. Baldwin presided at a meeting of the **Standards** Committee on Thursday, September 13 at IRE Headquarters. It was reported that A. E. Martin will be appointed as IRE representative to a subcommittee of ASA Sectional Committee (Y10) on Letter Symbols and Abbreviations.

The Proposed Standards on Electron Tubes: Physical Electronics Definitions was discussed, amended and unanimously approved on motion by P. A. Redhead and seconded by C. H. Page.

The Proposed Standard on Electron Tubes: Camera Tube and Phototube Definitions was discussed and amended. Further consideration will be given to this proposed standard at the next meeting of the Standards Committee.

Books

Automatic Digital Calculators, Second rev. ed. by A. D. Booth and K. H. V. Booth

Published (1956) by Academic Press Inc., 125 E. 23 St., N. Y. 10, N. Y. 234 pages+21 pages of bibliography+5 index pages+ix pages. Illus. 8¾×5¾. $6.00.

This book, now in its second revised edition, surveys a large slice of the digital computer field. It includes sections on history, organization, control, arithmetic, input-output, components, circuits, programming, and applications. The subject matter is technical, but the simple style suggests that the book was aimed at the novice. The serious student is aided by an extensive bibliography.

To provide detailed illustrations, the authors have understandably borrowed from their experience with the series of computers developed at the University of London, England, which go by such unpronounceable names as APE(X)C. To this are added many examples of different techniques used in other computers. Except for tidbits from the smorgasbord of games, machine learning, and language translation, the programming section is restricted to mathematical applications. There is no attempt to discuss business applications.

Since the 1953 edition of the book was written, the computer field has grown tremendously. In the second edition the authors have attempted to keep abreast by inserting new paragraphs on magnetic core and ferroelectric storage, transistors, and automatic programming. Unfortunately they have not succeeded. The book still retains the flavor of the days when there were but a few computers scattered around various universities. It takes no account of the enormous effects of commercial production, both in England and in the United States. The authors' estimates of "current practice" and performance levels have not been revised, and the grafting on of a paragraph on the high-speed NORC computer merely serves to point up the contrast.

One can only conclude that writing an up-to-date textbook on digital computers is a herculean task.

Werner Buchholz
IBM Research Laboratory
Poughkeepsie, N. Y.

Electromagnetic Waves by G. T. DiFrancia

Published (1953) by Interscience Publishers, Inc., 250 Fifth Ave., N. Y. 1, N. Y. 314 pages+6 index pages+xiii pages. 56 figures. 9¼×6¼. $6.00.

This book is particularly suited for use in engineering and physics curricula as an introductory text on electromagnetic theory at about the first-year graduate level. As stated in the preface, the purpose of the book "is to give a clear and readily understandable introduction to those students who will later engage in theoretical research and also to those who will be concerned with the more and more brilliant applications of electromagnetic waves. To accomplish this the author, while attempting to present the classical theory, has always borne in mind the new and elegant standpoints suggested by modern applications."

The use of analytical tools such as diadics, tensors, and Green's functions goes beyond that usually given in a course for radio engineers. However, in view of the widespread application of such techniques in modern research on waveguides and on radiation problems, this appears to be a desirable step in preparing the student to cope with published papers and to engage in research himself. The required mathematical background is given in the first forty pages of the book. This is followed by material on basic electromagnetic theory, fields in moving systems, circuits and transmission lines, wave phenomena, waveguides, and resonators. Of these subjects, the author's specialties—geometric optics and diffraction—receive particularly complete attention from the standpoint of mathematical fundamentals. The theory of waveguides and cavities is given briefly, including orthogonality conditions, but students who intend to specialize in microwave engineering will find it necessary to refer to texts that contain greater practical detail. Similarly, the treatment of radiators and of microwave optics provides the mathematical background for the study of antennas, but does not cover applications.

The translation from the original Italian is clear, and the vector symbols and other notation are similar enough to those in use in this country to cause no difficulty to the reader. Problem lists are not included in this book.

S. B. Cohn
Stanford Research Institute
Menlo Park, Calif.

Studien über einkreisige Schwingungssysteme mit zeitlich veränderlichen Elementen by B. R. Gloor

Published (1955) by Verlag Leeman, Zurich, Switzerland. 230 pages+3 pages of bibliography+viii pages. Illus. 8¼×6. 15.60 S.Fr.

Translation of title: "Single oscillating Circuits with time-variable Elements" Contributions to Theory and Applications of Superregenerative Receivers.

This erudite study is a dissertation for the degree of Dr. Sc. techn. at the Federal Institute of Technology in Zurich, Switzerland.

Its thoroughness is indicated by 54 references (of which seven, beginning with Armstrong's basic paper, were published in the IRE Proceedings). Strange to say, the list omits all the important Japanese contributions to the art.

The author's original work comprises 1) a new approximate solution method for the homogeneous differential equations of an oscillatory circuit with variable elements, 2) experimental support of his mathematical results.

The experiments were carried out at reduced carrier frequencies and quenching period. The test results are illustrated by oscillograms and numerous graphs.

In spite of the prodigious amount of labor expended the results are somewhat meager, hedged in by many reservations and rather obvious, such as:

"To minimize distortions, there should be no coherence between successive quenching periods."

"F.M. signals are received and detected with less distortion in unsaturated, quasi-linear systems than in saturated systems."

Furthermore, the analysis is limited to circuits with separate quenching in order to avoid the increased complexity of self-quenching circuits.

The type looks like the photographic reduction of a typewritten stencil and is hard on the eyes.

Regardless of these defects the book is a worth-while addition to super-regeneration theory and offers valuable diagrams to design engineers.

W. J. Albersheim
Bell Telephone Labs.
Whippany, N. J.

Transistors in Radio and Television by M. S. Kiver

Published (1956) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 302 pages+4 index pages+5 pages of bibliography+vii pages. Illus. 9¼×6¼. $6.50.

According to the author's preface, this book is directed towards "radio and television technicians and . . . other technical workers." In this reviewer's opinion the intended audience will welcome this book as a really useful contribution to the transistor literature.

A good idea of the scope of the book may be gained from its contents. It is divided into ten chapters: Introduction to Modern Electron Theory; Point-Contact and Junction Transistors; Comparison of Point-Contact and Junction Transistors; Transistor Amplifiers; Transistor Oscillators; Transistor Radio Receivers; Transistors in Television Receivers; Additional Transistor Developments; Servicing Transistor Circuits; and Experiments with Transistors. The author is to be particularly commended for his handling of modern electron theory. This is an area which is sometimes a neglected corner even in the knowledge of a transistor application engineer. The author's treatment is largely qualitative, but is clearly written. The chapters on amplifiers and oscillators will be of assistance to those who wish to gain some insight into modern transistor design practices. The chapter on amplifiers contains information on such pertinent subjects as feedback amplification, transformerless audio output systems and direct-coupled systems. The section on oscillators ranges through multivibrators, frequency standards, tunable broadcast band oscillators, and low distortion audio oscillators. The material on transistor radio receivers includes a discussion of several modern broadcast receiver designs and their automatic gain control system arrangements.

In the latter part of the book will be found a collection of material which should be of interest not only to technicians but also to engineers. The survey of current new trends and developments in semiconductors gathers many pieces of information into one place. The section on servicing etched wiring will make good reading for anyone who may be confronted with this task.

A service technician who reads this book will approach his first transistor service job with some measure of confidence. The engineer will find in it a profusely illustrated and easily read review of his basic transistor knowledge.

R. P. Burr
Burr-Brown Research Corporation
Cold Spring Harbor, N. Y.

Linear Transient Analysis, Vol. II by Ernst Weber

Published (1955) by John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 412 pages+28 pages of appendix+10 index pages+xiv pages. Illus. 9¼×6¼. $10.50.

This second volume of Professor Weber's two-volume text on linear transient analysis is—like his other works—remarkable for its thoroughness, precision and clarity. While the first volume dealt, in the main, with the theory of Laplace transformation and its application to the analysis of transients in two-terminal networks, the present volume presents a systematic and definitive account of the transient phenomena in two-terminal-pair networks and transmission lines. To make the volume self-contained, a brief but adequate review of the Fourier and Laplace transform methods is given in the first chapter.

To describe the contents of the volume in greater detail, this reviewer can do no better than quote from the author's preface, since he cannot paraphrase it to any advantage.

"Chapter 2 introduces the concept and matrix description of the two-terminal-pair network for which the term fourpole is preferred as shorter and unambiguous if we use the equivalence of the two terms as definition rather than accept the possible broader meaning of a four-terminal network. Because of the great advantage in the systematic treatment of extended networks composed of fourpoles in cascade, as in nearly all practical communication systems, the matrix notation is used throughout, though more as a matter of convenience in notation than as a real application of matrix analysis. All the necessary relations of simple matrix algebra are given in Appendix 4 together with selected references for further study for those interested in broader applications. The very considerable generalization of solutions for fourpole problems made possible by matrix notation should readily prove its desirability. A brief section on response to frequency-modulated signals concludes his chapter. Wave filters or passive fourpole lines are treated in Chapter 3 with the mathematical discussions needed to cover the extension of the Laplace transform method to difference equations of the particular kind arising here. A brief review of mechanical and thermal analogues is included because of the identical mathematical formulation. The complexity of the general fourpole response and its broad aspect of limited frequency characteristics as either a low-pass or band-pass network invites idealization of the network characteristic as first introduced by K. Kupfmuller. Chapter 4 is devoted to a fairly extensive discussion of this application of the Fourier transform which so far has only been included sketchily in books on network analysis. Chapter 5 gives a systematic exposition of active fourpoles, such as electron tubes and transistors as far as they operate in the small signal region and can thus be considered linear devices. Feedback control circuits and systems have not been included because a number of excellent books are available on this subject. The basic theory of feedback systems is, of course, covered in the section on feedback amplifiers and can readily be transcribed for feedback control systems if one makes the pertinent adjustments for the usual differences in notation and in nomenclature. In Chapter 6, the physical phenomena on transmission systems with distributed parameters are discussed first in a qualitative way in order to stress the background and limits of validity of the conventional engineering concept of transmission-line theory which is particularly applicable to very low frequency systems. The concept of traveling waves is developed with care and solutions for lossless and distortionless lines are derived. The standing wave solution and its significance for the transient behavior of lines concludes the chapter. Chapter 7 is devoted entirely to the ideal cable because of its considerable practical importance. The first, and simple, solution was given by W. Thomson (Lord Kelvin) when he analyzed the electrical characteristics of a transatlantic telegraph cable. New extensions of the inverse Laplace transform are developed as required and pertinent series expansions are discussed. Finally, Chapter 8 presents approximations and the rigorous solution for the general transmission line. Because of the tremendous mathematical complexities encountered, only the simplest types of terminations are considered. The appendices give, as in the first volume, a list of symbols and brief reviews of matrices and functions of a complex variable so as to render the volume nearly self-sufficient. However, for details of the various methods of linear transient analysis and illustrations on simple lumped circuits, it will be necessary to consult the first volume."

Professor Weber's exposition of the topics mentioned in the preface is very lucid and in many places considerably more detailed than that found in other texts on transients in linear systems. As a result, the reader is provided not only with an excellent introduction to the subject, but also with a reference work to which he may return again and again, either for specific results or to gain better general insight. However, this reviewer feels that the scope of the present volume is perhaps a little too narrow, if it is intended to serve as a text for a basic graduate course on transient analysis. One mises, in particular, a more extensive treatment of such subjects as the solution of difference equations by the Laplace transform methods the general solution of partial differential equations, the properties of delta-functions of various orders, and, perhaps, a brief treatment of the problem of approximation in the time and frequency domains. True, one can find treatments of these subjects in various texts and in the periodical literature. Nonetheless, in this reviewer s opinion their sketchy exposition in the present volume detracts somewhat from its suitability as a text for a basic course on the Laplace transform methods.

In any case, the choice of topics in a text is always a matter that admits of much argument. What does not admit of argument is the fact that Professor Weber has once again produced an outstanding text that will be regarded as a definitive work in its field for a long time to come.

L. A. Zadeh
Columbia University
New York, New York

Professional Groups†

Aeronautical & Navigational Electronics—James L. Dennis, General Technical Films, 3005 Shroyer, Dayton, Ohio.

Antennas & Propagation—H. G. Booker, School of Physics and Elec. Engrg., Cornell Univ., Ithaca, N. Y.

Audio—D. W. Martin, The Baldwin Piano Company, 1801 Gilbert Ave., Cincinnati 2, Ohio.

Automatic Control—J. C. Lozier, Bell Tel. Labs., Whippany, N.J.

Broadcast & Television Receivers—L. R. Fink, Research Lab., General Electric Company, Schenectady, N. Y.

Broadcast Transmission Systems—O. W. B. Reed, Jr., Jansky & Bailey, 1735 DeSales St., N.W., Washington, D. C.

Circuit Theory—H. J. Carlin, Microwave Res. Inst., Polytechnic Inst. of Brooklyn, 55 Johnson St., Brooklyn 1, N. Y.

Communications Systems—F. M. Ryan, American Telephone and Telegraph Co., 195 Broadway, New York 7, N. Y.

Component Parts—R. M. Soria, American Phenolic Corp., 1830 S. 54 Ave., Chicago 50, Ill.

Electron Devices—R. R. Law, CBS-Hytron, Danvers, Mass.

Electron Computers—J. D. Noe, Div. of Engineering, Research, Stanford Research Institute, Stanford, Calif.

Engineering Management—Rear Adm. C. F. Horne, Jr., Convair, Pomona, Calif.

Industrial Electronics—C. E. Smith, Consulting Engineer, 4900 Euclid Ave., Cleveland 3, Ohio.

Information Theory—M. J. Di Toro, Polytech. Research & Dev. Corp., 200 Tillary St., Brooklyn, N. Y.

Instrumentation—F. G. Marble, Boonton Radio Corporation, Intervale Road, Boonton, N. J.

Medical Electronics—V. K. Zworykin, RCA Labs., Princeton, N. J.

Microwave Theory and Techniques—H. F. Englemann, Federal Telecommunication Labs., Nutley, N. J.

Military Electronics—C. L. Engleman, 2480 16 St., N.W., Washington 9, D. C.

Nuclear Science—W. E. Shoupp, Westinghouse Elec. Corp., Commercial Atomic Power Activities, P.O. Box 355, Pittsburgh 30, Pa.

Production Techniques—R. R. Batcher, 240-02—42nd Ave., Douglaston, L. I., N. Y.

Reliability and Quality Control—Victor Wouk, Beta Electric Corp., 333 E. 103rd St., New York 29, N. Y.

Telemetry and Remote Control—C. H. Hoeppner, Stavid Engineering, Plainfield, N. J.

Ultrasonics Engineering—J. F. Herrick, Mayo Foundation, Univ. of Minnesota, Rochester, Minn.

Vehicular Communications—Newton Monk, Bell Labs., 463 West St., N. Y., N. Y.

† Names listed are group Chairmen.

Sections*

Akron (4)—C. O. Lambert, 1144 Roslyn Ave., Akron 20, Ohio; M. L. Kult, 1006 Sackett Ave., Cuyahoga Falls, Ohio.

Alamogordo-Holloman (7)—O. W. Fix, Box 915, Holloman AFB, N. M.; T. F. Hall, Box 824, Holloman AFB, N. M.

Alberta (8)—J. W. Porteous, Alberta Univ. Edmonton, Alta., Canada; J. G. Leitch, 13024—123A Ave., Edmonton, Alta., Canada.

Albuquerque-Los Alamos (7)—G. A. Fowler, 3333—49 Loop, Sandia Base, Albuquerque, N. M.; S. H. Dike, Sandia Corp., Dept. 5120, Albuquerque, N. M.

Atlanta (3)—M. D. Prince, 3821 Stoland Dr., Chamblee, Ga.; P. C. Toole, 605 Morningside Dr., Marietta, Ga.

Baltimore (3)—M. I. Jacob, 1505 Tredegar Ave., Catonsville 28, Md.; P. A. Hoffman, 514 Piccadilly Rd., Baltimore 4, Md.

Bay of Quinte (8)—R. L. Smith, Northern Electric Co., Ltd., Box 400, Belleville, Ont., Canada; M. J. Waller, R.R. 1, Foxboro, Ont., Canada.

Beaumont-Port Arthur (6)—W. W. Eckles, Jr., Sun Oil Company, Prod. Laboratory, 1096 Calder Ave., Beaumont, Tex.; E. D. Coburn, Box 1527, Beaumont, Tex.

Binghamton (1)—Arthur Hamburgen, 926 Glendale Drive Endicott, N. Y.; Y. M. Hill, 2621 Smith Dr., Endwell, N. Y.

Boston (1)—R. L. McFarlan, 20 Circuit Rd., Chestnut Hill 67, Mass.; T. F. Jones, Jr., 62 Bay St., Squantum, Mass.

Buenos Aires—A. H. Cassiet, Zavalia 2090 1 "B," Buenos Aires, Argentina; O. C. Fernandez, Transradio Internacional, 379 San Martin, Buenos Aires, Argentina.

Buffalo-Niagara (1)—G. F. Buranich, Route 1, Lewiston, N. Y.; Earl Whyman, 375 Mt. Vernon Rd., Snyder 21, N. Y.

Cedar Rapids (5)—A. H. Wulfsberg, 3235—14 Ave., S.E., Cedar Rapids, Iowa; W. B. Bruene, 2769 Franklin Ave., N.E., Cedar Rapids, Iowa.

Central Florida (3)—K. A. West, 1345 Indian River Dr., Eau Gallie, Fla.; J. M. Kaeser, 1453 Thomas Barbour Dr., Loveridge Heights, Eau Gallie, Fla.

Chicago (5)—R. M. Soria, 1830 S. 54th Ave., Chicago 50, Ill.; G. H. Brittain, 3150 Summit Ave., Highland Park, Ill.

China Lake (7)—H. W. Rosenberg, 217-B Fowler St., N.O.T.S., China Lake, Calif.; C. E. Hendrix, 211-B Byrnes St., China Lake, Calif.

Cincinnati (4)—W. S. Alberts, 6533 Elwynne Dr., Silverton, Cincinnati 36, Ohio; E. M. Jones, 148 Parkway Ave., Cincinnati 16, Ohio.

Cleveland (4)—J. F. Keithley, 22775 Douglas Rd., Shaker Heights 22, Ohio; C. F. Schunemann, Thompson Products, Inc., 2196 Clarkwood Rd., Cleveland, Ohio.

Columbus (4)—W. E. Rife, 6762 Rings Rd., Amlin, Ohio; R. L. Cosgriff, 2200 Homestead Dr., Columbus, Ohio.

Connecticut Valley (1)—B. R. Kamens, 45 Brooklawn Circle, New Haven, Conn.; J. D. Lebel, Benedict Hill Rd., New Canaan, Conn.

Dallas (6)—G. K. Teal, Texas Instruments Inc., 6000 Lemmon Ave., Dallas, Texas; John Albano, 4134 Park Lane, Dallas, Texas.

Dayton (4)—R. W. Ittelson, 724 Golfview Dr., Dayton 6, Ohio; Yale Jacobs, 310 Ryburn Ave., Dayton 5, Ohio.

Denver (6)—R. C. Webb, 2440 S. Dahlia St., Denver 22, Colo.; S. B. Peterson, 1295 S. Jackson, Denver 10, Colo.

Des Moines-Ames (5)—A. D. Parrott, 1515—45 St., Des Moines 11, Iowa; W. L. Hughes, E. E. Department, Iowa State College, Ames, Iowa.

Detroit (4)—M. B. Scherba, 5635 Forman Dr., Birmingham, Mich.; R. H. Reust, 20078 Westbrook, Detroit 19, Mich.

Egypt—H. M. Mahmoud, Faculty of Engineering, Fouad I University, Giza, Cairo, Egypt; E. I. El Kashlan, Main E.S.B. Stations, 4, Sherifein, Cairo, Egypt.

Elmira-Corning (1)—J. P. Hocker, Pilot Plant No. 2, Corning Glass Works, Corning, N. Y.; J. H. Fink, 26 Hudson St., Bath, N. Y.

El Paso (6)—J. C. Nook, 1126 Cimarron St., El Paso, Texas; J. H. Maury, 328 Olivia Circle, El Paso, Texas.

Emporium (4)—D. A. Dander, 22 S. Cherry St., Emporium, Pa.; R. J. Bisso, 99 Meadow Rd., Emporium, Pa.

Evansville-Owensboro (5)—D. D. Mickey, Jr., Engineering Department, General Electric Co., Owensboro, Ky.; C. L. Taylor, 2301 N. York, Owensboro, Ky.

Fort Wayne (5)—T. L. Slater, 1916 Eileen Dr., Waynedale, Ind.; F. P. Smith, 2109 Dellwood Dr., Sunnymede, Fort Wayne, Ind.

Fort Worth (6)—G. C. Sumner, 3900 Spurgeon, Fort Worth, Texas; C. W. Macune, 3132 Forest Park Blvd., Fort Worth, Texas.

Hamilton (8)—A. L. Fromanger, Box 507, Ancaster, Ont., Canada; C. J. Smith, Gilbert Ave., Dancaster Courts, Sub. Serv. 2, Ancaster, Ont., Canada.

Hawaii (7)—R. R. Hill, 46–029 Lilipuna Rd. Kaneohe, Oahu, T. H.; L. R. Dawson, 432 A Kalama St., Lanikai, Hawaii.

Houston (6)—L. W. Erath, 2831 Post Oak Rd., Houston, Texas; R. W. Olson, Box 6027, Houston 6, Texas.

Huntsville (3)—A. L. Bratcher, 308 E. Holmes St., Huntsville, Ala.; W. O. Frost, Box 694, Huntsville, Ala.

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Israel—Franz Ollendorf, Box 910, Hebrew Inst. of Technology, Haifa, Israel; A. A. Wulkan, P.O. B. 1, Kiryat Motzkin, Haifa, Israel.

Ithaca (1)—R. L. Wooley, 110 Cascadilla St., Ithaca, N. Y.; W. H. Murray, General Electric Co., Ithaca, N. Y.

Kansas City (6)—R. W. Butler, Bendix Aviation Corp., Kansas City Division, Kansas City 10, Mo.; Mrs. G. L. Curtis, Radio Industries, Inc., 1307 Central Ave., Kansas City 2, Kan.

Little Rock (6)—D. L. Winn, 10th and Spring Sts., Little Rock, Ark.; F. J. Wilson, 1503 W. 21st St., Little Rock, Ark.

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Long Island (2)—David Dettinger, Wheeler Laboratories, Inc., Great Neck, Long Island, N. Y.; T. C. Hana, 59–25 Little Neck Parkway, Little Neck, Long Island, N. Y.

Los Angeles (7)—V. J. Braun, 2673 N. Raymond Ave., Altadena, Calif.; J. N. Whitaker, 323—15th St., Santa Monica, Calif.

Louisville (5)—O. W. Towner, WHAS Inc., 525 W. Broadway, Louisville 2, Ky.; L. A. Miller, 314 Republic Bldg., Louisville, Ky.

Lubbock (6)—J. B. Joiner, 2621—30th St., Lubbock, Texas; E. W. Jenkins, Jr., Shell Oil Co., Production Department, Box 1509, Midland, Texas.

Miami (3)—E. C. Lockwood, 149 N.W. 105th St., Miami 50, Fla.; E. W. Kimball, 209 Alhambra Circle, Coral Gables 34, Fla.

Milwaukee (5)—W. A. Van Zeeland, 4510 N. 45th St., Milwaukee 16, Wis.; L. C. Geiger, 2734 N. Farwell Ave., Milwaukee 11, Wis.

Montreal (8)—F. H. Margolick, Canadian Marconi Co., 2442 Trenton Ave., Montreal, Quebec, Canada; A. H. Gregory, Northern Elec. Co., Dept. 348, 1261 Shearer St., Montreal, Que., Canada.

Newfoundland (8)—Col. J. A. McDavid, Hdqtrs. DIR-Comm., N.E. Air Command, APO 862, N. Y., N. Y.; J. H. Wilks, 57B Carpasian Rd., St. John, Newfoundland, Canada.

New Orleans (6)—J. A. Cronvich, Dept. of Electrical Engineering, Tulane University, New Orleans 19, La.; N. R. Landry, 620 Carol Dr., New Orleans 21, La.

New York (2)—H. S. Renne, Bell Telephone Laboratories, Inc., Publication Department, 463 West St., New York 14, N. Y.; O. J. Murphy, 410 Central Park W., New York 25, N. Y.

North Carolina-Virginia (3)—M. J. Minor, Route 3, York Rd., Charlotte, N. C.; E. G. Manning, Elec. Engrg. Dep't., N. Carolina State College, Raleigh, N. C.

Northern Alberta (8)—J. E. Sacker, 10235—103rd St., Edmonton, Alberta, Canada; Frank Hollingworth, 9619—85th St., Edmonton, Alberta, Canada.

Northern New Jersey (2)—A. M. Skellett, 10 Midwood Terr., Madison, N. J.; G. D. Hulst, 37 College Ave., Upper Montclair, N. J.

Northwest Florida (3)—F. E. Howard, Jr., 573 E. Gardner Dr., Fort Walton, Fla.; W. W. Gamel, Canoga Corp., P.O. Box 188, Shalimar, Fla.

Oklahoma City (6)—C. M. Easum, 3020 N.W. 14th St., Oklahoma City, Okla.;

(*Sections cont'd*)

* Numerals in parentheses following Section designate region number. First name designates Chairman, second name, Secretary.

Nicholas Battenburg, 2004 N.W. 30th St., Oklahoma City 6, Okla.

Omaha-Lincoln (5)—M. L. McGowan, 5544 Mason St., Omaha 6, Neb.; C. W. Rook, Dept. of Electrical Engineering, University of Nebraska, Lincoln 8, Neb.

Ottawa (8)—C. F. Pattenson, 3 Braemar, Ottawa 2, Ont., Canada; J. P. Gilmore, 1458 Kilborn Ave., Ottawa, Ont., Canada.

Philadelphia (3)—M. S. Corrington, RCA Victor TV Division, Cherry Hill 204-2, Camden 8, N. J.; I. L. Auerbach, 1243—65th Ave., Philadelphia 26, Pa.

Phoenix (7)—Everett Eberhard, 30 E. Colter St., Phoenix, Ariz.; R. V. Baum, 1718 East Rancho Dr., Pheonix, Ariz.

Pittsburgh (4)—Gary Muffly, 715 Hulton Rd., Oakmont, Pa.; H. R. Kaiser, WHC-WWSW, Sherwyn Hotel, Pittsburgh 22, Pa.

Portland (7)—J. M. Roberts, 4325 N.E. 77, Portland 13, Ore.; D. C. Strain, 7325 S.W. 35 Ave., Portland 19, Ore.

Princeton (2)—J. L. Potter, Rutgers Univ., New Brunswick, N. J.; P. K. Weimer, RCA Laboratories, Princeton, N. J.

Regina (8)—William McKay, 2856 Retallack St., Regina, Saskatchewan, Canada; J. A. Funk, 138 Leopold Crescent, Regina, Saskatchewan, Canada.

Rochester (1)—W. F. Bellor, 186 Dorsey Rd., Rochester 16, N. Y.; R. E. Vosteen, 473 Badkus Rd., Webster, N. Y.

Rome-Utica (1)—M. V. Ratynski, 205 W. Cedar St., Rome, N. Y.; Sidney Rosenberg, 907 Valentine Ave., Rome, N. Y.

Sacramento (7)—E. W. Berger, 3421—5th St., Sacramento 20, Calif.; P. K. Onnigian, 4003 Parkside Ct., Sacramento, Calif.

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Salt Lake City (7)—V. E. Clayton, 1525 Browning Ave., Salt Lake City, Utah. A. L. Gunderson, 3906 Parkview Dr., Salt Lake City 17, Utah.

San Antonio (6)—Paul Tarrodaychik, 215 Christine Dr., San Antonio 10, Texas; J. B. Porter, 647 McIlvaine St., San Antonio 1, Texas.

San Diego (7)—R. A. Kirkman, 3681 El Canto Dr., Spring Valley, Calif.; A. H. Drayner, 4520—62 St., San Diego, Calif.

San Francisco (7)—J. S. McCullough, 1781 Willow St., San Jose 25, Calif.; E. G. Goddard, 2522 Webster St., Palo Alto, Calif.

Schenectady (1)—J. S. Hickey, Jr., General Electric Co., Box 1088, Schenectady, N. Y.; C. V. Jakowatz, 10 Cornelius Ave., Schenectady 9, N. Y.

Seattle (7)—K. R. Willson, 1100—17th Ave 206, Seattle 22, Wash.; W. J. Siddons, 6539—39th N.E., Seattle 15, Wash.

Southern Alberta (8)—W. Partin, 448—22nd Ave. N.W., Calgary, Alberta, Canada; R. W. H. Lamb, Radio Station CFCN, 12th Ave. and Sixth St. E., Calgary, Alberta, Canada.

Syracuse (1)—P. W. Howells, Bldg. 3, Room 235, General Electric Co., Electronics Division, Syracuse, N. Y.; G. M. Glasford, Electrical Engineering Department, Syracuse Univ., Syracuse 10, N. Y.

Tokyo—Hidetsugu Yagi, Musashi Kogyo Daigaku, 2334 Tamagawa Todoroki 1, Setagayaku, Tokyo, Japan; Fumio Minozuma, 16 Ohara-Machi, Meguro-Ku, Tokyo, Japan.

Toledo (4)—M. E. Rosencrantz, 4744 Overland Parkway, Apt. 204, Toledo, Ohio; L. B. Chapman, 2459 Parkview Ave., Toledo 6, Ohio.

Toronto (8)—F. J. Heath, 830 Lansdowne Ave., Toronto 4, Ont., Canada; H. F. Shoemaker, Radio College of Canada, 86 Bathurst St., Toronto, Ont., Canada.

Tucson (7)—R. C. Bundy, Department 15, Hughes Aircraft Co., Tucson, Ariz.; Daniel Hochman, 2917 E. Malvern St., Tucson Ariz.

Tulsa (6)—J. D. Eisler, Box 591, Tulsa 2, Okla.; J. M. Deming, 5734 E. 25th St., Tulsa, Okla.

Twin Cities (5)—J. L. Hill, 25—17th Ave. N.E., North St. Paul 9, Minn.; W. E. Stewart, 5234 Upton Ave. S., Minneapolis 10, Minn.

Vancouver (8)—J. S. Gray, 4069 W. 13th Ave., Vancouver, B. C., Canada; L. R. Kersey, Department of Electrical Engineering, Univ. of British Columbia, Vancouver 8, B. C., Canada.

Washington (3)—R. I. Cole, 2208 Valley Circle, Alexandria, Va.; R. M. Page, 5400 Branch Ave., Washington 23, D. C.

Williamsport (4)—F. T. Henry, 1345 Pennsylvania Ave., Williamsport, Pa.; W. H. Bresee, 818 Park Ave., Williamsport, Pa.

Winnipeg (8)—H. T. Wormell, 419 Notre Dame Ave., Winnipeg, Manitoba, Canada; T. J. White, 923 Waterford Ave., Fort Garry, Winnipeg 9, Manitoba, Canada.

Subsections

Berkshire (1)—A. H. Forman, Jr., O.P. 1-203, N.O.D., General Electric Co., 100 Plastics Ave., Pittsfield, Mass.; E. L. Pack, 62 Cole Ave., Pittsfield, Mass.

Buenaventura (7)—W. O. Bradford, 301 East Elm St., Oxnard, Calif.; M. H. Fields, 430 Roderick St., Oxnard, Calif.

Centre County (4)—W. L. Baker, 1184 Omeida St., State College, Pa.; W. J. Leiss, 1173 S. Atherton St., State College, Pa.

Charleston (3)—W. L. Schachte, 152 Grove St., Charleston 22, S. C.; Arthur Jonas, 105 Lancaster St., North Charleston, S. C.

East Bay (7)—H. F. Gray, Jr., 2019 Mira Vista Dr., El Cerrito, Calif.; D. I. Cone, 1257 Martin Ave., Palo Alto, Calif.

Erie (1)—R. S. Page, 1224 Idaho Ave., Erie 10, Pa; R. H. Tuznik, 905 E. 25 St., Erie, Pa.

Fort Huachuca (7)—J. H. Homsy, Box 123, San Jose Branch, Bisbee, Ariz.; R. E. Campbell, Box 553, Benson, Ariz.

Lancaster (3)—W. T. Dyall, 1415 Hillcrest Rd., Lancaster, Pa.; P. W. Kaseman, 405 S. School Lane, Lancaster, Pa.

Memphis (3)—R. N. Clark, Box 227, Memphis State College, Memphis, Tenn. (Chairman)

Mid-Hudson (2)—R. E. Merwin, 13 S. Randolph Ave., Poughkeepsie, N. Y.; P. A. Bunyar, 10 Morris St., Saugerties, N. Y.

Monmouth (2)—W. M. Sharpless, Box 107, Bell Tel. Labs., Red Bank, N. J.; Arthur Karp, Box 107, Bell Tel. Labs., Red Bank, N. J.

Orange Belt (7)—F. D. Craig, 215 San Rafael, Pomona, Calif.; C. R. Lundquist, 6686 De Anza Ave., Riverside, Calif.

Palo Alto (7)—W. W. Harman, Electronics Research Laboratory, Stanford University, Stanford, Calif.; W. G. Abraham, 611 Hansen Way, c/o Varian Associates, Palo Alto, Calif.

Pasadena (7)—R. M. Ashby, 3600 Fairmeade Rd., Pasadena, Calif.; J. L. Stewart, Department of Electrical Engineering, California Institute of Technology, Pasadena, Calif.

Piedmont (3)—C. W. Palmer, 2429 Fairway Dr., Winston-Salem, N. C.; C. E. Bertie, 1828 Elizabeth Ave., Winston-Salem, N.C.

Quebec (8)—R. E. Collin, 41-B Boulevard des Allies, Quebec, P. Q., Canada; R. M. Vaillancourt, 638 Ave. Mon Repos, Ste. Foy, Quebec, Canada.

Richland (7)—W. G. Spear, 1503 Birch, Richland, Wash.; P. C. Althoff, 1800 Thompson, Richland, Wash.

San Fernando (7)—J. C. Van Groos, 14515 Dickens St., Sherman Oaks, Calif. (Chairman).

Tucson (7)—R. C. Eddy, 5211 E. 20 St., Tucson, Ariz.; P. E. Russell, Elect. Eng. Dept., Univ. Ariz., Tucson, Ariz.

USAFIT (5)—L. D. Williams, USAF Institute of Technology, MCLI, Box 3039, Wright-Patterson AFB, Ohio; G. P. Gould, Box 3274, USAFIT, Wright-Patterson AFB, Ohio.

Westchester County (2)—F. S. Preston, Norden Laboratories, 121 Westmoreland Ave., White Plains, N. Y.; R. A. LaPlante Philips Laboratories, Inc., S. Broadway, Irvington, N. Y.

Western North Carolina (3)—Officers to be elected.

Wichita (6)—M. E. Dunlap, 548 S. Lorraine Ave., Wichita 16, Kan.; English Piper, 1838 S. Parkwood Lane, Wichita, Kan.

Symposium on Optics and Microwaves

SPONSORED BY THE PROFESSIONAL GROUP ON ANTENNAS AND PROPAGATION
NOVEMBER 14–16, LISNER AUDITORIUM, GEORGE WASHINGTON UNIVERSITY, WASHINGTON, D. C.

In cooperation with the George Washington University, the Optical Society of America, and the Office of Naval Research, the IRE Professional Group on Antennas and Propagation is presenting the following technical symposium in Washington. In conjunction with this symposium there will be presented the "Instruments of Science" technical exhibit with fifteen equipment demonstration and informational displays of primary interest to scientists working in the optics and microwave fields.

The tentative program for the symposium is as follows:

WEDNESDAY, NOVEMBER 14

9:30 a.m.

SESSION I. THE REGIONS OF THE FREQUENCY SPECTRUM

Microwave Optics: John Brown, Lecturer, University College, London

Infrared Optics: John A. Sanderson, Head, Optical Division, Naval Research Laboratory

Modern Optics: A. Bouwers, N. V. Optische Industries, De Oude Delft, Holland

Electron Optics: L. L. Marton, Head, Electron Optics Division, National Bureau of Standards

SESSION II. OPTOMETRY AND MICROWAVE OPTICS

Microwave Analog of Rods and Cones: J. M. Enoch, School of Optometry, Ohio State University

Lens of the Human Eye: H. A. Knoll, University of California Medical Center

Inhomogeneous Lenses: K. S. Kelleher, Head, Antenna Laboratory, Melpar, Inc.

Optical Experiments at Millimeter Waves: W. Culshaw, Microwave Physics, National Bureau of Standards

THURSDAY, NOVEMBER 15

9:30 a.m.

SESSION III. DIFFRACTION AND ABERRATIONS

Luneberg-Kline Theory: M. Kline, Institute of Mathematical Sciences, New York University

Applications of the Luneberg-Kline Theory: J. B. Keller, Institute of Mathematical Sciences, New York University

The Imaging Properties of Microwave Lenses: G. W. Farnell, Professor, McGill University

Spherical Earth Diffraction: N. A. Logan, Air Force Cambridge Research Center

SESSION IV. OPTICS AND INFORMATION THEORY

Historical Highpoints: O. H. Schade, Radio Corporation of America

Microwave Optics and Information Theory: G. Toraldo di Francia, Vice Director, National Institute of Optics, Florence, Italy

Experimental Aspects of Filtering: M. A. Marechal, Professor, Institute of Optics, Paris, France; Secretary General, French Society of Physics

Microwave-Optical Filter Analysis: A. I. Kohlenberg, Consultant, Melpar, Inc.

FRIDAY, NOVEMBER 16

9:30 a.m.

SESSION V. ATMOSPHERIC AND STELLAR OPTICS

Radio Astronomy: F. T. Haddock, Astronomer, University of Michigan

New Aurora Theory: W. H. Bennett, Staff, Naval Research Laboratory

Radio Atmosphere: M. Katzin, President, Electromagnetic Research Corporation

Reduction of Contrast by Atmosphere: W. F. K. Middleton, Staff, National Research Council, Ottawa, Canada

SESSION VI. OPTICS AND MICROWAVES IN ROCKET FLIGHT

Problems Associated with Atmospheric Flight: F. J. Tischer, Research Laboratories, OML, Redstone Arsenal

Optical Tracking of the Earth Satellite: Karl Henize, Harvard University Observatory

Problems Associated with Rocket Landing: L. M. Hartman, G. E. Special Products Division

PGVC Annual National Conference

FORT SHELBY HOTEL, DETROIT, MICHIGAN
NOVEMBER 29–30, 1956

The theme of this year's annual national conference of the Professional Group on Vehicular Communications will be "Mobile Communications Promote Our Expanding Economy." The conference will be held at the Fort Shelby Hotel, Detroit, Michigan, November 29–30, 1956.

Registration arrangements should be made with H. A. Penhollow, 12249 Woodward Ave., Detroit 3, Michigan. The registration fee is $4.00 for IRE members; $2.00 for IRE student members; and $5.00 for non-members. Banquet, cocktail, and luncheon tickets are also available at $6.50, $1.00, and $5.00, respectively.

Ladies' arrangements include trips to the Plymouth division of the Chrysler Motor Car Company, Windsor, Canada, and the Northland Shopping Center.

Members of the conference committee are as follows: M. B. Scherba, *Section Chairman;* A. B. Buchanan, *General Chairman;* E. C. Denstaedt, *Vice-Chairman;* R. C. Stinson, *Secretary-Treasurer;* W. J. Norris, *Exhibits;* T. P. Rykala, *Program;* N. G. Jackson, *Arrangements;* W. B. Williams, *Publicity;* Zoltan Kato, *Hospitality;* and H. A. Penhollow, *Registration.*

Arrangements for exhibits may be made by contacting W. J. Norris, Michigan Bell Telephone Company, 118 Clifford Street, Detroit 26, Michigan.

THURSDAY, NOVEMBER 29

8:00–9:30 a.m.

Registration

9:30–10:30 a.m.

Opening remarks, Newton Monk, PGVC National Chairman.

Field Application of Transmission Quality Control in Mobile Radio Systems, R. B. Smith, New York Telephone Company.

10:30–11:00 a.m.
Coffee break

11:00 a.m.–Noon

Railroad Radio Communications, L. E. Kearney, Association of American Railroads.

A Selective Calling System to 106A Standards Employing Cold Cathode Thyratrons, W. Ornstein, Canadian Marconi.

Noon–2:00 p.m.
Lunch

2:00–3:00 p.m.

The Important Role of Mobile Communications in the Growing Gas Industry, T. G. Humphries, Alabama Gas Corporation.

Design and Life of Planar UHF Transmitting Tubes, H. D. Doolittle, Machlett Laboratories.

3:00–3:30 p.m.
Coffee break

3:30–4:30 p.m.

Electronics Application in the County of Los Angeles, W. C. Collins, Los Angeles County, California.

Mobile Radio Doesn't Cost—It Pays, R. L. Abel, American Trucking Association.

5:15 p.m.
Cocktail party

7:15 p.m.
Banquet

FRIDAY, NOVEMBER 30

9:30–10:30 a.m.

A Lower Power Industrial Communications Unit, A. W. Freeland, Bendix Radio.

Noise in Communications Antennas, M. W. Scheldorf, Andrew Corporation.

10:30–11:00 a.m.
Coffee break

11:00 a.m.–Noon

Radio Speeds the Flow of Oil, J. E. Keller Dow, Lohnes & Albertson.

Adjacent Channels and the Fourier Curse, J. S. Smith, General Electric Company.

Noon–2:00 p.m.

Luncheon. The speaker will be C. Plummer, Federal Communications Commission.

2:00–2:30 p.m.

Use of Single Sideband for VHF Mobile Service, H. Magnuski, W. M. Firestone, and and R. Richardson, Motorola Inc.

2:30–3:00 p.m.
Coffee break

3:00–4:30 p.m.

Single Sideband AM for Mobile Communications. Panel discussion by C. Plummer, H. Magnuski, J. S. Smith, J. C. Walter, and J. E. Keller. Moderator: A. B. Buchanan.

Second Midwest Symposium on Circuit Theory

KELLOGG CENTER, MICHIGAN STATE UNIVERSITY, EAST LANSING, MICHIGAN
DECEMBER 3–4, 1956

SPONSORED BY THE PROFESSIONAL GROUP ON CIRCUIT THEORY AND THE AIEE

MONDAY, DECEMBER 3

8:00 a.m.
Registration

9:00 a.m.

Opening remarks by I. B. Baccus, Michigan State University.

9:15 a.m.

TOPOLOGY & CIRCUIT THEORY

Chairman: L. A. Pipes, University of California, Los Angeles.

The Vertex, Circuit, Cut-Set and Tie-Set Aspects of Linear Graphs, S. Seshu, Syracuse University, and M. B. Reed, Michigan State University.

Kron's Method of Tearing and Its Applications, F. H. Branin, Jr., Shell Development Company.

Philosophy of the Network vs. the Mathematical Theory of Networks, M. B. Reed, Michigan State University.

Noon
Lunch

1:30 p.m.

SYSTEMS ANALYSIS & SYNTHESIS

Chairman: M. Van Valkenburg, University of Illinois.

Time-Varying Sampled-Data Systems, B. Friedland, Columbia University.

Schwarz Distributions, P. W. Ketchum, University of Illinois.

Sensitivity Considerations in Active Network Synthesis, J. G. Truxal and I. Horowitz, Polytechnic Institute of Brooklyn.

Synthesis of Minimum Phase Transfer Functions, R. H. Pantell, University of Illinois.

6:30 p.m.
Banquet

8:30 p.m.

Introduction of speaker, J. J. Gershon, DeVry Technical Institute.

Engineering Education for the Future, J. D. Ryder, Michigan State University.

TUESDAY, DECEMBER 4

8:30 a.m.

CIRCUIT THEORY & APPLICATIONS

Chairman: L. A. Zadeh, Columbia University.

Systematic Method for Solving Feedback Amplifier Circuits, R. A. Sharpe, Iowa State College.

Topological Graphs of Electromechanical Systems, H. E. Koenig and W. A. Blackwell, Michigan State University.

Equalization of Transistor Low-Pass Amplifiers, H. Hellerman and C. R. Zimmer, Syracuse University.

Patterns of Driving Elements Related to Tubes and Transistors, G. B. Reed, Michigan State University.

Noon
Lunch

1:30 p.m.

THE PLACE OF CIRCUIT THEORY IN EDUCATION

Chairman: W. Boast, Iowa State College.

The Place and Content of Circuit Theory Courses in the Electrical Engineering Curriculum, J. S. Johnson, E. M. Sabbagh and G. R. Cooper, Purdue University.

Panel discussion, moderated by W. R. LePage, Syracuse University.

Second IRE Instrumentation Conference and Exhibit

Sponsored by the Professional Group on Instrumentation and the Atlanta Section
December 5–7, Biltmore Hotel, Atlanta, Georgia

Wednesday, December 5

8:00 a.m.

Registration

10:30 a.m.

Chairman: B. J. Dasher, Georgia Institute of Technology.

Welcome address to be announced.

2:30 p.m.

Industrial Applications of Instrumentation

Chairman: Richard Rimbach, Instruments Publishing Company.

Development of the Transistor Inverter at 20 KC Using Power Transistors, W. A. Martin, Westinghouse Electric Corporation.

Automatic Damping Recorder for Wind Tunnel Application, C. O. Olsson, Oltronix Company.

A Liquid Level Detector Using a Radioactive Source, R. W. Wheeler, Robertshaw-Fulton Controls Company.

Use of the Compensated Hot Thermopile Principle in Industrial Instrumentation, C. E. Hastings and R. T. Doyle, Hastings-Raydist, Inc.

The Principles and Application of Radioisotopes to Non-Contact Measurements for Continuous Processes, O. Bauschinger, Y. M. Chen and F. H. London, Curtiss-Wright Corporation.

Thursday, December 6

9:30 a.m.

Laboratory Instrumentation

Chairman: F. G. Marble, Boonton Radio Corporation.

Setting Up A Standardization Laboratory for Electrical Measuring Instruments, J. O. Reece and P. Greenspan, Motorola, Inc.

Measurement of the Temperature Coefficient of Capacitance and Inductance Over the Range of 5 to 50 Megacycles, Isidore Bady, Signal Corps Engineering Laboratories.

A New High Stability Micro-Microammeter, J. Praglin, Keithley Instruments, Incorporated.

A Barometric Pressure to Current Transducer, F. A. Lapinski, Brown Instrument Division, Minneapolis-Honeywell Regulator Co.

Application of a Gamma Radiation Vapor-Liquid Meter to a Jet Fuel System, Mario Goglia and Henderson Ward, Georgia Institute of Technology.

2:30 p.m

Radiological Instrumentation for Industry and Civil Defense

Chairman: To be announced.

Man-Instrument Relationships in the Design of Nuclear Instrumentation, F. W. Trabold and G. J. Coe, Crosley Division, Avco Manufacturing Corporation.

A Self-Checking Radiation Monitor, W. E. Landauer and K. C. Speh, Airborne Instruments Laboratory, Inc.

Radiological Defense Instrumentation, Jack Greene, Federal Civil Defense Administration.

The HASL Aerial Radiological Monitoring System, Melvin Cassisy, Atomic Energy Commission.

Fall-Out Measurements for Instrument Design Specification, J. H. Tolan, Lockheed, Georgia Division.

Friday, December 7

9:30 a.m.

Aircraft Instrumentation and Acceleration Measurement

Chairman: Ernest Bevans, Massachusetts Institute of Technology, Lincoln Laboratories.

Phase Angle Analogues in Out-of-Sight Control Instrumentation, C. L. Parish, Chance Vought Aircraft, Incorporated.

An Airborne Electric Field Meter, G. C. Rein, Brown Instrument Division, Minneapolis-Honeywell Regulator Company.

Some Instrumentation Problems in Future Geomagnetic Navigational Aids, J. B. Chatterton, Sperry Gyroscope Company, Division of Sperry Rand Corporation.

The Instrumentation of Human Endurance, S. R. Smith, Lockheed, Georgia Division.

Trends in Acceleration Measurement, Anthony Orlacchio and George Hieber, Gulton Industries, Inc.

A Subminiature Self-Recording Accelerometer for High Shock Duty, Herman Erichsen and D. J. Ettelman, Gulton Industries, Inc.

High Frequency, High "G" Calibration, Al Gillen and Earl Feder, Gulton Industries, Inc.

2:30 p.m.

Solid State Devices and Their Application

Chairman: R. R. Law, CBS-Hytron.

Silicon Junction Diodes as Precision Reference Devices, Kurt Enslein, University of Rochester.

The Application of Miniature Saturable Reactors to Electronic Instrument Design, R. S. Melsheimer, Berkeley Division of Beckman Instruments, Inc.

Magnetic Cores for A Transistorized Memory, Frank McNamara, Massachusetts Institute of Technology, Lincoln Laboratories.

Circuit Considerations for A Transistorized Magnetic Core Memory, R. E. McMahon, Massachusetts Institute of Technology, Lincoln Laboratories.

New Solid State Devices for Computer Application, Dick Baker, Massachusetts Institute of Technology, Lincoln Laboratories.

The Cryotron—A Superconductive Computer, Dudley Buck, Massachusetts Institute of Technology, Lincoln Laboratories.

Committee members handling conference details are B. J. Dasher, General Manager; W. B. Wrigley, Exhibits; M. D. Prince, Program; W. B. Miller, Jr., Arrangements; and R. B. Wallace, Jr., Publicity.

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*
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Antennas and Propagation

Vol. AP-4, No. 3, July, 1956

(Proceedings of the Symposium on Electromagnetic Wave Theory, University of Michigan, June 20–25, 1955)

Introduction—K. M. Siegel

Welcoming Address—Samuel Silver

On Field Representations in Terms of Leaky Modes or Eigenmodes—N. Marcuvitz

Solutions to source-excited field problems are frequently represented as superpositions of source-free field solutions. The latter are in general of two types: eigenmodes and non-eigenmodes which are related to the zeros of the total impedance or alternatively the poles of the scattering coefficient of a system. The eigenmodes are everywhere finite and comprise a complete orthogonal set. The noneigenmodes become infinite in the infinitely remote spatial limits of a region and are not in general members of a complete orthogonal set; examples are "radio-active states," "damped resonances," and "leaky waves." Despite their physically singular behavior, the nonmodal solutions can be employed to represent field solutions in certain ranges.

The Interpretation of Numerical Results Obtained by Rigorous Diffraction Theory for Cylinders and Spheres—H. C. Van de Hulst

The classical solutions of the scattering problems for homogeneous spheres and circular cylinders are taken as the starting point for a physical discussion. The limiting cases arising if two or three of the parameters x, m, and $x(m-1)$ are very small or very large are surveyed and interpreted. The remaining paper deals with bodies fairly large compared to the wavelength. It is shown that exact transformations and/or approximate theories may help in the problem of interpolating between rigorous results. It is also shown that the extinction by large bodies is due to a combination of the classical effects of diffraction and geometrical optics with the less familiar edge effects and surface waves. The sign and magnitude of the edge effects for bodies of different refractive index admits of a simple explanation.

Creeping Waves for Objects of Finite Conductivity—W. Franz and P. Beckmann

It is shown that it is not necessary to apply the van der Pol—Bremmer expansion in order to obtain the Watson residue series without remainder integral. There appear two kinds of residual waves. Those of the first kind do not enter the object and correspond to the usual creeping waves for objects of infinite conductivity. They arise from poles in the vicinity of the zeros of $H\nu^{(1)}(ka)$. Residual waves of the second kind correspond to waves transversing the object and arise from poles in the vicinity of the zeros of $J_\nu(nka)$. They are of no importance in the case of strongly absorbing materials. Waves which are expected according to geometrical optics are obtained—as in the case of infinite conductivity—by splitting off an integral. Primary and reflected waves arise from two different saddle points of the same integrand which was thought of till now as only yielding the reflected waves. On the other hand the terms corresponding to the ingoing part of the primary wave give no contribution at all, but must be kept in order to assure the convergence of the integrals when shifting the path of integration.

A Method for the Asymptotic Solution of Diffraction Problems—R. Timman

The equation for the propagation of harmonic waves in a homogeneous medium is considered as the transform of an hyperbolic equation in one more variable. The boundary value problem of diffraction theory can, by this Laplace transform, be related to Cauchy's problem. The transformed problems are solved for 2+1 variables by methods introduced by Evvard and Ward in supersonic airfoil theory. As an example the diffraction problem for a strip is worked out and an asymptotic expression for the transmission cross section is given.

The Modeling of Physical Systems—R. K. Ritt

On the Diffraction Field Near a Plane-Screen Corner—W. Braunbek

It is shown that the diffraction field of an incident plane scalar wave in the vicinity of a plane-screen corner of arbitrary angle can be found approximately by solving Laplace's equation. An approximate solution, which satisfies the boundary conditions exactly, is presented as a simple closed expression by generalizing the known solution of the half-plane problem. A special corner condition, in addition to Meixner's edge condition, is not necessary.

Electromagnetic Radiation Patterns and Sources—Claus Muller

A Refinement of the WKB Method and Its Application to the Electromagnetic Wave Theory—Isao Imai

When dealing with the problem of diffraction of waves, certain special functions appear which are defined by ordinary differential equations of the second order; for example, Bessel functions, Legendre functions, and Mathieu functions appear for the case of a circular cylinder, a sphere, and an elliptic cylinder, respectively. Exact solutions are obtained in the form of infinite series of such functions, which are, in general, poorly convergent when the wavelength is comparable with or smaller than the dimension of the body. In such a case the series can be transformed into contour integrals and then evaluated by the method of steepest descents or by forming the residue series. For this purpose asymptotic expressions for the special functions are needed. A similar situation arises also in the problem of propagation of short radio waves in a horizontally stratified atmosphere.

In this paper a refinement of the WKB method is presented which enables one to obtain very accurate and compact expressions for such functions, which are particularly suited for the evaluation of zeros. The application of the method is illustrated for the case of Bessel functions, parabolic cylinder functions, Coulomb wave functions, etc.

Approximate Method for Scattering Problems—C. E. Schensted

Electromagnetic Research at the Institute of Mathematical Sciences of New York University—Morris Kline

This paper presents current electromagnetic research efforts and research completed during the past few years at New York University. In the research itself emphasis has been placed on basic problems involving appreciable mathematical complexity and mathematical methodology. However, this account describes the results from the standpoint of their contribution to microwave problems, ionospheric and tropospheric propagation, diffraction, the inverse propagation and synthesis problem, antenna and waveguide theory, and other physical problems.

Asymptotic Developments and Scattering Theory in Terms of a Vector Combining the Electric and Magnetic Fields—H. Bremmer

The vector combination

$$\vec{M} = \left(\frac{\mu}{\epsilon}\right)^{1/4}\vec{H} + j\left(\frac{\epsilon}{\mu}\right)^{1/4}\vec{E}$$

which was in principle introduced by Bateman and Silberstein in order to shorten Maxwell's equations for homogeneous media, also proves to be useful for the treatment of inhomogeneous media (ϵ and μ not depending on the time). The vector $M\rightarrow$ is to be considered together with its conjugated quantity $M^X\rightarrow$ obtained by replacing the imaginary unit j by $-j$. In a source-free medium the Maxwell equations reduce to

$$\text{curl}\,\vec{M} + \frac{j}{c}(\epsilon\mu)^{1/2}\partial\vec{M}/\partial\cdot = \frac{1}{4}\,\text{grad}\log\frac{\mu}{\epsilon}\wedge M^x, \quad (1)$$

and to the equation obtained by taking the conjugated complex value.

This relation shows how an interaction between $\vec{M}$ and $\vec{M^X}$ is produced only by the inhomogeneity of the medium. The theory of scattering by special volume elements, as well as that of partial reflections against layers with rapidly changing ϵ and μ, can be based on the single relation (1) while fully accounting for the vectorial character of the field. The introduction of $\vec{M}$ and $\vec{M^X}$ also enables one to put many results of Luneberg-Kline's theory concerning asymptotic developments in a very simple form. As an example we mention the equation:

$$\text{grad } S \wedge \vec{m}_r - i(_r\mu)^{1/2} m_r = c \text{ curl } \vec{m}_{r-1}$$
$$- \frac{c}{\epsilon} \text{grad} \log \frac{\mu}{\epsilon} \wedge \vec{m}_{r-1}{}^x,$$

which fixes all recurrence relations between the consecutive terms of geometric-optical expansions; these expansions are defined by the asymptotic development

$$\vec{M} = e^{ik^s} \sum_{r=0}^{\infty} \left(\frac{i}{kc}\right)^r \vec{m}_r$$

for monochromatic solutions corresponding to some eiconal function S.

The Theoretical and Numerical Determination of the Radar Cross Section of a Prolate Spheroid—K. M. Siegel, F. V. Schultz, B. H. Gere and F. B. Sleator

The exact curve is found for the nose-on radar cross section of a perfectly conducting prolate spheroid whose ratio of major to minor axis is 10:1, for values of π times the major axis divided by the wavelength less than three. The exact acoustical cross section is also found. The mathematical solution is obtained by setting up a series expansion for the scattered wave in terms of two sets of solutions of the vector Helmholtz equation and evaluating the undetermined coefficients in this series by applying the boundary conditions on the surface of the spheroid.

Solution of Problems in Electromagnetic Wave Theory on a High Speed Digital Calculating Machine—E. K. Ritter

This paper contains references to several problems in electromagnetic wave theory which have been solved by numerical methods. In particular, it treats the methods and machines employed by groups at the University of Michigan, Willow Run Research Center, and at the U. S. Naval Proving Ground, Dahlgren, Virginia, in obtaining on a high-speed computing machine a numerical solution for the radar cross section of a prolate spheroid. At the University of Michigan the work was under the direction of K. M. Siegel, while R. A. Niemann was responsible for that done at the Naval Proving Ground.

Edge Currents in Diffraction Theory—P. C. Clemmow

A comparatively simple method for obtaining an asymptotic approximation to the electromagnetic field diffracted by a large aperture in a perfectly conducting, infinitely thin, plane screen is suggested. The method is based on two assumptions: first, that in some regions the scattered field is nearly the same as the field that would be generated by certain currents located on the edge of the aperture; secondly, that at any point on the edge of the aperture these currents are nearly the same as the corresponding currents for a half-plane lying in the plane of the diffracting screen, the straight edge of which is locally coincident with the edge of the aperture. In the crudest approximation the calculation is made on the basis that the half-planes are excited by the incident field alone; higher order approximations arise from a consideration of the interaction between the different parts of the edge of the aperture.

Applications of the method to the cases of a plane wave normally incident on (1) a slit of infinite length with parallel straight edges, and (2) a circular aperture are considered. In the former case several terms of the asymptotic development of the transmission cross section in inverse powers of the slit width are given; in the latter case the aperture and axial fields based on the zero-order approximation which neglects interaction are compared with experimental data published by various authors and with some rigorous calculations of Andrejewski.

On Discontinuous Electromagnetic Waves and the Occurrence of a Surface Wave—B. Van der Pol

Two problems are considered: (1) The field around a dipole free in space. Contrary to the usual treatment, where the moment of the dipole is considered to vary harmonically in time, here the moment is assumed initially to be zero but at the instant $t=0$ to jump to a constant value, which it further maintains. (2) The same dipole is placed vertically on a horizontal plane separating two media of different refractive index. It is shown that the resulting disturbance on the plane is composed of two space waves and one surface wave. First the Hertzian vector at a distance ρ from the dipole is zero. At $t=t_1$ the disturbance arrives there through the less dense medium, and slowly begins to rise till, at the moment $t=t_2$, when the disturbance has had time to reach the same distance through the second (denser) medium it reaches its final static value and further stays constant. During the transitory interval $t_1<t<t_2$ the disturbance is found to be representable, apart from a constant, by a pure surface wave.

The two problems are solved with the help of the modern form of the operational calculus based on the two-sided Laplace transform. The analytical tools of the operational calculus needed are explained in a separate paragraph.

The Excitation of a Perfectly Conducting Half-Plane by a Dipole Field—A. E. Heins

Starting with the solution of two scalar problems in diffraction theory derived by MacDonald in 1915, it is shown that the following problem may be solved. An electric or magnetic dipole is situated in the presence of a semi-infinite, perfectly conducting, thin plane. This problem may be solved by appealing to an appropriate representation of the electromagnetic field. When the formulation is complete, we are left merely with a two-dimensional Poisson equation. The method serves to show why some orientations of the dipole are simpler to handle than others.

A Critique of the Variational Method in Scattering Problems—D. S. Jones

It is shown that the variational method of dealing with the integral equations of scattering problems is equivalent to solving the integral equation directly by Galerkin's method and using the standard formula for the amplitude of the scattered wave. The second method also satisfies the reciprocity theorem. It is therefore suggested that the reciprocity theorem be used as the basis of approximation without the introduction of variational formulas.

The error involved in using an approximate solution is discussed and it is shown that only a special set of approximations can lead to accuracy at low frequencies. Some ways in which bounds for the error may be obtained in special problems are also given.

The Mathematician Grapples With Linear Problems Associated With the Radiation Condition—C. L. Dolph

Diffraction by a Convex Cylinder—J. B. Keller

The leading term in the asymptotic expansion for large $k=2\pi/\lambda$, of the fields reflected and diffracted by any convex cylinder are constructed. The cross section of the cylinder is assumed to be a smooth curve which may be either closed or open and extending to infinity. The method employed is an extension of geometrical optics in two respects. First, diffracted rays are introduced. Secondly, fields are associated with the rays in a simple way. The results are applicable when the wavelength is small compared to the cylinder dimensions.

Near-Field Corrections to Line-of-Sight Propagation—A. D. Wheelon

This study considers the line-of-sight propagation of electromagnetic waves in a turbulent medium. Interest here centers on the received signal's phase stability. The field equation describing propagation through a region characterized by random dielectric fluctuations is first developed. Solutions of this equation which represent the scattered field are derived with ordinary perturbation theory. These solutions are next used to calculate the rms phase error for an arbitrary path in the troposphere. This approach includes both a three-dimensional and near-field description for the multipath, scattered amplitudes, thereby overcoming the limitations of previous treatments. The phase correlation between signals received on two parallel transmission paths is derived last to illustrate the role of overlapping antenna beams.

On the Scattering of Waves by an Infinite Grating—Victor Twersky

Using Green's function methods, we express the field of a grating of cylinders excited by a plane wave as certain sets of plane waves: a transmitted set, a reflected set, and essentially the sum of the two "inside" the grating. The transmitted set is given by $\psi_0+2\Sigma C_v G(\theta_v, \theta_0)\psi_v$, where the ψ's are the usual infinite number of plane wave (propagating and surface) modes; $G(\theta_v, \theta_0)$ is the "multiple scattered amplitude of a cylinder in the grating" for direction of incidence θ_0 and observation θ_v; and the C's are known constants. (For a propagating mode, C_v is proportional to the number of cylinders in the first Fresnel zone corresponding to the direction of mode v.) We show (for cylinders symmetrical to the plane of the grating) that

$$G(\theta, \theta_0) = g(\theta, \theta_0)$$
$$+ \left(\Sigma_v - \int dv\right) C_v [g(\theta, \theta_v) G(\theta_v, \theta_0)$$
$$+ g(\theta, \pi - \theta_v) G(\pi - \theta_v, \theta_0)],$$

where g is the scattering amplitude of an isolated cylinder. This inhomogeneous "sum-integral" equation for G is applied to the "Wood anomalies" of the analogous reflection grating; we derive a simple approximation indicating extrema in the intensity at wavelengths slightly longer than those having a grazing mode. These extrema suggest the use of gratings as microwave filters, polarizers, etc.

Measurement and Analysis of Instantaneous Radio Height-Gain Curves at 8.6 Millimeters over Rough Surfaces—A. W. Straiton and C. W. Tolbert

By the use of an array of ten vertically-spaced antennas and a rotating wave guide switch, a portion of the height-gain pattern for a short radio path was obtained as a function of time for a wave length of 8.6 millimeters.

In the analysis of the data taken across a small lake, the reflection from the water is assumed to be made up of two components. One component is a constant value equal to the median signal received at the antennas over the sampling period and the other component is a variable signal of the proper phase and magnitude to give the measured total signal at each instant.

The angle of arrival, phase and magnitude of the fluctuating signal are obtained for a short sample of data and their characteristics described.

Measurements of the Phase of Signals Received over Transmission Paths with Electrical Lengths Varying as a Result of Atmospheric Turbulence—J. W. Herbstreit and M. C. Thompson

A system for the measurement of the variations in effective lengths of radio propagation paths is described. The observed path-length instabilities are considered to be caused by the same atmospheric turbulence responsible for the existence of VHF and UHF signals far beyond the radio horizon. Preliminary results obtained on 172.8 mc and 1046 mc along a $3\frac{1}{2}$ mile path are reported. It is pointed out that measurements of this type should provide a powerful tool for the study of the size and intensity of the refractivity variations of the atmosphere giving rise to the observed phenomena.

Conditions of Analogy Between the Propagation of Electromagnetic Waves and the Trajectories of Particles of Same Spin with Application to Rectifying Magnetrons—J. Ortusi

The object of this article is the study of the biunivocal correspondence established by the Pauli principle between the internal energy of a particle and the frequency of the associated wave in media which are the seat of strong coupling between the particles and when their spins are in a favored direction.

In Section I, the determinantal forms of antisymmetrical wave functions are investigated, these being valid both for crystals and for electronic plasmas. It is shown that, starting from this determinant, two complementary series of wave functions can be constructed. Depending on the internal energy, two types of complementary particles are thus obtained: (1) free electrons associated with real waves, and (2) holes associated with evanescent waves.

In Section II, a study is made of the mathematical analogy between the Schrödinger equation and the tropospheric propagation equation. It is shown that the potential energy can be assimilated to the refraction modulus and that the group velocity of the propagation around the earth can be assimilated to the group velocity of the complementary particle.

By a very simple correspondence, the real modes of propagation predict the formation of holes while the imaginary modes of propagation predict the formation of free electrons. A special study is made of the analogy between the index barriers of the inversion layers and the potential barriers of the barrier layers. This analogy enables the existence to be predicted of purely electronic barrier layers without the need for any material support.

In Section III, the rectification and photoconduction properties of these electronic barrier layers in magnetrons and in traveling wave magnetron detectors are considered. Their analogies with and differences from the barrier layers of *p-n* junctions are examined. Finally, in the conclusion, the advantages and description of radar detection arrangements devised, on these principles, by the Compagnie Générale de T.S.F. in Paris, are set out.

Scattering at Oblique Incidence From Ionospheric Irregularities—D. K. Bailey

Forward- and Back-Scattering From Certain Rough Surfaces—W. S. Ament

Heuristic relations are derived between the specular reflection coefficient, R, and the radar echoing power of rough surfaces in which induced current elements are constrained to radiate equal powers in the reflected ray's direction and back toward the radar. To the extent that currents in the surface and fields scattered by it are calculable through a self-consistent formulation, a simple Fresnel-zone computation of R shows that σ_0, the radar area per unit area of mean plane, is proportional to $|R^2| \sin^2 \theta$, where θ is the angle incident rays make with the mean plane. It is plausibly assumed that large scatterers on the surface cast shadows with "beamwidth" proportional to radar wavelength λ; here the argument leads to $\sigma_0 \alpha(|R^2| \sin^2\theta)/\lambda$. In two appendices the law $\sigma_0 = 4 \sin^2 \theta$ is derived for a lossless surface obeying Lambert's law, and a known self-consistent "solution" of a rough surface problem is examined by three generally applicable criteria.

Cerenkov and Undulator Radiation—H. Motz

Nonreflecting Absorbers for Microwave Radiation—Hans Severin

The absorption of very short electromagnetic waves by absorbing systems, which avoid reflection of the incident wave is a problem of practical interest. Three different methods are applicable: (1) Complete absorption of the incident energy can be obtained for one wavelength by using resonance systems of relatively small thickness; e.g., a resistance card having a surface resistivity equal to the wave impedance of free space and placed a quarter of the wavelength in front of a metal sheet; a dielectric layer of lossy material on a metal sheet, with the thickness of the layer equal to about a quarter of the wavelength in the material; a two-dimensional periodic structure of concentric resonant circuits arranged within the metal sheet itself. (2) The reflecting object can be covered by a thick layer of absorbing material, so that in a wide wavelength range most energy of the incident wave will be absorbed before reaching the reflecting surface. To avoid reflection, the absorption material can be tapered or arranged in different layers in such a manner that the loss tangent steadily increases towards the base plate. (3) The bandwidth of resonance absorbers can be widened without an increase of its thickness by combination of two specially dimensioned resonant circuits.

Theory of the Corner-Driven Square Loop Antenna—Ronold King

The general problem of determining the distribution of current and the driving point impedances of a square loop or frame antenna is formulated when arbitrary driving voltages are applied at each corner or when up to three of these voltages are replaced by impedances. The loop is unrestricted in size and account is taken of the finite cross-section of the conductors. Four simultaneous integral equations are obtained and then replaced by four independent integral equations using the method of symmetrical components. These equations are solved individually by iteration and first-order formulas are obtained for the distributions of current and the driving-point admittances. By superposition the general solution for the arbitrarily driven and loaded loop is obtained. Interesting special cases include a corner-reflector antenna and the square rhombic (terminated) antenna. An application of the principle of complementarity permits the generalization of the solution to the square slot antenna in a conducting plane when driven from a double-slot transmission line at one corner.

The Radiation Pattern and Induced Current in a Circular Antenna with an Annular Slit—Josef Meixner

A finite plane antenna is considered which has holes on one side that act as sources of radiation and which is on the remaining parts of this side and on the whole other side perfectly conducting. The purpose of this paper is to develop an approximation method for the computation of the radiation pattern which works well if the finite plane and the distance of the holes from its boundary are large compared with the wave length. This is achieved by computing the radiation field of a corresponding infinite plane antenna and subtracting from it the field produced by the current induced in the infinite plane outside the finite plane antenna. Numerical results for the circular antenna with annular slit show that this approximation method is very satisfactory.

Aberrations in Circularly Symmetric Microwave Lenses—M. P. Bachynski and G. Bekefi

The electric field intensity distribution was measured in the image space of solid dielectric microwave lenses at a wavelength of 1.25 cm for various displacements of the source from the principal axis of the system. The experimental results are presented in the form of contours of constant intensity in several receiving planes and also as plots of field intensity versus radial positions of the point of observation. It was found that the deviations of the intensity patterns from the ideal, Airy aberration-free distribution could be interpreted quantitatively in terms of the third order aberrations of optics. The very good agreement obtained with the scalar diffraction theory of aberrations suggests the usefulness of the optical concepts in their application to the centimetric wavelength range.

Spherical Surface-Wave Antennas—R. S. Elliott

Solutions of Maxwell's equations are presented which approximately satisfy the boundary conditions for corrugated and dielectric-clad conducting spheres. These solutions have the physical interpretation of leaky latitudinal surface waves. Values of the complex propagation constant are given as functions of the geometry. For large spheres the leakage is small and the transmission properties approach those of a trapped cylindrical wave on a flat surface.

A corrugated spherical cap, used to support surface waves, has been found to have interesting possibilities as a low-drag omnidirectional antenna. Preliminary experimental results are offered as an illustration of the theory.

Application of Periodic Functions Approximation to Antenna Pattern Synthesis and Circuit Theory—J. C. Simon

Recently, mathematicians gave results on the approximation of periodic functions $f(x)$ by trigonometric sums $P_n(x)$. These results can be useful for antenna radiation and circuit theory problems. Rather than the least mean-square criterion which leads to Gibbs' phenomenon, it has been adopted that the maximum in the period of the error, $|f-P_n|$, is to be minimized. By linear transformation of the Fourier sum, a P_n sum can be obtained to give an error of the order $1/n^p$. The Fourier sum would give Log n/n^p. Limitations on the maximum of P_n derivatives are introduced allowing one to obtain the order of maximum error.

Antenna power diagram synthesis is then looked at with these results. The power radiation v^2 of an array of n isotropic independent sources equally spaced can always be written under the form of a P_n sum. Thus it is possible to give general limitations for the derivatives of v^2 in the broadside case and the endfire case. These limitations depend upon the over-all antenna dimension vs wavelength a/λ and the maximum error. A practical problem of shaped beam antenna is examined. It is shown that, by using the mathematical theory, improvements can be made on the diagram from what is usually obtained.

For circuit theory, physically evident limitations in time T and spectrum F allow one to write the most general function under the form of a P_n sum, and thus to apply the mathematical results to that field. Formal analogy allows comparison of antenna pattern and circuit theories.

A Theoretical Analysis of the Multi-Element End-Fire Array with Particular Reference to the Yagi-Uda Antenna—Yasuto Mushiake

Self and mutual impedances of a multi-element antenna system are discussed, and a method of approximation for these impedances is shown. The impedances derived by this method are applied to a theoretical analysis of the multi-element parasitic end-fire array. Various characteristics of the Yagi-Uda antenna computed by the theory are given in charts, and a procedure for designing the Yagi-Uda antenna is shown. Comparisons between the theory and experimental results are also discussed.

Resolution, Pattern Effects, and Range of Radio Telescopes—J. D. Kraus

Important source parameters and the characteristics of an ideal radio telescope are outlined. The resolution of a telescope antenna is given by Rayleigh's criterion as one-half the beamwidth between first nulls. The effect of source extent on the observed antenna pattern and the inverse problem of determining the source distribution from the measured pattern are considered. The range of a radio telescope is discussed and it is shown that some types of celestial sources could be detected far beyond the celestial horizon if such did not exist. The range of the largest optical telescopes is only half the distance to the celestial horizon, and it is pointed out that observations with large radio telescopes may be vital in determining whether a celestial horizon does in fact exist. The ultimate number of celestial sources that can be resolved with any radio telescope is given by Ko's criterion as numerically equal to the directivity of the telescope antenna.

Radiation from Ring Quasi-Arrays—H. L. Knudsen

The present paper constitutes a summary of investigations of certain antenna systems with rotational symmetry, so-called ring arrays and ring quasi-arrays, which have turned out to be or can be supposed to become of practical importance.

Particular stress has been laid on an investigation of the field radiated from homogenous ring arrays of axial dipoles and homogeneous ring quasi-arrays of tangential and radial dipoles; i.e., systems of respectively axial, tangential, and radial dipoles placed equidistantly along a circle and carrying currents of the same numerical value but with a phase that increases uniformly along the circle.

At first a calculation has been made of the radiated field in the case where the number of elements in the antenna system is infinitely large. After that the influence of the finite number of elements is accounted for by the introduction of correction terms. Subsequently, the radiation resistance and the gain have been calculated in a few simple cases.

The antenna systems described above may display super-gain. On the basis of the theory of super-gain an estimate is made of the smallest permissible radius of these antenna systems.

Further an investigation is made of the field from a directional ring array with a finite number of elements to ascertain in particular the influence on the field of the finite number of elements.

Directivity, Super-Gain and Information—G. Toraldo Di Francia

In this paper some analogies between antenna theory and the theory of optical resolving power are analyzed. The effect of the finite size of a rotating antenna on the informational content of the echo is discussed, without taking into account noise. From this point of view, the most important feature of the aerial is the highest angular frequency which is contained in its radiation pattern. Super-gain is possible because no upper limit exists for this frequency. A simple method is pointed out for synthesizing a radiation pattern containing any prescribed set of finite angular frequencies. A numerical example is worked out.

Exact Treatment of Antenna Current Wave Reflection at the End of a Tube-Shaped Cylindrical Antenna—Erik Hallén

Propagation in Circular Waveguides Filled with Gyromagnetic Material—L. R. Walker and H. Suhl

Using a specific form for the dependence of the permeability tensor components of a ferromagnetic medium on frequency and magnetizing field, the characteristic equation for the propagation constant in circular waveguide is written down. A method for discussing the complete mode spectrum of this equation is outlined. The general behavior of the spectrum is discussed.

The Low-Frequency Problem in the Design of Microwave Gyrators and Associated Elements—C. L. Hogan

The introduction of ferrite microwave circuit elements has allowed considerable simplification in the realization of many system functions. However, to date practical low loss ferrite devices have not been built to operate at frequencies below 3,000 mc. Many problems arise when one attempts to build devices to operate below this frequency. Some of these problems arise from the fact that mechanisms of loss occur in the ferrites at lower frequencies which are negligible at the higher microwave frequencies. In addition, at frequencies below 1,000 mc, one can seldom neglect the existence of internal anisotropy fields in the ferrite materials. The most fundamental limitation to the operation of ferrite devices at very low microwave frequencies, however, is that one is approaching the relaxation frequency for ferromagnetic resonance, and as a result the performance of all ferrite microwave devices must deteriorate at sufficiently low frequencies, regardless of whether one assumes a ferrite whose other properties are ideal. All these problems are discussed and quantitative expressions are obtained for the ultimate low-frequency limitation of ferrite isolators, circulators, and microwave gyrators.

Some Topics in the Microwave Application of Gyrotropic Media—A. A. van Trier

The Faraday effect of plane and guided waves is reviewed in Sections I and II. Section III deals with a cavity technique for measuring Faraday rotations in a circular waveguide with a coaxial ferrite pencil. In Section IV some experimental results are discussed, including the evaluation of the permeability tensor components, the relation between Faraday rotation and pencil radius, and ferromagnetic resonance in circularly polarized waves. The problem of the rectangular gyrotropic waveguide is taken up in Section V. A simple method of successive approximations is described and applied to the case of the square waveguide.

The Seismic Pulse, an Example of Wave Propagation in a Doubly Refracting Medium—C. L. Pekeris

An exact and closed solution is given for the motion produced on the surface of a uniform elastic half-space by the sudden application of a concentrated pressure-pulse at the surface. The time variation of the applied stress is taken as the Heaviside unit function, and its concentration at the origin is such that the integral of the force over the surface is finite. This problem gives an instructive illustration of wave propagation in a doubly refracting medium, since both shear waves and compressional waves are excited, and they travel with different speeds. There is, in addition, the Rayleigh surface wave. For a medium in which the elastic constants λ and μ are equal, the vertical component of displacement w_0 at the surface is given by:

$$w_0 = 0, \qquad \tau < \frac{1}{\sqrt{3}},$$

$$w_0 = -\frac{Z}{\pi\mu r}\left\{\frac{3}{16} - \frac{\sqrt{3}}{32\sqrt{\tau^2 - \frac{1}{4}}} - \frac{\sqrt{5+3\sqrt{3}}}{32\sqrt{\frac{3}{4}+\frac{\sqrt{3}}{4}-\tau^2}} + \frac{\sqrt{3\sqrt{3}-5}}{32\sqrt{\tau^2+\frac{\sqrt{3}}{4}-\frac{3}{4}}}\right\} \qquad \frac{1}{\sqrt{3}} < \tau < 1,$$

$$w_0 = -\frac{Z}{\pi\mu r}\left\{\frac{3}{8} - \frac{\sqrt{5+3\sqrt{3}}}{16\sqrt{\frac{3}{4}+\frac{\sqrt{3}}{4}-\tau^2}}\right\}, \qquad 1 < \tau < \frac{1}{2}\sqrt{3+\sqrt{3}},$$

$$w_0 = -\frac{Z}{\pi\mu r}\,\frac{3}{8}, \qquad \tau > \frac{1}{2}\sqrt{3+\sqrt{3}},$$

where $\tau = (ct/r)$, c-shear wave velocity, and $-Z$ is the surface integral of the applied stress.

The horizontal component of displacement is obtained similarly in terms of elliptic functions. A discussion is given of the various features of the waves.

It is pointed out that in the case of a buried source, an observer on the surface will, under certain circumstances, receive a wave which travels to the surface as an S wave along the ray of total reflection, and from there along the surface as a *diffracted* P wave. An exact expression is given for this diffracted wave.

The question of the suitability of automatic computing machines for the solution of pulse propagation problems is also discussed.

On the Electromagnetic Characterization of Ferromagnetic Media: Permeability Tensors and Spin Wave Equations—G. T. Rado

Various constitutive equations applicable to ferromagnetic and ferrimagnetic media are discussed systematically, the emphasis being on a formulation and analysis of the underlying assumptions. A distinction is made between the "ordinary" (Maxwellian) and certain "average" field vectors. The latter are useful in the presence of domain structure; they include appropriately defined spatial averages, $\langle\vec{b}\rangle$ and $\langle\vec{h}\rangle$, of the time-dependent components of the ordinary $\vec{B}$ and $\vec{H}$, respectively. In cases where $\langle\vec{b}\rangle$ and $\langle\vec{h}\rangle$ are connected by a "point relation," the general form of Polder's permeability tensor is extended to nonsaturated media; the special tensors due to Polder, the writer, and Wangsness, are then reviewed. In cases where $\langle\vec{b}\rangle$ and $\langle\vec{h}\rangle$ are not so connected, the "exchange effect" and the "spin wave equation" are discussed. Following Ament and Rado, three consequences of this equation are treated: the new boundary conditions, and the triple refraction and "equivalent isotropic permeability" in metals.

Plasma Oscillations—D. Gabor

This paper is a report on the investigations by the author and collaborators F. Berz, E. A. Ash, and D. Dracott at Imperial College. F. Berz has theoretically investigated wave propagation in a uniform plasma and found that even in the absence of collisions only damped waves can arise, because the fluctuating velocity distribution contains a term, overlooked by previous authors, which represents a flowing-apart of the electron density. The cutoff due to this effect alone is at about 1.15 of the Langmuir frequency, and the shortest wavelength at about 20 Debye lengths.

Experimental investigations by E. A. Ash and D. Dracott extending over 5 years have at last elucidated the paradox of the existence of Maxwellian electron distributions in the positive column of arcs at low pressures. The interaction is not between electrons and electrons but between these and an oscillating boundary sheath. The sheath was explored by an electron beam probe and oscillations of about 100 mc observed under conditions when the plasma frequency in the arc was about 500 mc. Electrons diving into the boundary sheath spend about one cycle in it, during which time they can gain or lose energies of the order of several volts. Possible applications to radio astronomy are briefly suggested.

Theory of Ferrites in Rectangular Waveguides—K. J. Button and B. Lax

Reciprocal and nonreciprocal propagation of electromagnetic energy in an infinitely long rectangular waveguide partially filled with one or two ferrite slabs is described. Methods for obtaining exact solutions of the transcendental equations usually encountered in these boundary value problems are demonstrated for several structures. Calculations are carried out for a lossless ferrite and the phase constant is plotted as a function of the ferrite slab thickness. The cutoff conditions for the lowest TE mode are evaluated in terms of the ferrite slab thickness. New modes, not associated with the empty waveguide modes, are analyzed as ferrite dielectric modes, their propagation characteristics are discussed and the rf electric and magnetic field patterns are plotted. The rf electric fields are plotted for all reciprocal and nonreciprocal modes and the appropriate field configurations are used to explain the operation of ferrite cutoff isolators, the field-displacement isolator, the field-displacement circulator, and the nonreciprocal phase shifter. Solutions above ferromagnetic resonance are shown and the E-fields are plotted. A brief comparison of the operation of dispersive devices at high and low frequencies is made. The calculations are extended to include absorption loss, and nonreciprocal attenuation is plotted as a function of slab position near resonance.

Panel Discussion on Boundary Value Problems of Diffraction and Scattering Theory—(I)

Panel Discussion on Boundary Value Problems of Diffraction and Scattering Theory (II)

Panel Discussion on Forward and Multiple Scattering

Panel Discussion on Antenna Theory and Microwave Optics

Combined Panel Session on Propagation in Doubly-Refracting Media and Future Directions for Research in Electromagnetic Wave Theory in Modern Physics

Appendix

Index to Authors

Audio

Vol. AU-4, No. 4, July–August, 1956

PGA News

Letters to the Editor

List of Published Standards that May Be Applied to High Fidelity Equipment

The Use of Transistors in Airborne Audio Equipment—V. P. Holec

The need for light weight, low power consumption, reliable audio amplifiers in airborne intercommunication systems led to the development of a new series of amplifiers to meet these requirements. Careful evaluation of the influence of temperature on operating points and circuit stability is an essential part of obtaining a satisfactory and reliable design.

Engineering Consideration of Ceramic Phonograph Pickups—B. B. Bauer

Performance of ceramic pickups is compared to Rochelle salt and magnetic pickups. Whereas voltage-temperature characteristics of barium titanate ceramics and Rochelle salt crystals are relatively constant over a range of temperatures, ceramics exhibit a more stable capacity vs temperature characteristic than does Rochelle salt, and are not subject to damage due to arid and tropical conditions.

The performance of piezoelectric pickups and magnetic pickups is analyzed with respect to the standard recording characteristic. It is concluded that crystal pickups are outstanding when high output is the principal requirement, where quality requirements are moderate, and climatic conditions are benign. Ceramic pickups are the logical choice when quality and economy are both important or where climatic conditions are severe or when magnetic induction is a problem. Current magnetic or dynamic pickups are indicated when the available amplifying equipment, or the present-day public opinion are the principal factors.

Stereo Reverberation—R. Vermeulen

Investigating the reasons why reproduced music gives an impression different from that which a listener receives during a concert, it was found that the distribution of the sound over the room is essential. Although stereophonic reproduction can give a sufficiently accurate imitation of an orchestra, it is necessary to imitate also the wall reflections of the concert hall, in order that the reproduction may be musically satisfactory. This can be done by means of several loudspeakers, distributed over the listening room, to which the signal is fed with different time-lags. The diffused character of the artificial reverberation thus obtained seems to be even more important than the reverberation time. Likewise, when a live orchestra is playing in an acoustically unsatisfactory hall (*e.g.*, a theater), the diffuseness of the sound field and the reverberation time may both be improved by picking up the music by means of a directional microphone and repeating it through loudspeakers with different retardations. The audience does not experience the improvement consciously and ascribe it to the orchestra playing better. The performers, however, are aware of the change in the acoustics as making the hall more playable.

Contributors

Broadcast & TV Receivers

Vol. BTR-2, No. 2, July, 1956

The RETMA Color Television Test Stripe Signal—R. J. Farber

In a television receiver installation, reflections on the antenna feeder line or multipath transmission to the receiving antenna can give rise to selective reinforcement and cancellation throughout any given channel, so that a relatively nonuniform transmission characteristic results. When a monochrome television receiver is involved, this response characteristic is generally only of secondary interest. Since a color television receiver makes more complete use of the available spectrum, it becomes more important in this latter case to have a more or less flat transmission characteristic from the transmitter to the receiver terminals if satisfactory performance is to be had.

By observing the relative transmission of sideband components due to modulation by frequencies in the neighborhood of the color subcarrier, the usefulness of an antenna for a color television receiver can be determined. A color test stripe signal has been devised so that this observation can be made when only monochrome program material is being transmitted. This paper describes the test stripe and its application to color television receivers.

The Synchrotector, A Sampling Detector for Television Sound—Kurt Schlesinger

The paper describes an efficient and economic demodulator for intercarrier television sound. The circuit uses the method of sampling near zero passage of the carrier. This is accomplished in one-half of a double triode. The other half operates as a locked oscillator, whose cathode output is used to drive the sampler cathode. The phase angles between grid and cathode of the sampler are not in quadrature. A centering method to obtain coincidence between optimum fm detection and best AM rejection is described.

Using a conventional double triode 12AU7 this *Synchrotector* locks on signals upward of 10 millivolts, and produces an audio-output of 25 volts with AM rejection ratios between 40 and 50 db.

Technical Standards for Color Television—J. W. Wentworth

This paper consists of a simplified technical derivation of the standards for compatible color television as approved by the Federal Communications Commission for broadcast use. It is shown that compatible *color* television is based upon principles which are logical extensions of the principles used in *monochrome* television, in that means for controlling hue and saturation are added to the conventional means for controlling brightness in the reproduced images. The role of the primary color process in color television is explained, and the electronic multiplexing techniques used to combine the three independent components of a color signal for transmission through a single channel of limited bandwidth are described. The paper is concluded by a summary of all the major processes used in compatible color television from the camera input to the receiver output.

A Printed Circuit IF Amplifier for Color TV—Linus Ruth

This paper deals with the design of a 41 mc IF strip for color TV, in which inductances and wiring are etched on the same board. Advantages of this method, together with problems encountered and their solution, are covered. It is shown that one of the least expensive and most satisfactory methods of tuning printed inductances is with vanes. Graphs of Q variations with distance from coil and tuning range are presented for both vanes and powdered iron slugs. The problems of shielding the large field of the printed coils and elimination of undesirable ground currents are covered. Performance data and response curves of a representative strip are given.

Electron Devices

Vol. ED-3, No. 3, July, 1956

Positive Ion Oscillations in Long Electron Beams—T. G. Mihran

Positive ion oscillations occurring in a long electron beam were investigated experimentally. The predominant direction of oscillation was found to be transverse to the direction of electron flow, and the frequency of oscillation was found to be three times higher than existing theory predicts.

In pulsed beams the onset time of irregularities in current flow due to positive ion formation was found to be inversely proportional to current, and in some cases positive ion effects were observed to take place within four microseconds of the beginning of the pulse.

A New Higher Ambient Transistor—J. J. Bowe

Interest in high speed transistor switching circuits whose operation is unaffected by large changes in ambient temperature led to an investigation of silicon-germanium alloy point-contact transistors because of the larger forbidden energy gap of silicon-germanium alloys. In germanium transistors, as far as temperature stability is concerned, I_{c0} is particularly poor. I_{c0} is the value of collector current, at a given collector voltage, with no emitter current. The I_{c0} of germanium units tested rose rather linearly from 20°C to about 65°C, with a gradient of 25 μa/°C but then entered a region of *run away*. A number of point-contact transistors have been manufactured using 3 per cent silicon-germanium (10 ohm-cm, *n*-type), and the parameters r_{11}, r_{12}, r_{22}, α, f_{c0} and I_{c0} at room temperature, and values of I_{c0} as a function of temperature have been measured.

Results show that 3 per cent silicon-germanium transistors are as good as germanium transistors in all respects and better in temperature stability. The values of I_{c0} for silicon-germanium transistors rose linearly from 18° to about 95°C, with a gradient comparable to that of the germanium units below 65°C.

A Low Voltage One Centimeter Retarding-Field Oscillator—C. J. Carter and W. H. Cornetet, Jr.

The retarding-field oscillator is similar in operation and applications to the reflex klystron but is simpler in structure. In this paper a new low-voltage design is described and some of its experimental characteristics are presented. These include a power output in excess of 40 milliwatts in the wavelength range 0.9–1.1 cm with an anode potential of 400 volts and a beam current of 26 milliamperes. A brief comparison is made between this low voltage retarding-field oscillator and known available reflex klystrons.

Microwave Shot Noise and Amplifiers—F. N. H. Robinson

Several recent papers have used the analogy between an electron beam and a transmission line to discuss beam noise and the minimum noise figure of amplifiers. Despite their basic similarity the treatments given differ so much that it has seemed worthwhile to attempt to review the field and relate the different approaches.

The Gaussistor, A Solid State Electronic Valve—Milton Green

The property of magnetoresistivity can be employed to produce tuned amplifiers and oscillators principally for the sub-audio and possibly for the audio range. To accomplish this, a strip or a coil of a magnetoresistive material, such as bismuth, is placed in the magnetic circuit of a laminated or a ferrite core of an inductor and appropriately wired into an electric circuit containing a dc power supply. The circuitry is simple and the device can be constructed to match a wide range of input and output impedances. The recently developed semiconductor indium antimonide, having an exceptionally high magnetoresistive coefficient, offers hope of obtaining useful power gain at room temperature.

The basic theoretical concepts are presented and experimental results with bismuth and indium antimonide are given.

The Spike in the Transmit-Receiver (TR) Tubes—A. A. Dougal and L. Goldstein

The spike leakage signal from high-Q and band-pass tr tubes was recorded as a function of time by using high-speed oscillographic techniques. Transients as short as 0.5 millimicroseconds (0.5×10^{-9} seconds) were resolved.

The variation of the 1B24 high-Q tr spike was determined as a function of time for several experimental parameters including the gas type and pressure, initial number of electrons in the tr gap, and peak incident power supplied by the transmitter.

Oscillographic recordings show the tr spike leakage power from a commercial gas-filled 1B24 tr tube rises to a peak of 0.3w in a time interval of 6 mμsec. The spike leakage power from a 5863 three-gap band-pass tr tube rises to a peak of 3.6w in 7.5 mμsec.

Threshold of microwave gas discharge breakdown measurements in helium gas are used to determine the electric field intensity in the 1B24 tr gap as a function of the waveguide power. From this, the electron motion during the spike interval is calculated. The results indicate that production of electrons in the gap can occur through ionization of the gas by the electrons' radio frequency energies, and by secondary emission at the gap surfaces, as well as through ionization of the gas by the electrons' energies of random motion.

The Experimental Determination of Equivalent Networks for a Coaxial Line to Helix Junction—W. H. Watson

Equivalent networks were determined for a right angle transition between a coaxial line and a shielded helix. By employing a movable mercury short on the helix it was possible to determine these equivalent circuits through the use of well-known microwave measurement techniques.

Utilizing the possible physical connection which might exist between the junction and its equivalent circuit, an attempt was made to measure quantitatively the effect of varying various parameters in the junction.

For the limited number of cases studied, no simple connection between the elements of the equivalent circuit and the physical parameters of the junction was discovered.

Although the results for the equivalent networks were very sensitive to small experimental errors, by using these networks it was possible to calculate reasonably accurate values of input impedance in the coaxial line for known impedance terminations on the helix.

Forward Transients in Point Contact Diodes—C. G. Dorn

This paper discusses some of the factors which have to be taken into account in the evaluation of point contact diodes for computer work in view of the forward transients which may be present. Oscillograms of forward transients are shown and comparisons of various diodes and operating conditions made. Material is presented to acquaint the engineer with the forward transients attributed to the spreading resistance of point contact diodes, and illustrate why they should be considered by designers of high speed pulse circuits.

A Developmental Intrinsic-Barrier Transistor—R. M. Warner, Jr. and W. C. Hittinger

The intrinsic-barrier design extends transistor frequency range without sacrificing power-handling capacity. A Germanium p-n-i-p transistor has been developed to serve as an oscillator in the neighborhood of 200 mc and to yield approximately 20 mw of useful output at the oscillation frequency. The structure of this developmental unit is described, and some performance and parameter distribution data are given for a group of 53 transistors which were selected on the basis of $\alpha_0>0.7$ and estimated commonbase $f_\alpha>80$ mc. The most efficient unit tested as an oscillator delivered 37 mw at 225 mc with an input power of 160 mw.

Experimental Notes and Techniques

Information Theory

Vol. IT-2, No. 2, June, 1956

Norbert Wiener

What is Information Theory?—Norbert Wiener

Optimum, Linear, Discrete Filtering of Signals Containing a Nonrandom Component—K. R. Johnson

The problem of filtering nonrandom signals from stationary random noise has recently received considerable attention. The filter design procedure developed by Wiener is not applicable in this case since that procedure is predicted on the assumption that the signal to be filtered is stationary and random. Lately, both Booton and the team of Zadeh and Ragazzini have developed optimum filters for the smoothing of nonrandom signals; however, both of these filters are of the continuous type, whereas in many applications in which discontinuous control is used there is need for discrete filters for such signals. This paper presents equations governing the design of a discrete version of the Zadeh-Ragazzini filter. The input signal is assumed to be the sum of a nonrandom polynomial and a stationary random component and is assumed to be obscured by stationary random noise.

An approximate formula for the output noise power of an optimum filter designed to make a zero-lag estimate of either its input function or one of the derivatives thereof is derived for the important special case in which the noise is white and the signal is a nonrandom polynomial. A brief discussion is given of the use of the filter with nonrandom, nonpolynomial signals.

Spatial Filtering in Optics—E. L. O'Neill

Starting with the formulation of H. H. Hopkins for the image forming properties of an optical system in terms of a coherence factor over the object plane, the two extreme cases of complete coherence and incoherence are considered. The incoherent case is treated briefly as a low-pass spatial frequency filter.

In the case of coherent illumination, it is shown that the optical analog of such well-known electrical concepts as equalization, edge-sharpening, and the detection of periodic and isolated signals in the presence of noise can be carried out with relative ease. A detailed theoretical treatment of the problem together with illustrations emphasizes the analogy between optical and electrical filtering.

Effects of Signal Fluctuation on the Detection of Pulse Signals in Noise—Mischa Schwartz

The Neyman-Pearson statistical theory on testing hypotheses has in previous work been applied to the problem of the detection of nonfluctuating constant-amplitude signals embedded in noise. This work is extended in this paper to the case of signal power fluctuating according to a prescribed probability distribution. The effect on system performance of possible correlation between successive signal pulses is taken into account.

The introduction of signal fluctuation leads in general to some loss in system performance as compared to the case of nonfluctuating signals. This loss is most pronounced when there is complete correlation between successive signals, and is quite small when successive signals are independent of one another.

Solution of an Integral Equation Occurring in the Theories of Prediction and Detection—K. S. Miller and L. A. Zadeh

In many of the theories of prediction and detection developed during the past decade, one encounters linear integral equations which can be subsumed under the general form $\int_a^b R(t,\tau)x(\tau)d\tau=f(t)$, $a\leq t\leq b$. This equation includes as special cases the Wiener-Hopf equation and the modified Wiener-Hopf equation $\int_0^T R(|t-\tau|)x(\tau)d\tau = f(t),\ 0 \leq t \leq T$.

The type of kernel considered in this note occurs when the noise can be regarded as the result of operating on white noise with a succession of not necessarily time-invariant linear differential and inverse-differential operators. For this type of noise, which is essentially a generalization of the stationary noise with a rational spectral density function, it is shown that the solution of the integral equation can be expressed in terms of solution of a certain linear differential equation with variable coefficients.

Generalization of the Class of Nonrandom Inputs of the Zadeh-Ragazzini Prediction Model—Marvin Blum

The prediction theory presented in this paper is an extension of the prediction theory of Zadeh and Ragazzini. It differs from their theory in that the nonrandom component of the input signal in the Zadeh-Ragazzini model is restricted to a polynomial of known degree n. In the theory developed here, the nonrandom component of the input signal may be any orbitrary linear function of a subset of known analytic functions where the subset of functions are known *a priori* but the linear relationship need not be. As in the previous solution, the determination of the impulsive admittance of the optimum predictor reduces to the solution of a modified Wiener-Hopf integral equation.

The Correlation Function of a Sine Wave Plus Noise after Extreme Clipping—J. A. McFadden

This paper presents a simple formula for the correlation function of an extremely clipped signal when the input is Gaussian noise plus a sine wave of small amplitude.

A Note on Two Binary Signaling Alphabets—David Slepian

A generalization of Hamming's single error correcting codes is given along with a simple maximum likelihood detection scheme. For small redundancy these alphabets are unexcelled. The Reed-Muller alphabets are described as parity check alphabets and a new detection scheme is presented for them.

Generating a Gaussian Sample—S. Stein and J. E. Storer

The general theoretical difficulties in analyzing the effect of a random input signal on a known system are pointed out. Basically, if certain output statistics are computed directly, each statistic represents a complete, separate problem. An alternative analytical computational procedure is suggested, using a Monte Carlo type technique in which the output is obtained by numerical integration from sequences of values which represent members of the statistical ensemble of the input process. For such applications, or for other possible uses such as in testing, it is necessary to generate statistical sequences, analogous to tables of random numbers.

Techniques are discussed for analytically generating such sequences, to correspond to gaussian probability distributions which are further characterized by arbitrarily specified power spectra or autocorrelation functions. The procedure makes use of the standard tables of random numbers, these numbers being distributed uniformly and without correlation. The exact statistical generation of N values of a sequence is shown to require, in general, the diagonalization (or solution for the eigenvalues and eigenvectors) of an Nth order matrix; two simpler approximate procedures are also described.

A Bibliography of Soviet Literature on Noise, Correlation, and Information Theory—P. E. Green, Jr.

Abstract—On the Information Invariant—Satio Okada

Correspondence

Contributors

Information Theory

Vol. IT-2, No. 3, September, 1956

(1956 Symposium on Information Theory held at Massachusetts Institute of Technology, Cambridge, Massachusetts, September 10–12, 1956)

The Zero Error Capacity of a Noisy Channel—C. E. Shannon

The zero error capacity C_0 of a noisy channel is defined as the least upper bound of rates at which it is possible to transmit information with zero probability of error. Various properties of C_0 are studied; upper and lower bounds and methods of evaluation of C_0 are given. Inequalities are obtained for the C_0 relating to the "sum" and "product" of two given channels. The analogous problem of zero error capacity C_{0F} for a channel with a feedback link is considered. It is shown that while the ordinary capacity of a memoryless channel with feedback is equal to that of the same channel without feedback, the zero error capacity may be greater. A solution is given to the problem of evaluating C_{0F}.

A Linear Circuit Viewpoint on Error-Correcting Codes—D. A. Huffman

A linear binary filter has as its output a binary sequence, each digit of which is the result of a parity check on a selection of preceding output digits and of present and preceding digits of the filter input sequence. The terminal properties of these filters may be described by transfer ratios of polynomials in a delay operator. If two binary filters have transfer ratios which are reciprocally related then the filters are mutually inverse in the sense that, in a cascade connection, the second filter unscrambles the scrambling produced by the first. The coding of a finite sequence of binary information digits for protection against noise may be accomplished by a binary sequence filter, the output of which becomes the sequence to be transmitted. The inverse filter is utilized at the receiver.

Theory of Information Feedback Systems—S. S. L. Chang

A general information feedback system is defined and formulated in a way broad enough to allow coded or uncoded channels with total or partial information feedback. Basic theorems governing change in information rate and reliability are derived with full consideration of the transition probabilities of both direct and feedback channels, including message words as well as the confirmation—denial signal.

A Linear Coding for Transmitting a Set of Correlated Signals—H. P. Kramer and M. V. Mathews

A coding scheme is described for the transmission of n continuous correlated signals over m channels, m being equal to or less than n. Each of the m signals is a linear combination of the n original signals.

On an Application of Semi-Group Methods to Some Problems in Coding—M. P. Schutzenberger

We give an abstract model of some sort of language and try to show how semi-group concepts apply fruitfully to it with the hope that some of them may be of interest to specialists working on natural languages. In a first part, the model and its main properties are discussed at a concrete level on the simplest cases: coding and decoding with length-bounded codes. In a second part a selection of theorems are proved whenever the necessary semi-group-theoretic preliminaries are not exacting.

The Logic Theory Machine—A. Newell and H. A. Simon

In this paper we describe a complex information processing system, which we call the logic theory machine, that is capable of discovering proofs for theorems in symbolic logic. This system relies heavily on heuristic methods similar to those that have been observed in human problem solving activity. The present paper is concerned with specification of the system, and not with its realization in a computer.

Tests on a Cell Assembly Theory of the Action of the Brain, Using a Large Digital Computer—N. Rochester, J. H. Holland, L. H. Haibt and W. L. Duda

Theories by D. O. Hebb and P. M. Milner on how the brain works were tested by simulating neuron nets on the IBM Type 704 Electronic Calculator. The cell assemblies do not yet act just as the theory requires, but changes in the theory and the simulation offer promise for further experimentation.

The Measurement of Third Order Probability Distributions of Television Signals—W. F. Schreiber

A device has been built for the rapid, automatic measurement of the third order probability density of video signals. Examples are presented of second and third order distributions, and of entropies calculated for a variety of scenes.

Gap Analysis and Syntax—V. H. Yngve

A statistical procedure has been tried as a method of investigating the structure of language with the aid of data processing machines. The frequency of gaps of various lengths between occurrences of two specified words is counted. The results are compared with what would be expected if the occurrences of the two words were statistically independent. Deviations from the expected number give clues to the constraints that operate between words in a language.

Three Models for the Description of Language—A. N. Chomsky

We investigate several conceptions of linguistic structure to determine whether or not they can provide simple and "revealing" grammars that generate all of the sentences of English and only these. We find that no finite-state Markov process that produces symbols with transition from state to state can serve as an English grammar. We formalize the notion of "phrase structure" and show that this gives us a method for describing language which is essentially more powerful. We study the properties of a set of grammatical transformations, showing that the grammar of English is materially simplified if phrase-structure description is limited to a kernel of simple sentences from which all other sentences are constructed by repeated transformations, and that this view of linguistic structure gives a certain insight into the use and understanding of language.

Some Studies in the Speed of Visual Perception—G. C. Sziklai

Statistical studies of television signals indicated a high degree of correlation between successive elements, lines and frames. Some tests were devised to measure the perception speed of observers. These tests included certain reading and character recognition tests and finally a test consisting of object recognition in precisely measured periods was devised. Several series of these tests indicated that the visual perception speed of a normal observer is between 30 and 50 bits per second, that this value holds for periods of one-tenth to two seconds, and that the first thing observed is the center of the picture.

Human Memory and the Storage of Information—G. A. Miller

The amount of selective information in a message can be increased either by increasing the variety of the symbols from which it is composed or by increasing the length of the message. The variety of the symbols is far less important than the length of the message in controlling what human subjects are able to remember.

The Human Use of Information III—Decision-Making in Signal Detection and Recognition Situations Involving Multiple Alternatives—J. A. Swets and T. G. Birdsall

A general theory of signal detectability, constructed after the model provided by decision theory, is applied to the performance of the human observer faced with the problem of choosing among multiple signal alternatives on the basis of a fixed, finite observation interval. The results indicate that a highly simplified theory is adequate for prediction of the obtained payoff and response-frequency tables to within a few per cent. They also indicate the fairly large extent to which intelligence may influence a sensory process usually assumed to involve fixed parameters.

On Optimum Nonlinear Extraction and Coding Filters—A. V. Balakrishnan and R. F. Drenick

The problem of determining optimal nonlinear least-square filters is solved for a class of stationary time series. This theory is then used as the basis for developing a band-width reduction scheme using non-linear encoding and decoding filters, for the same class of signals. A simple illustrative example is included.

Final-Value Systems with Gaussian Inputs—R. C. Booton, Jr.

A final-value system controls a response variable $r(t)$ over a time interval $(0, T)$ with the objective of minimizing the difference between a desired value ρ, and the final response value $r(T)$. Physical limitations of the element being controlled result in a maximum-value constraint on the system velocity $r'(t)$. Earlier results suggest that a system consisting of an estimator followed by a "bank-bang" servo is approximately optimum. The estimator uses the input to produce an estimate ρ^* of the desired response and the servo results in a system velocity as large in magnitude as possible and with the same sign as the difference ρ^*-r. The present paper shows that this system is the true optimum when the joint distribution of the input and the desired response is Gaussian and the error criterion is minimization of the average of a nondecreasing function of the magnitude of the error.

An Extension of the Minimum Mean Square Prediction Error Theory for Sampled Data—M. Blum

A method is developed for finding the ordinates of a digital filter which will produce a general linear operator of the signal $S(t)$ such that the mean square error of prediction will be a minimum. The input to the filter is sampled at intervals t. The samples contain stationary noise $N(jt)$, a stationary signal component, $M(jt)$, and a nonrandom signal component. The solution is obtained as a matrix equation which relates the ordinates of the digital filter to the autocorrelation properties of $M(t)$ and $N(t)$ and the nature of the prediction operation.

A New Interpretation of Information Rate—J. L. Kelly

If the input symbols to a communication channel represent the outcomes of a chance event on which bets are available at odds consistent with their probabilities (i.e., "fair" odds), a gambler can use the knowledge given him by the received symbols to cause his money to grow exponentially. The maximum exponential rate of growth of the gambler's capital is equal to the rate of transmission of information over the channel. Thus we find a situation in which the transmission rate has significance even though no coding is contemplated.

A Radar Detection Philosophy—W. McC. Siebert

This paper attempts to present a short, unified discussion of the radar detection, parameter estimations, and multiple-signal resolution problems—mostly from a philosophical rather than a detailed mathematical point of view. The purpose is to make it possible in at least some limited sense to reason back from appropriate measures of desired radar performance to specifications of the necessary values of the related radar parameters.

An Outline of a Purely Phenomenological Theory of Statistical Thermodynamics: Canonical Ensembles—B. Mandelbrot

Since the kinetic foundations of thermodynamics are not sufficient in the absence of further hypotheses of randomness, are they necessary in the presence of such hypotheses? The aim of the paper is to show (partly after Szilard) that a substantial part of the results, usually obtained through kinetic arguments, could be obtained by postulating from the outset a statistical distribution for the properties of a system, and following up with a purely phenomenological argument. It is of interest to the communication engineer to have a unified treatment of the foundations of fluctuation phenomena and of methods of fighting noise.

Production Techniques

PGPT-1, September, 1956

Message from the Editor

Guest Editorials—Electronic Production Techniques—C. L. Munroe

Mechanized Production of Electronics—C. W. Stirling

Automation, the Path to Reliability—M. L. DeGuire

(Symposium on the Automatic Factory in the Production of Electronic Equipment)

Introductory Remarks to the Symposium—R. J. Bibbero, Chairman

Introductory Remarks to Round Table Discussion—John Diebold, Moderator

The Chairman's Notebook and National News

Development and Application of Automatic Assembly Techniques for Miniaturized Electronic Equipment—F. M. Hom

A report is made on some of the development and application of automatic assembly techniques which have been under investigation since 1951 for the U. S. Air Force. Application of automation to miniaturized electronic equipment is stressed. Studies are described covering electronic configurations of various types. Emphasis is placed on heat sinking, component-part density, and on the use of readily available JAN and commercial component parts.

One More Step . . .—Walter Hausz

The speaker emphasizes that other automation systems seem to neglect the job-shop operator. Quoting a machine tool manufacturer, 75% of the metal working operations of the country are in job lots of ten to fifty; and at GE-Syracuse, military and commercial job-lot or semi-job-lot production exceeds by two-to-one the mass production of TV sets and radios. The author proposes that GE introduce automation to small quantity production. Flexibility, minimum set-up time and skill, minimum tool investment, and minimum down time, are named as some of the prime requirements.

A report is made on the development of an Automatic Component Assembly System for the Army Signal Corps. The indicated approach uses various combinations of widths and lengths, from 1 to 12 inches, of printed wiring boards; and at a placement rate of 50 component parts per minute. Standard component parts and conventional constructional methods are only slightly modified in the interest of standardization. Programming of a Weiderman punch press is accomplished by punched cards.

Following his talk, the author showed a motion picture as a progress report.

"Project Tinkertoy"—R. L. Henry

This paper is based on NBS Summary Technical Report 1824. Formerly code named "Project Tinkertoy," Modular Design of Electronics is described as consisting of 4 to 6 steatite wafers stacked into a module. It is stated that flexibility of product design, standardization, and uniformity—the prerequisites for economical processing by automatic machinery—are integral parts of the system. It is shown how an "MDE work sheet" replaces the conventional circuit diagram.

Mechanized Production of Electronics is described as producing electronic parts of many varieties, from raw materials. Processes are outlined for the production of: steatite wafers and tube sockets, titanate capacitors, and asbestos tape resistors. Automatic methods are also discussed for materials handling, component-parts assembly, module assembly, automatic inspection, and final assembly.

It is pointed out that the automatic production of stacked-wafer modules makes possible a rapid conversion from civilian to military products (and back again) on short notice; and concurrently, allows a greatly expanded production capacity by storing "know how" in the form of punched cards and circuit stencil screens for converting raw materials directly into sub-assemblies. It is stated that performance is generally equivalent to that obtainable by conventional assemblies, but with an enhanced high quality level due to automatic production and 100% automatic inspection.

Solderless Wrapped Connection—R. F. Mallina

The solderless wrapped connection is introduced. The terms automatic factory and assembly are defined. Electronic industry growth over the last 30 years is discussed and its expansion during the next 10 years is predicted at 300 per cent. The importance of modular building blocks and modular terminal spacing, in the program of standardization for automation, is emphasized.

The solid-wire solderless wrapped connection is examined; hand wrapping tools, a wrapping gun and wrapping techniques are discussed. The high contact pressure principle is developed and documented by several photographs, drawings, and curves. Data is presented concerning the relaxation of contact-tension with time and temperature. It is shown that tension reaches the 50 per cent point in times varying from three hours (at 170 degrees C) to 40 years (at 57 degrees C). Reliability of the solderless wrapped connection and that of the solder connection are compared under vibration. Other advantages of the solderless wrapped connection are tabulated. Some applications to the wiring of telephone trunk circuit panels and general rack-and-panel wiring installations are given.

Mr. Mallina concluded his talk by showing a motion picture of the wrapping gun in action.

Development of Systems of Mechanized Assembly—W. H. Hannahs

It is pointed out that the radio-making machine of John Sargrove and the serigraphic methods of Brunetti and Khouri have repeatedly stimulated product engineering since 1946. The Sargrove method is briefly described. The early work of Snyder on printed wiring is discussed, and a unitized telemetering channel assembly—of 1947 vintage—is shown. The author poses the complex question of whether to stock parts, assemblies, or complete units—a prerequisite aspect to mechanized design which yet remains unanswered in both military and civilian electronics. As a general approach, however, a unitized type of construction is described as having a considerable long-range validity in terms of servicing, transferring of heat, and ease of sub-assembling.

The early work by Danko on dip soldering is mentioned, and hot-peel strength data for foil-clad laminates is presented. Early work on the printed wiring socket is outlined, and the flexible printed circuit shown at the 1951 IRE National Convention is reviewed.

Attention is given to a study using limited printed wiring, and standard art and practices, as sponsored by the Write Air Development Center. Using a transformer coupled intercomm as a vehicle, a common module was chosen—fixed in cross section but graduated stepwise in length, so the component parts could be stacked. A wrap-around flexible circuit was designed which was adaptable to automation using a cam-operated multihead soldering machine. A mechanical lock of the component parts before soldering, and the importance of accessibility are stressed in the development.

The paper is concluded with a series of curves on rejection percentages and on variables controlling printed circuits. A plea is made for the establishment of standard practices—a role already begun by various RETMA committees.

The Economics of Automation—Some Important Considerations—A. A. Lawson

It is suggested that both the machinery builder and the user need to know the savings reflected in the end product. The "break-even point" is described as the time when the variable costs due to automation are less than the variable manual costs, by the amount of the automation capital investment. A break-even load curve relates savings during a regular 8-hour day to a longer or shorter working day. By observing the distinctions between fixed costs and variable costs and by using conservative estimates, the capabilities of the Mini-Mech system are presented.

Three years is suggested as the useful life for electronic automation equipment, in contrast with the usual 15-year amortization period for other machinery, and with some confidence that the Internal Revenue Service will see eye-to-eye with the user. It is estimated that the Mini-Mech system might require an initial total investment of $25,000; machinery modification costs are estimated at 50% of this figure over the three-year period, as component parts and techniques are improved. At an insertion rate of 24 component parts per minute, the quoted variable saving is approximately $275 per day, and with a break-even point at 123 days. It is stated that, up to this point, hand methods are cheaper as a result of lower fixed costs; but beyond this point, machine methods are cheaper since accrued variable-cost savings will have liquidated the higher fixed costs of the automatic equipment.

Administrative Committee of the IRE Professional Group on Production Techniques

Standing and Working Committees of the PGPT Administrative Committee (1953–1956)

Constitution of IRE Professional Group on Production Techniques

Calendar of Coming Events

News of the Chapters

Ire Professional Group on Production Techniques—Membership Directory

Reliability & Quality Control

PGRQC-8, September, 1956

Reliability Criterion for Constrained Systems—M. J. DiToro

In a system designed to within a constraint such as a given over-all weight wherein failure of any one or more of the system's components causes system failure a design criterion for the weight vs. failure probability parameters of each component is given to achieve maximum system reliability.

Some Reliability Aspects of Systems Design—Fred Moskowitz and J. B. McLean

Elementary principles of probability theory are used and a systematic development is presented which leads to formulas, charts and guide rules for engineers involved in the design of systems and equipments. Examples are given which illustrate the use of the formulas and the principles derived.

This study attempts to show that when the problem of obtaining reliable equipment which consists of unreliable parts is present the solution is redundancy. Complexity by itself need not necessarily lead to unreliability if complexity is used correctly. Two very simple redundancy schemes are described and analyzed. It is shown that it is possible to obtain a desired reliability at relatively reasonable cost in terms of increased size and weight.

Designing for Reliability—S. A. Meltzer

A mathematical model is presented which will enable the designer of electronic equipment to compute its survival probability. Thus, he can make the engineering decision on whether or not the equipment meets its reliability specification. Although only electronic circuits are discussed here, the model is equally applicable to mechanical and electrical systems.

In the model the performance parameters (gain, power output, bandwidth, etc.) are expressed as a function of the individual components of the circuit. The statistical distributions of the component characteristics (dependent on both operating time and environment) are then used in conjunction with the circuit equations to yield a simplified expression for the circuit survival probability. This simplification involves an approximation method using the multivariate Gram-Charlier expansion. These methods are applicable even if the component characteristics are correlated and are described by any probability density function. The fact that individual performance parameters may be correlated is taken into account automatically.

A Developmental Approach to Reliability in Missile System Equipment—J. H. Yueh

Some concepts of reliability based on failure rates of components are presented. These concepts are further interpreted from a statistical point of view for better understanding of the basic reasons for component failures—hence unit, subsystem and system failures. Based on these concepts a program of testing, data collection and analysis is outlined as an approach to estimating and controlling these failure rates to a sufficiently low level so that system reliability is consistent with design objectives.

The Background of Reliability—T. A. Smith

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, 1955 through January, 1956 is published by the PROC. IRE, June, 1956, Part II. It is also published by *Wireless Engineer* and included in the March, 1956 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534.112 2929

Investigation of the Dependence of the Number of Antinodes on a Linear Elastic Body on the Tension of the Individual Mass Elements, as demonstrated by Transverse Waves on Strings—H. Fark. (*Frequenz*, vol. 10, pp. 89–91; March, 1956.)

534.121.1 2930

Vibrations of a Rectangular Plate with Distributed Added Mass—H. Cohen and G. Handelman. (*J. Franklin Inst.*, vol. 261, pp. 319–329; March, 1956.)

534.2-14 2931

An Estimate of the Effect of Turbulence in the Ocean on the Propagation of Sound—J. A. Knauss. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 443–446; May, 1956.)

534.2-8-14 2932

The Absorption of Ultrasonic Waves in Water and its Dependence on the Temperature and Air Content of the Water—S. K. Mukhopadhyay. (*Acustica*, vol. 6, pp. 25–34; 1956. In German.)

534.21-16:549.514.5 2933

Propagation of Longitudinal Waves and Shear Waves in Cylindrical Rods at High Frequencies—H. J. McSkimin. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 484–494; May, 1956.) General theory is presented, and experimental results are reported for propagation at 10–25 mc in fused-silica rods of radius 1.13 cm.

534.22-14:546.212 2934

Temperature Coefficient of the Speed of Sound in Water near the Turning Point—M. Greenspan, C. E. Tschiegg, and F. Breckenridge. (*J. Acoust. Soc. Amer.*, vol. 28, p. 500; May, 1956.) Results of measurements at a frequency of 15.3 kc and temperatures between 70° and 77.5°C. are presented graphically. The temperature coefficient α_c is given by the formula $-25.9\ (T-73.95)\times10^{-6}$/°C. and the calculated velocity is 1555.5 mps at the turning point 73.95°C.

534.22-14:546.212 2935

Effect of Dissolved Air on the Speed of Sound in Water—M. Greenspan and C. E. Tschiegg. (*J. Acoust. Soc. Amer.*, vol. 28, p. 501; May, 1956.) The effect of dissolved air does not exceed 1 part in 10^5 at temperatures of 31.8° and 0°C.

534.231 2936

The Radiation Force on a Spherical Obstacle in a Cylindrical Sound Field—T. F. W. Embleton. (*Canad. J. Phys.*, vol. 34, pp. 276–287; March, 1956.) A general expression is obtained for the radiation force in terms of the complex amplitudes of spherical harmonics required to synthesize the incident field. The results are qualitatively the same as for a spherical field (1636 of 1954), but the point at which the force changes from attraction to repulsion, for a given obstacle size and sound frequency, is nearer the source.

534.232 2937

Directional Circular Arrays of Point Sources—W. Welkowitz. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 362–366; May, 1956.) The Fourier-series solution for the radiation field of a circular current sheet presented by LePage, *et al.* (31 of 1951) is applied to the synthesis of a sound field expressed in the form of a Tchebycheff polynomial. This leads to an exact solution in closed form for the amplitude and phase of excitation of the transducer elements when the main lobe width of the radiation pattern, the sidelobe suppression and the array circle diameter are specified. Some numerical results are given.

534.24+[538.566:535.42 2938

Fourier-Transform Method for the Treatment of the Problem of the Reflection of Radiation from Irregular Surfaces—W. C. Meecham. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 370–377; May, 1956.)

534.52 2939

Scattering of Sound by Sound—U. Ingard and D. C. Pridmore-Brown. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 367–369; May, 1956.) "Calculations and measurements are reported of the summation and difference frequency components which are scattered from the interaction region of two sound beams in air intersecting each other at right angles."

534.64+621.317.73 2940

An Impedance Measuring Set for Electrical, Acoustical and Mechanical Impedances—Ayers, Aspinall, and Morton. (See 3149.)

534.64 2941

Some Notes on the Measurement of Acoustic Impedance—O. K. Mawardi. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 351–356; May, 1956.) The theory of a plane-wave method of measuring the acoustic impedance of a specimen in a tube is developed and the effect of surface irregularities is investigated.

534.75 2942

Intelligibility of Diphasic Speech—G. E. Peterson, E. Sivertsen, and D. L. Subrahmanyam. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 404–411; May, 1956.) The effect on intelligibility of the switching rate at which successive portions of the speech signal are reversed in phase was investigated; the intelligibility was high at switching frequencies up to 100 cps.

534.78:621.39 2943

The Vobanc—a Two-to-One Speech Bandwidth Reduction System—B. P. Bogert. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 399–404; May, 1956.) The Vobanc (voice band compression) system is described and the characteristics of experimental equipment are given. See also 2605 of October (Kock).

534.79 2944

The Significance of the "Frequency Group" for the Loudness of Sounds—H. Bauch. (*Acustica*, vol. 6, pp. 40–45; 1956. In German.) Report of an experimental investigation of the effect on the subjective loudness of a complex tone of the frequency separation of the component tones, for different absolute frequencies, intensity levels, and relative phases. The "frequency group" is the term applied to a critical bandwidth; for lower values of bandwidth the ear assesses the loudness as for a single tone. Intensity fluctuations can still be perceived at frequencies <3 cps.

534.833.4:538.569.2/.3].029.6 2945

Absorption Devices for Centimetre Electromagnetic Waves and their Acoustic Analogues—Meyer and Severin. (See 3037.)

534.844/.845 2946

Determination of the Form of Reverberation Chambers for Measurements—G. Venzke. (*Acustica*, vol. 6, pp. 2–11; 1956. In German.) The influence of the shape and size of the reverberation chamber on the measurement results obtained has been investigated experimentally; procedure recommended in German Standard DIN 52212 was used. For comparison, calculations were made of the curvature to be expected in the reverberation curves for rectangular rooms of different sizes. In addition, independent measurements were made of the absorption coefficient of a particular material in two separate sets of rooms, each set ranging in volume from 83 to 258 m³. The results indicate that the uncertainty of measurements increases with decrease of room size, especially for highly absorbent materials. Rooms with nonparallel plane walls are not necessarily better than rectangular rooms.

534.846.6 2947
Accuracy of Matching for Bounding Surfaces of Acoustic Models—A. F. B. Nickson and R. W. Muncey. (*Acustica*, vol. 6, pp. 35–39; 1956.) A theoretical study is reported of the extent to which precise matching of the specific acoustic impedance of the surfaces of the model with those of the original space is possible or necessary.

534.861.1:621.376.223 2948
Sound Transformations in Broadcasting Studio Technique, Particularly by Application of Frequency Conversion—L. Heck and F. Bürck. (*Elektronische Rundschau*, vol. 10, pp. 1–7; January, 1956.) The production of special sound effects particularly by means of a ring modulator, is discussed.

621.395.616 2949
Air-Stiffness Controlled Condenser Microphone—T. J. Schultz. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 337–342; May, 1956.) The construction and characteristics of small microphones using rubber hydrochloride (pliofilm), vinylidene chloride (saran), or polyethylene terephthalate (mylar) membranes, are described.

621.395.623.7.012 2950
A Method for the Measurement of the Directivity Factor[of loudspeakers]—G. Sacerdote and C. B. Sacerdote. (*Acustica*, vol.6, pp. 45–48; 1956.)

2951
621.395.625.3+621.395.92+621.396.62]
:621.314.7
Transistor Circuity in Japan—(See 3204.)

621.395.625.3:534.86 2952
An Acoustic Time-Regulator for Sound Recordings—A. M. Springer. (*Elektrotech. Z.*, *Edn B*, vol. 8, pp. 93–96; March 21, 1956.) The reproduction of a magnetic-tape recording may be expanded or compressed in time, without changing the pitch, by changing the speed of the tape and simultaneously moving the pickup head so as to keep their relative speed constant. This is achieved by using a quadruple rotating pickup head in conjunction with the mechanical coupling to the tape-drive motor described. For a short description, in English, see *Electronics*, vol. 29, pp. 184, 188; June, 1956.

ANTENNAS AND TRANSMISSION LINES

621.315.212.011.3 2953
The Mean Geometrical Distances of a Circle—H. Schering. (*Elektrotech. Z., Edn A*, vol. 77, pp. 12–13; January 1, 1956.) Simple formulas are derived for the mean distances of a circle from itself and from a surface enclosed by it. The formulas involve power series whose convergence is such that they need not be taken beyond the quadratic term. Further formulas are derived for the inductance of coaxial lines with inner conductors of various cross sections.

621.372.2 2954
Theory of Helical Lines—S. Kh. Kogan. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 107, pp. 541–544; April 1, 1956. In Russian.) The dispersion characteristics of helical lines are derived, taking into account both of the orthogonal components of the current in the case of a thin helical strip or of the field in a helical slit cut in a thin metallic cylinder. Calculated characteristics for three different conductor-width/spacing ratios are presented graphically.

621.372.2 2955
The Propagation of Electromagnetic Waves along a Helical Strip in a Circular Waveguide—E. V. Anisimov and N. M. Sovetov. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1965–1971; October, 1955.) Theory is developed for the case of an ideally conducting strip. The em wave at certain frequencies is the sum of a number of components; these are given by (16). The results can be applied to more complex helical systems.

621.372.2:621.372.8 2956
Electromagnetic Surface Waves in Rectangular Channels—M. A. Miller. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1972–1982; October, 1955.) The conditions necessary for a surface wave to be guided by a rectangular channel are discussed; no surface waves can be present in a channel with an ideally conducting bottom. Analysis is presented for two systems which would act as waveguides: a) channels with longitudinal or transverse partitions on the bottom, and b) channels with curved bottom. The optimum operating conditions are established for these two cases.

621.372.8 2957
Transitions from the TE_{01} Mode in a Rectangular Waveguide to the TE_{11} Mode in a Circular Waveguide—F. Mayer. (*J. Phys. Radium*, vol. 17, Supplement to no. 3, *Phys. Appl.*, pp. 52A–53A; March, 1956.) Brief descriptions are given of a graded cross section coupling of length about 12 cm, and of a $\lambda/4$ transformer giving satisfactory operation over a 500-mc band centered on 9.35 kmc.

621.372.8:538.221:538.614 2958
Ferrites in Waveguides—G. H. B. Thompson. (*J. Brit. IRE*, vol. 16, pp. 311–328; June, 1956.) A survey covering the theory of the gyromagnetic mechanism controlling the microwave permeability of a ferrite, and of wave propagation in circular or rectangular waveguides containing longitudinally or transversely magnetized ferrites. Devices based on resonance absorption or on nonreciprocal transmission are described, including gyrators, isolators, and phase circulators; the different types are compared in respect of ease of construction and performance at a single frequency or over a band. Methods of measuring the components of the ferrite permeability tensor are discussed.

621.372.8:621.318.134 2959
Broad-Band Nonreciprocal Phase Shifts—Analysis of Two Ferrite Slabs in Rectangular Guide—S. Weisbaum and H. Boyet. (*J. Appl. Phys.*, vol. 27, pp. 519–524; May, 1956.) A differential phase shift equalized over a wide frequency band can be produced by using two ferrite slabs of different thickness and magnetic properties but magnetized in the same direction. Analysis is given for the general case. Examination of a particular example indicates that a differential phase shift of π can be obtained constant to within ± 2.5 per cent over the frequency range 5.925–6.425 kmc using ferrite slabs 5.4 inches long in a waveguide 1.59 inches wide.

621.396.674.3 2960
Radiation from an Electric Dipole in the Presence of a Corrugated Cylinder—J. R. Wait. (*Appl. Sci. Res.*, vol. B6, pp. 117–123; 1956.) "A solution is outlined for the problem of an electric dipole which is located outside and parallel to the axis of a circular cylinder of infinite length. The corrugated surface of the cylinder is assumed to be described by an anisotropic boundary impedance which specifies the ratios of the tangential electric and magnetic fields. It is shown that, in general, the radiated field is elliptically polarized."

621.396.674.3:621.396.11 2961
Radiation from a Vertical Antenna over a Curved Stratified Ground—Wait. (See 3190.)

621.396.677.71 2962
Calculated Radiation Characteristics of Slots Cut in Metal Sheets: Part 2—J. R. Wait and R. E. Walpole. (*Canad. J. Technol.*, vol. 34, pp. 60–70; March, 1956.) "Theoretical radiation patterns are presented for antennas consisting of a notch cut in the edge of a perfectly conducting half-plane and a vanishingly thin elliptic cylinder. The principal plane patterns for these two cases are found to be very similar. The conductance of the notch is also considered." part 1: 3160 of 1955. See also 1309 of 1956 (Frood and Wait).

621.396.677.833 2963
Aerial with Wide-Lobe Radiation Pattern—L. Thourel. (*Ann. Radioélect.*, vol. 10, pp. 348–354; October, 1955.) The design of a parabolic-reflector antenna with a sector-shaped radiation pattern is discussed, such as is desirable for long-range surveillance radar installations. Analysis indicates that the optimum radiation pattern for the primary radiator consists of a principal lobe with two counter-phased side lobes; a suitable arrangement for producing such a pattern is a twin-horn radiator. Experimental results supporting the theory are presented.

621.396.677.85 2964
Designing Dielectric Microwave Lenses—K. S. Kelleher. (*Electronics*, vol. 29, pp. 138–142; June, 1956.) Design data for Maxwell, Luneberg, Eaton, Kelleher, and modified types of variable-refractive-index lenses.

AUTOMATIC COMPUTERS

681.142 2965
The Logical Design of an Idealized General-Purpose Computer—A. W. Burks and I. M. Copi. (*J. Franklin Inst.*, vol. 261, pp. 299–314; March, and pp. 421–436; April, 1956.) A detailed discussion emphasizing the distinction between the logic requirements and the particular physical form of a digital computer.

681.142 2966
Analog Computers for the Engineer—J. M. Carroll. (*Electronics*, vol. 29, pp. 122–129; June, 1956.) A review of computer techniques, with tabulated data for some 20 commercially available types.

681.142 2967
Electronic Methods of Analogue Multiplication—Z. Czajkowski. (*Electronic Engng.*, vol. 28, pp. 283–287; July, and pp. 352–355; August, 1956.) A general survey of the principles used; different systems are compared as to accuracy, speed, and complexity.

681.142 2968
High-Speed Electronic-Analogue Computing Techniques—D. M. MacKay. (*Proc. IEE*, part B, vol. 103, pp. 558–559; July, 1956.) Discussion on 3499 of 1955.

681.142 2969
An Analog Computer for the Solution of Tangents—F. S. Preston. (IRE TRANS., vol. EC-4, pp. 101–106; September, 1955.) A modified Wheatstone-bridge arrangement is described, permitting computation of the tangents of angles between 0° and 90°. Only linear elements are used. The accuracy achieved is within 1 part in 2500. The design of plug-in units is discussed.

681.142 2970
Design of Diode Function Simulators—A. D. Talantsev. (*Avtomatika i Telemekhanika*, vol. 17, pp. 129–139; February, 1956.)

681.142:061.3 2971
Digital Computer Techniques—D. B. G. Edwards. (*Nature, Lond.*, vol. 177, pp. 1069–1071; June 9, 1956.) Brief report of a convention held at the Institution of Electrical Engineers, London, in April, 1956. Fifty-eight papers were presented; the full text, together with reports of the discussion, is to be published

in three sections as a supplement to *Proc. IEE*, part B.

681.142:621.374.3 2972

A Variable Multiple Pulse-Stream Generator—W. Woods-Hill. (*Electronic Engng.*, vol. 28, pp. 306–307; July, 1956.) Apparatus designed for checking the logic of electronic computer circuits which require numerous pulse streams for their operation is described. The electrostatic pickup described previously (1632 of 1956) is used.

681.142:621.384.612 2973

Analog Computer for the Differential Equation $y''+f(x)y+g(x)=0$—E. Bodenstedt (*Rev. Sci. Instrum.*, vol. 27, pp. 218–221; April, 1956.) A high-precision electromechanical system developed from that mentioned previously (830 of 1955) uses a torsion pendulum whose motion corresponds to the given expression; the solutions are obtained from photographic records of the motion.

681.142:621.385.132 2974

Binary Adder uses Gas-Discharge Triode—Maynard. (See 3261.)

681.142.002.2 2975

Pulse Circuits fabricate Computer Code Disk—E. M. Jones. (*Electronics*, vol. 29, pp. 146–149; June, 1956.) "Frequency divider, counter, gate, and wave-shaping circuits control optical circle-dividing machine to produce 16-bit pattern on photosensitive glass disk. Used for analog-to-digital conversion, the code disk has a pattern accuracy of ± 0.0001 inch and can be made in about 2 hours." For another account, see *Proc. Nat. Electronics Conf., Chicago*, vol. 11, pp. 288–299; 1955 (Jones, *et al.*).

681.142 2976

An Introduction to Electronic Analogue Computers [Book Review]—C. A. A. Wass. Publishers: Pergamon Press, London, 237 pp.; 1955. (*Brit. J. Appl. Phys.*, vol. 7, p. 157; April, 1956.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.018.3 2977

An Experimental Investigation of Subharmonic Oscillations in a Nonlinear System—K. Göransson and L. Hansson. (*Kungl. Tek. Högsk. Handl., Stockholm*, no. 97, 16 pp.; 1956. In English.) Forced subharmonic oscillations in a circuit containing an iron core are studied. Damping is reduced by means of feedback, so that measurements can be effected at very low driving voltages and subharmonics up to the ninth. Results are in good agreement with theory developed by Lundquist (1269 of 1956) for low driving voltages.

621.316.8:621.372.44:621.314.26 2978

Frequency Conversion with Positive Nonlinear Resistors—C. H. Page. (*J. Res. Nat. Bur. Stand.*, vol. 56, pp. 179–182; April, 1956.) Positive nonlinear resistors are defined as two-terminal devices through which the current I is a real finite single-valued nondecreasing function of the voltage V across the terminals, with the added condition that $I(0)=0$. When subjected to an almost periodic voltage such a resistor will absorb power at some frequencies and supply power at other frequencies. Analysis indicates that modulation efficiency cannot exceed unity, that subharmonics are not produced, and that the efficiency of generating an nth harmonic cannot exceed $1/n^2$.

621.318.4 2979

Winding Focus Coils with Aluminum Foil—(*Electronics*, vol. 29, pp. 244, 252; July, 1956.) Coils of Al foil with a thin coating of Al oxide are wound with no additional insulation.

621.319.4:621.315.615.9 2980

Polychloronaphthalene-Impregnated-Paper Capacitors—J. Coquilion. (*Rev. Gén. Élect.*, vol. 65, pp. 185–193; March, 1956.) Polychloronaphthalene waxes are particularly suitable for use as impregnants in paper-dielectric capacitors, having stable characteristics at temperatures as high as 110°C. or over. In certain cases an aging effect is avoided by allowing the wax to cool from the liquid to the solid state under the influence of an ac or dc field. This phenomenon is discussed in relation to the dipole nature of the waxes.

621.319.4:621.317.3:681.142 2981

Industrial Measurement of the Temperature Coefficient of Ceramic-Dielectric Capacitors—Peyssou and Ladefroux. (See 3136.)

621.319.45 2982

Tantalum Solid Electrolytic Capacitors—D. A. McLean and F. S. Power. (Proc. IRE, vol. 44, pp. 872–878; July, 1956.) A capacitor with low volume/capacitance ratio is obtained by forming a layer of Ta_2O_5 on a porous Ta anode and then depositing a number of coatings of MnO_2 to form a solid electrolyte. The unit is further coated with graphite, and a layer of Pb-Sn alloy is sprayed on to form the cathode. Temperature, frequency, and life characteristics are reported.

621.372 2983

Inter-reciprocity Applied to Electrical Networks—J. L. Bordewijk. (*Appl. Sci. Res.*, vol. B6, pp. 1–74; 1956.) A new concept, "inter-reciprocity," is introduced which is useful in the study of nonreciprocal networks. When a particular topological operation termed "transposition" is performed on a given linear network, the initial network and the resulting transposed network are said to be interreciprocal. Application of the theory to a variety of general and special circuit problems is illustrated; the noise properties of gyrator, triode, and transistor networks are discussed.

621.372:621.3.018.752:621.397.8 2984

The Effect upon Pulse Response of Delay Variation at Low and Middle Frequencies—M. V. Callendar. (*Proc. IEE*, part B, vol. 103, pp. 475–478; July, 1956.) "Calculations are given for the magnitude and form of the distortion introduced into a square wave by a network or system which exhibits uniform transmission except for increasing (or decreasing) phase delay in the low-midfrequency region. The fractional peak distortion is found to be equal to twice the area under the curve relating T_n to frequency, where T_n is the delay relative to that at high frequencies. The waveform of the distortion is given for several simple shapes of curve for T_n. This distortion is especially characteristic of vestigial-sideband systems, and occurs in television as a 'preshoot' before a transition and as a smear (in principle equal, but opposite, to the preshoot) after it."

621.372.012 2985

Feedback Theory—Further Properties of Signal Flow Graphs—S. J. Mason. (Proc. IRE, vol. 44, pp. 920–926; July, 1956.) Continuation of theory presented previously (3531 of 1953).

621.372.41:621.318.424 2986

Transient Behavior in a Ferroresonant Circuit—J. G. Skalnik. (*J. Appl. Phys.*, vol. 27, pp. 508–513; May, 1956.) An analysis is made of the response to a sinusoidal voltage of a circuit including a nonlinear inductor. For certain frequency values of the applied voltage there are three possible values for the flux in the inductor, of which the middle value is unstable. The differential equation representing the circuit has been solved using an analog computer. For the case when the system is released in the region of the lower stable state, the solution corresponds to two sinusoidal oscillations of different amplitude and frequency. If the system is released in the region of the upper stable state, the solution corresponds to an oscillation modulated in amplitude and phase, for certain values of the parameters.

621.372.413:621.317.337 2987

Measurement of the Q-Factor of Cavity Resonators, using a Straight Test Line—Urbarz. (See 3141.)

621.372.44:621.372.6 2988

Some General Properties of Nonlinear Elements: Part 1—General Energy Relations—J. M. Manley and H. E. Rowe. (Proc. IRE, vol. 44, pp. 904–913; July, 1956.) An analysis is made of power relations in networks with reactive nonlinear elements. Two equations are derived relating the powers at different frequencies; the only assumption introduced is that the nonlinear characteristic is single-valued. The theory is relevant to the operation of modulators, demodulators, and harmonic generators.

621.372.5(083.5) 2989

Tables of Phase of a Semi-infinite Unit Attenuation Slope—D. E. Thomas. (*Bell Syst. Tech. J.*, vol. 35, pp. 747–749; May, 1956.) The five-figure tables published previously (968 of 1948) are to appear together with newly prepared seven-figure tables as *Bell System Monograph* 2550.

621.372.51:621.372.22 2990

Fundamentals in the Synthesis of Loss-Free Quadripoles from Lines with Continuous Non-uniformities—H. Meinke. (*Nachrichtentech. Z.*, vol. 9, pp. 99–106; March, 1956.) The synthesis is facilitated by appropriate choice of line coordinates and a polynomial representation of the characteristic impedance. Application to problems of wide-band transformation and matching are illustrated.

621.372.51:621.396.67 2991

Impedance Quadripoles for the Frequency Compensation of Aerial Input Impedance—R. Herz. (*Nachrichtentech. Z.*, vol. 9, pp. 128–133; March, 1956.) Networks with one or two frequency-independent resistances are discussed which are capable of effecting wide-band matching with lower losses than reactive circuits at frequencies up to 1 kmc or above. Composite coaxial-line sections are used; in an application to a dipole antenna for use with a parabolic-cylinder reflector, the compensating coaxial line serves as support for the dipole.

621.372.54:621.375.132:621.3.018.75 2992

Normalized Representation of Transients in Filter Amplifiers with Double-T Elements—H. Dobesch. (*Hochfrequenztech. u. Elektroakust.*, vol. 64, pp. 102–107; January, 1956.) The response of amplifiers with frequency-dependent negative feedback is analyzed for various pulse and step waveforms; the frequency spectrum corresponding to a train of square pulses is determined.

621.372.542.2 2993

A Solution to the Approximation Problem for RC Low-Pass Filters—K. L. Su and B. J. Dasher. (Proc. IRE, vol. 44, pp. 914–920; July, 1956.) A method of synthesizing filters is described in which elliptic functions are used to effect a transformation in the complex-frequency plane which results in a symmetrical arrangement of the zeros and poles. Some design charts are included.

621.372.57:[621.385+621.314.7 2994

A Particular Case of the Application of the Matrix Method to Radio Engineering—B. Ya. Yurkov. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1988–1993; October, 1955.) Use of the matrix method in analysis of the operation of quadripoles including thermionic tubes or transistors is discussed. A simple method is proposed for carrying out the necessary transformations to the formulas on passing from the one case to the other.

621.373+621.375.9]:538.561.029.6 2995
Application of Electron Spin Resonance in a Microwave Oscillator or Amplifier—Combrisson, Honig, and Townes. (See 3032.)

621.373.421 2996
Constant-Frequency Oscillators—L. B. Lukaszewicz. (*Wireless Engr.*, vol. 33, pp. 201–202; August, 1956.) Comment on 697 of 1956 (Gladwin).

621.373.421 2997
Bridge-Stabilized Oscillators and their Derivatives—E. J. Post and J. W. A. van der Scheer. (*J. Brit. IRE*, vol. 16, pp. 345–350; June, 1956.) Reprinted from *PTT-Bedrijf*, vol. 6, September, 1955.) General analysis is presented for the operation of the bridge-stabilized feedback oscillator, and modifications obtained by interchanging bridge elements crosswise or by unbalancing the bridge are discussed.

621.373.421.13:621.372.412:621.316.726 2998
Frequency Stability and Quartz-Controlled Oscillators—A. Erkens. (*Ann. Radioélect.*, vol. 10, pp. 399–405; October, 1955.) The operation of some commonly used types of crystal-controlled oscillator is reviewed. Frequency can be held constant to within a factor of 10^{-4} over a period of months by using a Y-bar crystal resonator.

621.373.431.1 2999
Bistable Circuits using Triode-Pentodes—H. L. Armstrong. (*Electronics*, vol. 29, pp. 210, 214; July, 1956.) Note on the operation of multivibrator-type circuits in which one feedback path is provided by connecting triode anode to pentode screen, leaving one grid free for triggering, gating, or modulation.

621.373.432 3000
Simple Method for producing H.F. Pulses of Short Duration and Large Amplitude—A. V. J. Martin. (*J. Phys. Radium*, vol. 17, p. 310; March, 1956.) Pulses of duration about 10 μs and peak-to-peak amplitude about 240 v are obtained from the tuned secondary of a transformer in the cathode circuit of a thyratron.

621.373.52+621.375.4]:621.314.7 3001
Applications for Tandem Transistors—H. E. Hollmann. (*Tele-Tech and Electronic Ind.*, vol. 15, pp. 58–59, 114; February, 1956.) The tandem transistor, consisting of two transistors housed in a single container and cascaded so that one acts as the base leak for the next, may be used as an amplifier with high input impedance and in various applications in which single grounded-emitter stages are normally used.

621.373.52.029.3 3002
Superregenerative Transistor Oscillator—R. J. Kircher and I. P. Kaminow. (*Electronics*, vol. 29, pp. 166–167; July, 1956.) The circuit described generates pulses of 500-cps tone at a rate of 7 per second. The performance with different values of quench capacitor, bias, feedback, etc. is shown graphically.

621.374 3003
Investigation of Special Frequency Dividers with Large Dividing Ratio—E. O. Philipp. (*Z. Angew. Phys.*, vol. 8, pp. 119–126; March, 1956.) Two frequency dividers and one pulse counter are developed on the basis of Kroebel's work (383 of 1955). These give stable dividing ratios of 100 and 200 at input pulse frequencies of 4 mc and 31.25 kc respectively. The counter can handle irregular pulses spaced at intervals of 1–50 ms.

621.374.3:621.385.5.032.24 3004
A New High-Slope Multigrid Valve and its Application in Pulse and Switching Circuits—Gosslau and Guber. (See 3262.)

621.374.32:621.314.7 3005
A Point-Contact Transistor Scaling Circuit with 0.4-μs Resolution—G. B. B. Chaplin. (*Proc. Inst. Elect. Engrs*, part B, vol. 103, pp. 505–509; July, 1956. Discussion, pp. 516–518.) Simple circuits using normal point-contact transistors are described; features contributing to the short resolving time are the prevention of bottoming of collector potential and the absence of capacitors. A typical scale-of-ten circuit uses seven transistors, seven pulse transformers, and fourteen crystal diodes. Wide tolerances on the transistor parameters are permissible.

621.374.32:621.314.7 3006
A Junction-Transistor Scaling Circuit with 2-μs Resolution—G. B. B. Chaplin and A. R. Owens. (*Proc. IEE* part B, vol. 103, pp. 510–515; July, 1956. Discussion, pp. 516–518.) The basic circuit discussed is a binary scaler using a differentiating transformer instead of capacitors for coupling; the speed of operation thus depends only on the transistor characteristics. Scale-of-5 and scale-of-10 circuits built up from the basic circuit are described.

621.375.2:621.385.3.029.63 3007
Disc-Seal Triode Amplifiers—G. Craven. (*Wireless Engr.*, vol. 33, pp. 179–183; August, 1956.) "The design of a resonant π-type coupling network for disc-seal triodes operating in the earthed-grid connection at frequencies in the range 300–3000 mc is considered. A coaxial form of line is adopted. Tuning for a small range can be by a 'screw' or, for a larger range, by a built-in capacitance. Complete amplifiers can give 100-w output and 30-db gain using three stages."

621.375.2:621.385.5:621.314.7 3008
Higher Pentode Gain—L. Levy. (*Electronics*, vol. 29, pp. 190, 196; July, 1956.) Note on the use of a transistor as an anode load.

621.375.232.029.3:621.396.822 3009
Noise in an Amplifier Stage with Negative Voltage Feedback—H. Nottebohm. (*Elektronische Rundschau*, vol. 10, pp. 57–62; March, 1956.) The problem is considered with particular reference to the input circuit of an amplifier for a magnetic tape recorder. Analysis indicates that frequency distortion inherent in the system can be corrected by use of negative feedback at the input tube, and indicates the existence of an optimum ratio for the input transformer, from the point of view of signal/noise ratio.

621.375.232.3.029.3 3010
Triode Cathode-Followers for Impedance Matching to Transformers and Filters—T. J. Schultz. (IRE Trans., vol. AU-3, pp. 28–37; March/April, 1955.) Design curves derived from measurements on five different types of triode are presented.

621.375.232.9 3011
An Improved Type of Differential Amplifier—J. C. S. Richards. (*Electronic Engng.*, vol. 28, pp. 302–305; July, 1956.) "A differential amplifier stage capable of giving a high rejection ratio with unselected tubes and components and without a balance control is analyzed, and a particular amplifier is described in some detail. The stage is particularly suitable for converting balanced to unbalanced signals."

621.375.3 3012
Comparison of Some Magnetic-Amplifier Circuits with Internal Feedback—A. B. Gorodetski. (*Avtomatika i Telemekhanika*, vol. 17, pp. 147–159; February, 1956.)

621.375.3 3013
Push-Pull Magnetic Amplifier with Direct-Current Output—R. Kh. Bal'yan. (*Avtomatika i Telemekhanika*, vol. 17, pp. 160–171; February, 1956.)

621.375.3:621-526 3014
Decicyle Magnetic-Amplifier Systems for Servos—L. J. Johnson and S. E. Rauch. (*Elect. Engng, N. Y.*, vol. 75, p. 243; March, 1956.) Digest of paper published in *Trans. AIEE part I, Communication and Electronics*, vol. 74, pp. 667–672; 1955. Improvements in circuitry and core materials, and the adoption of pulse techniques, make possible systems whose response times are one tenth to one hundredth of a cycle of the power-supply frequency.

621.376.22:621.318.134 3015
A Ferrite Microwave Modulator employing Feedback—W. W. H. Clarke, W. M. Searle, and F. T. Vail. (*Proc. IEE*, part B, vol. 103, pp. 485–490; July, 1956.) An amplitude modulator with good linearity is obtained by applying feedback to a gyrator comprising a ferrite rod in a circular waveguide section interposed between rectangular waveguide sections. The feedback circuit is based on linear detection of the amplitude modulation by means of a crystal. Limitations of the arrangement are discussed. Good results have been obtained with sinusoidal modulating signals of frequencies up to 20 kc.

621.37/.39(083.74) 3016
Handbook Preferred Circuits, Navy Aeronautical Equipment, NAVAER 16-1-519 [Book Review]—J. C. Muncy. Publishers: Government Printing Office, Washington, D. C. (*Tech. News Bull. Nat. Bur. Stand.*, vol. 40, pp. 66–67; May, 1956.) Gives design details and characteristics of the standardized circuits discussed previously (342 of 1956). Supplements are to be issued from time to time.

621.375.13 3017
Linear Feedback Analysis [Book Review]—J. G. Thomason. Publishers: Pergamon Press, London, 365 pp. (*J. IEE*, vol. 2, p. 187; March, 1956.) A useful introduction to the subject.

GENERAL PHYSICS

537:538.56 3018
Electron Plasma Oscillations in an External Electric Field—A. I. Akhiezer and A. G. Sitenko. (*Zh. Eksp. Teor. Fiz.*, vol. 30, pp. 216–218; January, 1956.) The oscillation frequency is calculated, assuming that the electron distribution function satisfies a given kinetic equation.

537.2 3019
Fields and Stresses in Dielectric Media—G. Power. (*Brit. J. Appl. Phys.*, vol. 7, pp. 137–144; April, 1956.) Expressions are obtained for the mechanical forces at the boundary of an isotropic dielectric, caused by an electric field. Results are verified in particular cases by electrolyte-tank experiments.

537.311.1 3020
On the Energy Dissipation of Conduction Electrons Undergoing Elastic Scattering by Impurities—T. Yamamoto, K. Tani, and K. Okada. (*Progr. Theor. Phys.*, vol. 15, pp. 184–185; February, 1956.) A brief theoretical note on the mechanism responsible for the energy dissipation in conduction in metals.

537.311.31:537.312.8 3021
Theory of Galvanomagnetic Phenomena in Metals—I. M. Lifshits, M. Ya. Azbel', and M. I. Kaganov. (*Zh. Eksp. Teor. Fiz.*, vol. 30, pp. 220–222; January, 1956.) The theory is developed without making any special assumptions regarding the conduction-electron dispersion law and the form of the collision integral.

537.311.62 3022
Anomalous Skin Effect Assuming Arbitrary Collision Integral—M. Ya Azbel', and E. A. Kaner. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 876–878; December, 1955.) Results of a calculation

of the surface impedance $Z_\alpha = R_\alpha + iX_\alpha$, show that the ratio X_α/R_α equals $\sqrt{3}$ for an arbitrary electron-dispersion law and an arbitrary collision integral; Z_α is proportional to $\omega^{2/3}$ and is independent of temperature in the anomalous skin-effect temperature range.

537.5 3023
Statistics of Electron Avalanches in a Uniform Field—L. Frommhold. (*Z. Phys.*, vol. 144, pp. 396–410; February 7, 1956.) The statistical distribution of the number of charge-carrier pairs about the mean was determined experimentally by measurements on discharges in ethyl alcohol. Results agree with theory.

537.523 3024
Surge Voltage Breakdown of Air in a Nonuniform Field—J. H. Park and H. N. Cones. (*J. Res. Nat. Bur. Stand.*, vol. 56, pp. 201–223; April, 1956.) Experiments on discharges between a spherical and a plane electrode are described, and a tentative explanation of the breakdown mechanism is presented.

537.525:538.56.029.5 3025
Investigation of a Discharge in the Frequency Region between High Frequency and Audio Frequency at Low Gas Pressure—N. A. Popov and N. A. Kaptsov. (*Zh. Eksp. Teor. Fiz.*, vol. 30, pp. 68–76; January, 1956. English summary, *ibid.*, Supplement, p. 5.)

537.525:538.56.029.6 3026
Investigation of the High-Frequency Discharge—G. M. Pateyuk. (*Zh. Eksp. Teor. Fiz.*, vol. 30, pp. 12–17; January, 1956. English summary, *ibid.*, Supplement, p. 3.) The dependence of the ignition and operating potentials in Ar, Ne, and H_2 on the gas pressure and the geometry of the discharge space was investigated in the frequency range 57–500 mc. Results, presented graphically, are in agreement with the diffusion theory of Herlin and Brown (690 of 1949).

537.533 3027
Influence of an Adsorbed Film of Dipole Molecules on the Electron Work Function of a Metal—N. D. Morgulis and V. M. Gavrilyuk. (*Zh. Eksp. Teor. Fiz.*, vol. 30, pp. 149–159; January, 1956. English summary, *ibid.*, Supplement, p. 7.) Experimental results indicate that films of CsCl molecules decrease the work function of w by up to 1.8 ev, as compared with a decrease of up to 3 ev produced by Cs, 3.5 ev by BaO and 2.9 ev. by Ba.

537.533.8 3028
Auger Electron Emission in the Energy Spectra of Secondary Electrons from Mo and W—G. A. Harrower. (*Phys. Rev.*, vol. 102, pp. 340–347; April 15, 1956.) Analysis of the observed energy distributions of the secondary electrons for a range of primary energies reveals subsidiary maxima at points along the energy axis characteristic of the target material but independent of the primary voltage; the positions of these points are consistent with an Auger-process origin for the electrons with these energies.

537.533.8:546.561-31 3029
Investigation of the Inelastic Reflection of Electrons by a Cuprous Oxide Surface—N. B. Gornyi. (*Zh. Eksp. Teor. Fiz.*, vol. 30, pp. 160–170; January, 1956. English summary, *ibid.*, Supplement, pp. 7–8.) The energy losses of electrons on reflection at monocrystalline or polycrystalline Cu_2O surfaces are equal to the energy required to transfer electrons of the crystal lattice from filled to permitted zones. The mechanism involved is similar to that responsible for the appearance of discrete groups of true secondary electrons (685 of 1955).

538.311 3030
Production and Use of High Transient Magnetic Fields: part 1.—H. P. Furth and R. W. Waniek. (*Rev. Sci. Instrum.*, vol. 27, pp. 195–203; April, 1956.) Technique for the production of pulsed magnetic fields of strength 5×10^5 G or over is discussed; capacitor-discharge arrangements are used, with impact-resistant solenoids comprising massive single-layer helices. Pulse durations range from 50 μs to 10 ms. Measurement of the magnetoresistance of Ge is one of the applications mentioned.

538.56:53 3031
Radio-Frequency Physics—J. G. Powles. (*Nature, London*, vol. 177, pp. 1022–1023; June 2, 1956.) Brief report of the 1956 annual conference, held at Geneva of the organization A.M.P.É.R.E. (Atomes et Molécules par Études Radioeléctriques), which is concerned with the use of radio frequencies in the various branches of physics. Some 50 papers were presented, the subjects including dielectric and magnetic properties, electron resonance of various types and associated effects, and microwave spectroscopy. See also *Onde Élect.*, vol. 35, pp. 437–505; May, 1955, which gives papers from the 1954 conference, held at Paris, covering a similar range of subjects and including also some material on atmospheric physics.

538.561.029.6:[621.373+621.375.9 3032
Application of Electron Spin Resonance in a Microwave Oscillator or Amplifier—J. Combrisson, A. Honig, and C. H. Townes, (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2451–2453; May 14, 1956.) A brief analysis indicates the condition for a paramagnetic substance located within a cavity resonator and subjected to a direct magnetic field to supply power instead of absorbing it from the field. Results of preliminary experiments indicate that it should be possible to produce oscillations using, *e.g.*, a small sample of Si containing a suitable impurity providing a concentration of about 10^{17} paramagnetic centers per $cm.^3$

538.566:535.337 3033
Radiation from Molecules in the Presence of a Strong High-Frequency Field—V. M. Fain. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 878–880; December, 1955.) It is shown that in addition to an absorption of the hf energy at a frequency $\omega \cong \omega_0 = (E_1 M E_2)/h$, where E_1 and E_2 are energy levels of the molecule, emission takes place at a frequency Ω_0 which is a function of the matrix element $\vec{\mu}_{12}$ of the dipole moment corresponding to the transition $E_1 \rightarrow E_2$ and the field strength of the hf field. In a typical case $|\mu_{12}| \cong 10^{-18}$ c.g.s.e. and the field strength is 1–10 c.g.s.e.; the value of Ω_0 is then approximately 10^9 $second^{-1}$–10^{10} $second^{-1}$. The radiation is only present if the elements $\vec{\mu}_{11}$ and $\vec{\mu}_{12}$ are not equal to zero.

538.566:535.42]+534.24 3034
Fourier Transform Method for the Treatment of the Problem of the Reflection of Radiation from Irregular Surfaces—C. Meecham. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 370–377; May, 1956.)

538.566:537.533.9 3035
Incidence of an Electromagnetic Wave on a 'Čerenkov Electron Gas'—M. A. Lampert. (*Phys. Rev.*, vol. 102, pp. 299–304; April 15, 1956.) Analysis is presented for the interaction of an em wave in a retarding medium (*e.g.*, a dielectric) with an electron gas moving through or near the medium at a velocity exceeding that of the wave in the medium. For electron densities exceeding a critical value, the gas acts as a mirror for the incident em wave. Possible laboratory experiments for investigating the problem are outlined.

538.566.2 3036
Method of Calculating Electromagnetic Fields Excited by an Alternating Current in Stratified Media—A. N. Tikhonov and D. N. Shakhsuvarov. (*Bull. Acad. Sci. U.R.S.S., Sér. Geophys.*, no. 3, pp. 245–251; March, 1956. In Russian.) The expressions for the field due to a dipole in the boundary of a stratified half-space are developed in a form suitable for evaluation by a modern computer. The em characteristics of the strata are assumed to be independent of time and of the field; the permeability is constant and the conductivities are arbitrary; the conductivity of the surface layer is finite.

538.569.2/.3].029.6:534.833.4 3037
Absorption Devices for Centimetre Electromagnetic Waves and their Acoustic Analogues—E. Meyer and H. Severin. (*Z. Angew. Phys.*, vol. 8, pp. 105–114; March, 1956.) A survey of the operating mechanism of three types of absorbers: a) homogeneous material, b) wedges, and c) resonance absorbers.

538.6:537.311.31 3038
Thermo- and Galvano-magnetic Effects in Strong Fields at Low Temperatures—G. E. Zil'berman. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 762–769; December, 1955.) Thermoelectric force, resistance, and Hall effect of a metal in a magnetic field at low temperatures are calculated using a two-zone model of the metal.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 3039
An Investigation of Monochromatic Radio Emission of Deuterium from the Galaxy—G. J. Stanley and R. Price. (*Nature, London*, vol. 177, pp. 1221–1222; June 30, 1956.

525.2:523.2 3040
Gravitational Influence of Jupiter on some Geophysical Phenomena—D. Argentieri. (*Ann. Geofis.*, vol. 8, pp. 457–473; October, 1955.) Consideration of astronomical observations from ancient times onwards has indicated apparent variations in astronomical time. Attention is drawn particularly to a variation having a period of 83 years; this is also the period taken by the sun, the earth, and Jupiter to return to the same alignment and relative distance. It is suggested that the combined gravitational action of the sun and Jupiter causes tidal motion in the earth's crust, the apparent variation of astronomical time corresponding to a real displacement of the meridian.

551.510.5:538.569.4.029.6:523.72 3041
Atmospheric Attenuation of Solar Millimeter-Wave Radiation—H. H. Theissing and P. J. Caplan. (*J. Appl. Phys.*, vol. 27, pp. 538–543; May, 1956.) Measurements have been made of the absorption of solar radiation by atmospheric water vapor at wavelengths down to about 1 mm. The results are combined with theoretical formulas for the absorption spectrum of water vapor [see, *e.g.*, 3100 of 1947 (Van Vleck)].

551.510.53:551.593 3042
Origin of the Meinel Hydroxyl System in the Night Airglow—D. R. Bates and B. L. Moiseiwitsch. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 305–308; June, 1956.)

551.510.53:551.593+551.594.5]:535.241 3043
A Photometric Unit for the Airglow and Aurora—D. M. Hunten, F. E. Roach, and J. W. Chamberlain. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 345–346; June, 1956.)

551.510.534 3044
Note on the Variations of Atmospheric Ozone as a Function of Height—E. S. Epstein,

C. Osterberg, and A. Adel. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 347–348; June, 1956.) Observations confirming those of Paetzold (748 of 1956) are reported.

551.510.535 3045

Symposium on Ionospheric Drifts—(*J. Sci. Industr. Res.*, vol. 14A, pp. 482–485; October, 1955.) Brief report of symposium held at New Delhi in July, 1955.

551.510.535 3046

Accurate Height Measurements using an Ionospheric Recorder—A. J. Lyon and A. J. G. Moorat. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 309–317; June, 1956.) "A method for the calibration of an ionospheric recorder is described, which corrects errors in height measurement arising from the distortion of the echo-pulse in its passage through the receiver. The amount of this error depends on the echo-amplitude, and is shown to vary in an approximately linear manner with the width of the recorded echo-trace. Several methods of checking the calibration confirm that it is reliable to within ±2 km. Using a calibrated recorder and an expanded timebase, it is possible to measure E-region equivalent heights to this order of accuracy. The systematic error due to pulse distortion will, in general, cause the heights recorded in routine ionospheric measurements to be from 5 to 15 km too high. Some consequences of this, *e.g.*, for muf predictions, are mentioned."

551.510.535 3047

Monthly Mean Values of Ionospheric Characteristics at Rome in the Period March 1949–April 1953—P. Dominici. (*Ann. Geofis.*, vol. 8, pp. 379–400; October, 1955.) Hourly values are tabulated for the critical frequency and virtual height of the F_2, F_1, and E layers and for the percentage of occurrences of the E_s layer. Brief particulars are given of the sounding schedule operated and the conventions adopted in the calculations.

551.510.535 3048

Sporadic Echoes from the E Region over Ahmdabad (23° 02′ N, 72° 38′ E)—K. M. Kotadia. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 331–337; June, 1956.) An analysis is made of $h'f$ records for the sunspot-minimum period 1953–1954. The diurnal and seasonal variations of E_s as a whole are interpreted as variations in the relative contributions of three distinct types of E_s, namely a) E_{se}, a thin layer observed at 95–100 km, with a maximum frequency of occurrence at late evening, b) E_{sn}, which is observed at 105–125 km with a minimum in the afternoon and maximum towards the end of the night, and c) E_{ss}, at 115–125 km, developed by the vertical downward movement of the E_2 layer and observed only during the daytime.

551.510.535 3049

A New Theory of Formation of the F_2 Layer—T. Yonezawa. (*J. Radio Res. Labs, Japan*, vol. 3, pp. 1–16; January, 1956.) Electron/ion pairs generated in the upper part of the F_2 region diffuse rapidly downwards under gravity, but at sufficiently low heights they are rapidly lost by the mechanism of charge transfer and dissociative recombination suggested by Bates and Massey (1944 of 1948), giving rise to a maximum ionization density at a greater height. This theory gives results in accordance with observations.

551.510.535 3050

The Structure of the F_2 Layer as deduced from its Daily Variations—T. Shimazaki. (*J. Radio Res. Labs, Japan*, vol. 3, pp. 17–43; January, 1956.) Observed variations in the F_2 region may be accounted for by assuming that a) in consequence of the decrease with height of the effective decay coefficient, the maximum electron density in the F_2 region is at a level above that of maximum electron production, and that b) vertical semidiurnal tidal drift is nonuniform. At the level of maximum electron production the rate of production varies inversely as temperature. An attachment coefficient of 8.3×10^{-5}/second at 300 km is indicated, with a solar temperature of 6000° K.

551.510.535 3051

Geomagnetic Control to the Diurnal Variation of the F_2 Layer on the Temperate Latitude—Syun-ichi Akasofu. (*Sci. Rep. Tohoku Univ.*, *5th Ser.*, *Geophys.*, vol. 7, pp. 45–50; November, 1955.) The "longitude effect" demonstrated by Appleton (882 of 1951) is examined. The observed diurnal variation is consistent with geomagnetic control of the thermal vertical flow in the F_2 region. Seasonal variations are also observed.

551.510.535:523.746.5 3052

Comparison of f_0F_2 at Four Observatories in Japan—I. Kasuya and K. Sawada. (*J. Radio Res. Labs., Japan*, vol. 3, pp. 45–53; January, 1956.) Observations over the solar-activity half-cycle 1947–1954 are correlated with sunspot numbers. On a long-term basis, the magnitude of the variation of f_0F_2 is a function of latitude.

551.510.535:537.56 3053

Negative Oxygen Ions in the Upper Atmosphere: the Affinity and Radiative Attachment Coefficient of Atomic Oxygen—L. M. Branscomb and S. J. Smith. (*Trans. Amer. Geophys. Union*, vol. 36, pp. 755–758; October, 1955.) "The influence of negative ions of atomic oxygen on the physics of the ionosphere and night airglow is re-examined in the light of new experimental determinations of the oxygen affinity (1.48 ±0.10 ev) and photodetachment cross section [396 of 1956 (Smith and Branscomb)]. The radiative attachment coefficient is calculated from the photodetachment cross section. There is no evidence of a resonance at the threshold, where the attachment coefficient is approximately 1.2×10^{-15} cm³/second."

551.510.535:621.396.11 3054

Observations of Ionospheric Absorption at the K.N.M.I. [Royal Netherlands Meteorological Institute]—C. J. van Daatselaar. (*Tijdschr. Ned. Radiogenoot.*, vol. 21, pp. 49–63; March, 1956.) Theory of ionospheric absorption is outlined and measurement difficulties due to fading are discussed. The procedure at the Netherlands station is to determine the apparent reflection coefficient for vertically incident waves, using pulse transmissions with cro display of the echo amplitude; total absorption and absorption index are hence derived. The equipment is described and some results are reported.

551.510.535:621.396.11 3055

On the Existence of a 'Q.L.'–'Q.T.' 'Transition-Level' in the Ionosphere and its Experimental Evidence and Effect—Lepechinsky. (*J. Atmos. Terr. Phys.*, vol. 8, No. 6, pp. 297–304; June 1956.) See 1767 of 1955 (Lepechinsky and Durand).

551.510.535:621.396.11.029.55 3056

Back-Scatter Ionospheric Sounder—Shearman and Martin. (See 3197.)

551.510.535:621.396.812.3 3057

A Correlation Treatment of Fading Signals—Barber. (See 3200.)

551.594.6 3058

On the Propagation of Whistling Atmospherics—G. R. Ellis. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 338–344; June, 1956.) "It is shown that the dispersion of whistling atmospherics propagated along the lines of force of the earth's magnetic field should be greatly dependent on the geomagnetic latitude of the observing point. The change in the magnetic-field intensity along a line of force produces an upper-frequency limit for whistler propagation which, at latitudes greater than 62°, should fall within the usually observed frequency region of 1–10 kc. Dispersion curves showing this critical frequency are given for geomagnetic latitudes 55°, 60°, and 65°."

551.594.6:523.75 3059

Sudden Decrease in Low-Frequency Atmospheric Noise during the Cosmic-Radiation Storm of February 23—C. A. McKerrow. (*Nature, London*, vol. 177, pp. 1223–1224; June 30, 1956.) Note of observations on 100 kc at Churchill, Manitoba. The relation of the disturbance to solar-flare conditions and to the proximity of the station to the auroral zone is briefly discussed.

551.594.6:538.566.029.43 3060

Influence of the Horizontal Geomagnetic Field on Electric Waves between the Earth and the Ionosphere Travelling Obliquely to the Meridian—W. O. Schumann. (*Z. Angew. Phys.*, vol. 8, pp. 126–127; March, 1956.) A more general case than that noted earlier (232 of 1956) is considered briefly. Results indicate that the differences in the type of atmospherics waveform arriving from south-east and from south-west [2809 of 1952 (Caton and Pierce)] are due to differences not in the propagation but in the nature of the discharge, which may occur over the sea in one case, over land in the other.

551.594.6:550.385 3061

The Low-Frequency Noise of the Geomagnetic Field—R. Benoit. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2534–2535; May 23, 1956.) Grenet's investigations of the source of af disturbances (1718 of 1956) are discussed. Observations made in the Sahara are reported; a telephone cable formed into a circular loop of diameter 300 m was used as antenna, in conjunction with a multistage amplifier and pen recorder; the total frequency band explorable was 10 cps–50 kc. The results indicate that the low-frequency pulses received are almost entirely due to atmospherics; this confirms Grenet's theory.

551.594.6:550.385 3062

Electromagnetic Phenomena of Natural Origin in the 1.0–150-c/s Band—P. A. Goldberg. (*Nature, London*, vol. 177, pp. 1219–1220; June 30, 1956.) A report is presented of observations made at an isolated region in Oregon in the summer of 1955. Air-core detector coils were used, one with an effective area of 12,800 m² for observing the vertical magnetic-field component, and another with an effective area of 5500 m² for the horizontal north-south component. Voltage waveforms proportional to the field and to its time rate of change were studied by means of photographic records from cr oscillographs. The signals recorded are predominantly of burst type, the horizontal component being more intense than the vertical. The level of activity exhibits a systematic diurnal variation. Comparison with the incidence of rf atmospherics suggests that the low-frequency signals are associated with lightning, while the timing of the daytime maximum indicates that the propagation mechanism is different from that for the rf atmospherics.

551.594.6:621.396.11 3063

The Propagation of a Radio Atmospheric—Srivastava. (See 3189.)

551.594.6:621.396.11.029.4 3064

Propagation of Audio-Frequency Radio Waves to Great Distances—Chapman and Macairo. (See 3192.)

LOCATION AND AIDS TO NAVIGATION

621.396.93 3065

Fluctuations in Continuous-Wave Radio Bearings at High Frequencies—W. C. Bain. (*Proc. IEE*, part B, vol. 103, p. 560; July, 1956.) The investigation reported previously (3265 of 1955), covering the frequency band 6–20 mc, is extended to cover the band 3–4 mc. The results differ from those obtained previously in that the standard deviation in a group of observations is not correlated significantly with the value of τ_0.

621.396.93 3066

The 'Wullenwever' Long-Base Direction-Finding Installation—H. Rindfleisch. (*Nachrichtentech. Z.*, vol. 9, pp. 119–123; March, 1956.) This system was developed during the war and is described in *Radio Research Special Report* No. 21, 1951, *Radio Direction Finding and Navigational Aids; some Reports on German Work issued in 1944–45.*

621.396.96:519.21:621.396.822 3067

Connection between the Detectability of an Object and the Number of Illuminating Pulses—G. N. Bystrov. (*Radiotekhnika, Moscow*, vol. 11, pp. 74–76; February, 1956.) The probability P, that a blip on the cr tube display is due to the object and not to the noise is $P=1-\exp(-na_v^2/2a_0^2)$, where n is the number of radar pulses, a_v the amplitude of the blip, and a_v the mean effective noise voltage. A Rayleigh-type noise-voltage amplitude distribution in the output of the second detector is assumed. The probability of detecting an object is then calculated in terms of the distance, transmitter power, antenna gain, wavelength, surface area of object and power input to receiver, as well as the absolute temperature, pass band and receiver noise factor.

621.396.96:621.316.726:621.385.029.6 3068

Klystron Control System—R. J. D. Reeves. (*Wireless Engr.*, vol. 33, pp. 135–143, 162–167; June, and pp. 184–189; August, 1956.) Wide-range tuning of reflex klystrons is discussed with particular reference to an afc system for primary radar. The problem is complicated because the optimum reflector voltage is not independent of the resonator frequency. The concept of a "control plane" is introduced to facilitate analysis of the klystron operation. Test equipment is described which presents the control plane on a cro and maps either klystron mode areas and frequency contours or servo-trajectories on to the plane. In the particular afc system described in detail, a sampling technique for mode centering is introduced which causes minimum disturbance of the controls and provides a slightly better error criterion than mode peak finding.

621.396.962.2:621.376.3:629.13 3069

A Frequency-Modulation Radio Altimeter—G. Collette and R. Labrousse. (*Ann. Radioélect.*, vol. 10, pp. 387–398; October, 1955.) The Type-AM.210 altimeter is discussed; the range is 1500 m and the frequency band 420–460 mc; the modulating function is a symmetrical sawtooth repeated 4050 or 810 times per minute. The problem of coupling between the slotted-cavity antennas is examined, and suitable values of antenna spacing and feeder length are indicated.

621.396.963.001.4:534.21-8 3070

Variable Delay Line Simulates Radar Targets—S. A. Gitlin. (*Electronics*, vol. 29, pp. 143–145; June, 1956.) "Two quartz transducers and movable corner reflector in 3½-ft water-filled copper tank give time delays ranging from 72 to 1400 μs for simulating moving targets during tests of new radar."

621.396.963.33.001.4 3071

Three-Dimensional Radar Video Simulator—P. Pielich. (*Electronics*, vol. 29, pp. 131–133; July, 1956.) Terrain is represented on a test slide with six contour lines defining range and azimuth at six heights. The slide is scanned by a flying-spot system and x, y, z voltages from the simulator unit are combined to give appropriate, X, Y deflection voltages for a cro. Detailed circuit diagrams are given.

621.396.969 3072

Frequency-Modulation Radar for Use in the Mercantile Marine—D. N. Keep. (*Proc. IEE*, part B, vol. 103, pp. 519–523; July, 1956. Discussion, pp. 523–526.) "The principles of fm radar are outlined and a comparison is made between pulse and fm techniques, particularly with respect to the requirements of the merchant service. It is concluded that multigate fm radars are too complex for this application and methods are outlined for overcoming the inherently low scanning rate of single sweeping-gate systems. Equipment is described which has an antenna beamwidth of 1.7° and a rotation rate of 10 rpm with a fractional range resolution of 1/30. The future of fm radar for mercantile marine use is critically examined, the conclusion being that it will be most useful where very-short-range high-resolution pictures are required. Before such equipment is economically available further developments in transmitting tubes must take place."

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 3073

Observations on the Characteristics of the Cold-Cathode Ionization Gauge—J. H. Leck and A. Riddoch. (*Brit. J. Appl. Phys.*, vol. 7, pp. 153–155; April, 1956.) A gauge of the type described by Penning and Nienhuis (1423 of 1950) has been calibrated for the pressure range 10^{-6}–10^{-9} mm Hg. Anode-cathode-voltage/current and pressure/current characteristics are given; in the latter a sharp discontinuity occurs at a pressure of about 10^{-8} mm Hg. A marked change in sensitivity occurs during the first 200 hours of operation; this may account for conflicting characteristics obtained by various workers.

533.56 3074

The Ultimate Vacuum Obtainable in Vapour Pumps—N. A. Florescu. (*Vacuum*, vol. 4, pp. 30–39; January, 1954.) Experiments with hydrogen are described; the results indicate that in a well-designed vapor pump the ultimate vacuum is limited not by the pressure of the gas diffused from the fore-pressure side but by the lowest total pressure of all gases and vapors leaving the nozzle, apart from the partial pressure of the vapor of the working fluid.

533.56 3075

Theory of Molecular Pumps at Very Low Pressures—C. Mercier. (*J. Phys. Radium*, vol. 17, Supplement to No. 3, *Phys. Appl.*, pp. 1A–11A; March, 1956.)

535.215+535.37 3076

A Theoretical Property of Relaxation Curves of Luminescence and Photoconductivity—N. A. Tolstoi and A. V. Shatilov. (*Zh. Eksp. Teor. Fiz.*, vol. 30, pp. 109–114; January, 1956. English summary, *ibid.*, Supplement, p. 6.) A note on the recombination mechanism of phosphors and photoconductors.

535.215:546.817.221 3077

A Photo-E.M.F. Dependent on Direction of Illumination in Polycrystalline PbS Films—G. Schwabe. (*Ann. Phys., Lpz.*, vol. 17, pp. 249–262; February 29, 1956.) Fuller account of work described previously (3271 of 1955).

535.215:[546.863.221+546.23 3078

Time-Lag in Photoconductors for Camera Tubes—W. R. Daniels. (*J. IEE*, vol. 2, pp. 150–151; March, 1956.) A brief note on preliminary observations of the time lag in amorphous Se Sb_2S_3.

535.37:546.472.21 3079

Reduction of the Luminous Output of Phosphors under Intense Excitation—V. V. Antonov-Romanovski and L. A. Vinokurov. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 830–833; December, 1955.) Measurements on ZnS-Cu, Co, comparison with earlier measurements on ZnS-Cu, indicate that the observed effect is due to an increase of the concentration of localized electrons and ionized luminescence centers resulting in an increase of the number of radiationless recombinations.

535.37:546.472.21 3080

Phosphorescence of ZnS-Cu Crystal Phosphor excited by an Electron Beam—T. P. Belikova. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 905–906; December, 1955.) Luminescence decay curves of a ZnS-Cu specimen excited by radiation of wavelength 365 mμ and by an electron beam (2000 v, up to 3μA/cm²) are compared. The initial-intensity/temperature curves are also given.

535.376 3081

Electroluminescence from Boron Nitride—S. Larach and R. E. Shrader. (*Phys. Rev.*, vol. 102, p. 582; April 15, 1956.) A preliminary note reporting observations of electroluminescence with alternating-field excitation, using an electrode isolated from the phosphor.

537.226+537.228.1]:546.431.824-31 3082

Elastic, Piezoelectric, and Dielectric Constants of Polarized Barium Titanate Ceramics and some Applications of the Piezoelectric Equations—R. Bechmann. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 347–350; May, 1956.) A complete set of the constants and the various electromechanical coupling factors is given and typical values are tabulated.

537.228.1:549.514.51 3083

Piezoelectric Structure of Quartz and of Minerals Containing Quartz—E. I. Parkhomenko. (*Bull. Acad. Sci. U.R.S.S., Sér. Géophys.*, No. 3, pp. 297–306; March, 1956. In Russian.)

537.311.31:539.23 3084

The Electrical Conductivity of Anisotropic Thin Films—R. Englman and E. H. Sondheimer. (*Proc. phys. Soc.*, vol. 69, pp. 449–458; April 1, 1956.) "It is shown that, when the electron free path is large, the theoretical electrical conductivity of single crystal metal films exhibits anomalous anisotropic properties similar to, but even more pronounced than, those found in the anomalous skin effect in anisotropic metals."

537.311.31:621.316.842(083.74) 3085

Nickel-Chromium-Aluminium-Copper Resistance Wire—A. H. M. Arnold. (*Proc. IEE*, part B, vol. 103, pp. 439–447; July, 1956.) Report of an investigation made at the National Physical Laboratory on the suitability of alloys for resistance standards. The alloy "evanohm," composed of Ni, Cr, Al, and Cu, has a resistivity three times that of Mn, and its temperature coefficient can be adjusted to zero by heat treatment. A number of resistance standards made of this wire are undergoing long-term stability tests.

537.311.33 3086

Grain-Boundary Structure and Charge-Carrier Transport in Semiconductor Crystals—H. F. Mataré. (*Z. Naturf.*, vol. 10a, pp. 640–652; August, 1955.) "The structural character of boundaries or interfaces between two perfect crystals of different orientation but equal chemical composition defines the behavior of grain boundaries with respect to carrier transport. The amount of misfit in the grain boundary zone, as well as the amount of energy stored by elastic deformation, defines the electrical properties. The number of free carriers (electrons) in boundary states increases with the cross-potential applied, while positive space charge regions build up on both sides of the boundary. The boundary zone itself has p-type character and becomes more conductive when the number of electrons bound to the dangling bonds increases. Grain boundary zones may be as thick as a few tenths of a mm. Extremely small zones are formed by disturbed twins. Two- and three-probe measurements on such bicrystals have been made in order to study the carrier transport phenomena. High current multiplication due to carrier density misfit and gate action in the case of opposite polarization have been found. In addition, contacts were plated to boundary zones and modulation through the bulk material, as in a *n-p-n* junction, was studied. Here current multiplication can reach high values even in a base-to-ground connection. Since those electrons bound to a grain boundary interface by a cross potential may be present only in the form of excitons, in the field of their dangling bonds before adjustment, their time constants for recharging processes might be very short such that it is probable that high-frequency response is improved. Basic elements and consequences of the developed theory and the correlation between boundary stress field and carrier transport are outlined." Similar material is presented in a paper entitled "Grain Boundaries and Transistor Action" in 1955 IRE CONVENTION RECORD, vol. 3, part 3, pp. 113–124.

537.311.33 3087
p-n Junction Theory by the Method of δ Functions—H. Reiss. (*J. Appl. Phys.*, vol. 27, pp. 530–537; May, 1956.) A concise method is presented for calculating the current flow in one-dimensional semiconductor structures with any number of junctions and contacts. The method indicates the importance of the space derivatives of the hole currents in the neighborhood of junctions.

537.311.33 3088
A Method for Measurement of Surface-Recombination Velocities in Semiconductors using the Photomagnetoelectric Effect in a Sinusoidal Regime—J. Grosvalet. (*Ann. Radioélect.*, vol. 10, pp. 344–347; October, 1955.) The method is based on the phase difference between the photomagnetoelectric and photoresistive voltages discussed previously (1062 of 1955).

537.311.33:536.21 3089
Thermal Conductivity of Semiconductors—J. M. Thuillier. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2633–2634; May 28, 1956.) Addendum to analysis presented previously (799 of 1956). An error in the calculation is corrected.

537.311.33:536.21 3090
Thermal Conductivity of Semiconductors—A. V. Ioffe and A. F. Ioffe. (*Bull Acad. Sci. U.R.S.S., Sér. Phys.*, vol. 20, pp. 65–75; January, 1956. In Russian.) A discussion of theoretical and experimental work.

537.311.33:537.533 3091
The Effect of Field Emission on the Behaviour of Semiconductor Contacts—R. Stratton. (*Proc. Phys. Soc.*, vol. 69, pp. 491–492; April 1, 1956.) Recent work by Sillars (1084 of 1956) is extended to include field emission across gaps of arbitrary width and fields varying with the distance from the center of the contact, such as might arise if a variable work function exists at the gap surfaces.

537.311.33:[546.28+546.289 3092
Chemical Interactions among Defects in Germanium and Silicon—H. Reiss, C. S. Fuller, and F. J. Morin. (*Bell Syst. Tech. J.*, vol. 35, pp. 535–636; May, 1956.) Chemical reaction mechanisms in semiconductor solid solutions are shown to be similar to those in aqueous solutions. A comprehensive report of experimental and theoretical investigations of these mechanisms is presented. The limits of validity of the mass-action principle are examined. 71 references.

537.311.33:546.28 3093
Theory of Electron Multiplication in Silicon—J. Yamashita. (*Progr. Theor. Phys.*, vol. 15, pp. 95–110; February, 1956.) General theory of the conductivity of nonpolar crystals in strong es fields, developed previously [ibid., vol. 12, pp. 443–453; October, 1954. (Yamashita and Watanabe)] on a kinetic-statistical basis, is extended to take account of the impact ionization process and is used to explain the electron multiplication in Si *p-n* junctions observed by McKay and McAfee (1079 of 1954).

537.311.33:546.28 3094
Measurement of Minority Carrier Lifetime in Silicon—R. L. Watters and G. W. Ludwig. (*J. Appl. Phys.*, vol. 27, pp. 489–496; May, 1956.) A method of measurement based on the decay of photocurrent in a specimen exposed to pulsed illumination is used. Limitations on the injection level are discussed. Trapping, barrier, and contact effects are taken into account in evaluating the results, which are checked by measurements using a drift technique. Lifetime values >1500 μs for *p*-type crystals and >2500 μs for *n*-type have been found. The temperature dependence of the lifetime was investigated. A value of about 3500 cms at 300°K was determined for the surface recombination velocity of a *p*-type crystal.

537.311.33:546.28 3095
Diffusion of Donor and Acceptor Elements in Silicon—C. S. Fuller and J. A. Ditzenberger. (*J. Appl. Phys.*, vol. 27, pp. 544–553; May, 1956.) The diffusion of Group-III and Group-V elements in Si has been measured over the temperature range 1050°–1350°C. Results are tabulated. In nearly all cases the acceptor elements diffuse more rapidly than the donor elements. Boron and phosphorus exhibit similar diffusional properties; they may form compounds with the Si under the conditions of diffusion.

537.311.33:546.28:535.37 3096
Photon Emission from Avalanche Breakdown in Silicon—A. G. Chynoweth and K. G. McKay. (*Phys. Rev.*, vol. 102, pp. 369–376; April 15, 1956.) Results obtained by Newman (1088 of 1956) are discussed. Further experiments were made using a junction very close to a surface; the results indicate that light is emitted from breakdown regions distributed over the whole of the junction area, not only where the junction intercepts the surface. The emitted light has a continuous spectrum. Recombination between free electrons and holes is thought to be responsible for the shorter wavelengths, and intra-band transitions for the longer ones. The emission efficiency over the visible spectrum is tentatively estimated as 1 photon per 10^8 electrons crossing the junction.

537.311.33:546.289 3097
Effect of Water Vapor on Germanium Surface Potential—A. R. Hutson. (*Phys. Rev.*, vol. 102, pp. 381–385; April 15, 1956.) A simple calculation based on the thickness and dielectric properties of the water film adsorbed on the Ge surface gives values of the surface potential in good agreement with the observed values for different degrees of humidity of the ambient atmosphere.

537.311.33:546.289 3098
Temperature-Dependent Factor in Carrier Lifetime—R. L. Longini. (*Phys. Rev.*, vol. 102, pp. 584–585; April 15, 1956.) Results of measurements on carrier recombination in Ge made by various workers are discussed. It is suggested that rapid recombination believed to occur at dislocations may be due to relaxation of momentum selection rules. When recombination does take place predominantly at dislocations, the lifetime is not necessarily temperature dependent.

537.311.33:546.289 3099
Time-Dependent Changes of Surface Lifetime in Germanium in the Presence of Electrical Fields—J. D. Nixon and P. C. Banbury. (*Proc. Phys. Soc.*, vol. 69, pp. 487–488; April 1, 1956.) Extension of the work of Henisch and Reynolds (3652 of 1955); curves show the relation between surface-recombination velocity and applied field for both *n*- and *p*-type Ge, and the time variation of the conductance on applying and removing the field.

537.311.33:546.289 3100
The Absorption of 39-kMc/s (39-Gc/s) Radiation in Germanium—A. F. Gibson. (*Proc. phys. Soc.*, vol. 69, pp. 488–490; April 1, 1956.) Experimentally determined values of the absorption coefficient over the temperature range 15°—55°C are in excellent agreement with theory, assuming the effective mass of charge carriers to be of the same order as the electronic mass. The results are not in agreement with those of Klinger (1088 of 1954), which indicate an effective mass about ten times greater.

537.311.33:546.289:548.24 3101
Growth Twins in Germanium—G. F. Bolling, W. A. Tiller, and J. W. Rutter. (*Canad. J. Phys.*, vol. 34, pp. 234–240; March, 1956.) The nucleation of twin crystals in Ge requires a certain degree of supercooling; the frequency of occurrence of twins increases with the degree of supercooling. The addition of Ga to the melt lowers the solid/liquid interface energy.

537.311.33:546.289:669.046.54 3102
Single Crystals of Exceptional Perfection and Uniformity by Zone Leveling—D. C. Bennett and B. Sawyer. (*Bell Syst. Tech. J.*, vol. 35, pp. 637–660; May, 1956.) Technique for producing semiconductors with very low impurity content and with very uniform impurity distribution is based on traversing a single liquid zone through the crystal. Ge crystals have been produced with transverse variations of resistivity as low as ± 3 per cent and longitudinal variations ± 7 per cent.

537.311.33:546.3–1–28–289 3103
Preparation of Alloys of Germanium with Silicon and Other Metalloids by Fusion Electrolysis—M. J. Barbier-Andrieux. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2352–2354; May 7, 1956.) A whole range of mixed Ge-Si crystals has been obtained by the technique described. Some experiments with Ge-Sn and Ge-As alloys are also mentioned.

537.311.33:546.561–31 3104
Excitation Spectrum of Excitons in a Solid—E. F. Gross. (*Bull. Acad. Sci. U R.S.S., Sér. Phys.*, vol. 20, pp. 89–104; January, 1956. In Russian.) A critical survey of literature with particular reference to Cu_2O. 45 references.

537.311.33:546.561–31 3105
Occlusions of Cupric Oxide in Cuprous Oxide Layers—A. I. Andrievski and M. T. Mishchenko. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1893–1897; October, 1955.) Statements made by various authors to the effect that layers of Cu_2O contain crystals of CuO have been confirmed by a microscope investigation. A report is presented including a number of photomicrographs.

537.311.33:546.561–31:539.23 3106
Investigation of the Structure of the Surface of Films of Cuprous Oxide on Different Faces of a Single Crystal of Copper and Determination of the Contact Potential Difference between these Surfaces—N. B. Gornyi. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 808–816; December, 1955.)

537.311.33:[546.682.18+546.681.19 3107
Preparation and Electrical Properties of InP and GaAs—O. G. Folberth and H. Weiss. (*Z. Naturf.*, vol. 10a, pp. 615–619; August, 1955.) Measurements were made of conductivity and Hall effect over the temperature range from $-180°$ to $+960$°C. Polycrystalline rod specimens were used. Results are shown

graphically. Values are deduced for the carrier mobilities and the widths of the energy gaps.

537.311.33:546.682.86 3108
Preparation of Indium Antimonide. Determination of the Effective Masses—M. Rodot, P. Duclos, F. Kover, and H. Rodot. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2522–2525; May 23, 1956.) Specimens of various degrees of purity were prepared; impurity concentrations down to 10^{15} centers/cm^3 were attained. Hall-effect and Seebeck-effect measurements indicate that the effective masses of electrons and holes depend greatly on temperature.

537.311.33:546.786–31 3109
The Preparation of Semiconducting Ceramics based on WO_3, and a Study of Some of their Electrical and Thermal Properties—G. I. Skanavi and A. M. Kashtanova. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1883–1892; October, 1955.) The preparation of the specimens is described in detail and results are given of numerous experiments. The main properties of the material are as follows: the conductivity varies within relatively wide limits, from 7×10^{-3} to 4 Ω^{-1}. cm^{-1}; the thermo-emf has negative sign, corresponding to n-type conductivity; the temperature coefficient of thermo-emf is relatively high (0.70–0.85 mv/deg). The material should find application in the production of thermocouples.

537.311.33:546.873–31 3110
The Electrical Conductivity of Bismuth Oxide—V. M. Konovalov, V. I. Kulakov, and A. K. Fidrya. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1864–1867; October, 1955.) Measurements are reported. In air, at room temperature, the resistivity varied from 10^8 to 10^{10} Ω. cm. When the specimens were heated up to 700°C, the resistivity fell to about 10^2Ω. cm. The conductivity depends to a great extent on the preparation of the specimens and on their moisture content. The results indicate that within the range of temperatures investigated the conductivity is predominantly of n-type, which is contrary to previous conclusions. A considerable positive photoeffect was also observed.

537.32:546.562–31 3111
A Thermoelectric Effect Exhibited by Cupric Oxide in Powder Form—M. Perrot, G. Peri, J. Robert, J. Tortosa, and A. Sauze. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2519–2522; May 23, 1956.) Experiments have been made with elements comprising powdered CuO compressed between two Cu electrodes. Graphs show the temperature variation of resistance of an element as a whole, and the variation of the thermo-emf as a function of the temperature difference between the electrodes for several elements; in one case the useful power is 22 mW/cm^2. Elements using Cu_2O powder give a greater emf for the same temperature conditions, but their resistance is also greater.

537.533:546.815 3112
Work Function of Lead—P. A. Anderson and A. L. Hunt. (*Phys. Rev.*, vol. 102, pp. 367–368; April 15, 1956.) The work function of Pb surfaces has been determined by measuring the contact difference of potential with respect to a Ba surface in a special tube. The value obtained is 4.00±0.01 ev. The results indicate that the work function is unaffected when an initially clean Pb surface is exposed to the residual gas in a sealed-off Ba-gettered tube.

538.22:621.318.134 3113
Micrographic Study of the Order-Disorder Transformation in Lithium Ferrite—I. Behar. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2465–2468; May 14, 1956.)

538.22:621.318.134 3114
Magnetic Properties of Garnet-Type Yttrium Ferrite $5Fe_2O_2 \cdot 3Y_2O_3$—R. Aléonard, J. C. Barbier, and R. Pauthenet. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2531–2533; May 23, 1956.)

538.221 3115
The Behavior of Ferromagnetics under Strong Compression—F. D. Stacey. (*Canad. J. Phys.*, vol. 34, pp. 304–311; March, 1956.) Magnetization curves are given for thin specimens of Ni and Ni-Cu alloys under nonhydrostatic pressures up to 10,000 atm. The saturation magnetizations increase markedly with pressure.

538.221 3116
Interpretation of Domain Patterns recently found in BiMn and SiFe Alloys—J. B. Goodenough. (*Phys. Rev.*, vol. 102, pp. 356–365; April 15, 1956.)

538.221:538.632 3117
Theory of the Hall Effect in Ferromagnetic Alloys—K. Meyer. (*Z. Naturf.*, vol. 10a, pp. 656–657; August, 1955.)

538.221:538.652 3118
Iron-Aluminum Alloys for Use in Magnetostrictive Transducers—M. T. Pigott. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 343–346; May, 1956.) A systematic determination of the electromechanical coupling coefficient k of Fe-Al alloys containing between 12 per cent and 14 per cent Al by weight and annealed at temperatures between 600° and 1000°C. is reported. For annealing temperatures near 1000°C., k^2 is about 0.05 and is nearly independent of composition; k^2 has a maximum value of 0.12 for an alloy containing 12.3 per cent Al annealed at 650°C. Eddy-current losses are smaller than for soft annealed "A" nickel.

538.221:621.318.134 3119
Resonance Widths in Polycrystalline Nickel-Cobalt Ferrites—M. H. Sirvetz and J. H. Saunders. (*Phys. Rev.*, vol. 102, pp. 366–367; April 15, 1956.) Brief report of measurements at a frequency of 10 kmc on ferrites of composition $Co_\alpha Ni_{1-\alpha}Fe_2O_4$. The variation of resonance-line width with variation of α up to 0.04 and with variation of temperature between 20° and 350°C is shown graphically and discussed in relation to the crystal properties.

538.221:621.318.134 3120
Investigation of the Magnetic Spectra of Solid Solutions of some NiZn Ferrites at Radio Frequencies—L. A. Fomenko. (*Zh. Eksp. Teor. Fiz.*, vol. 30, pp. 18–29; January, 1956. English summary, *ibid.*, Supplement, p. 3.) Results of an experimental investigation of the frequency dependence in the range 0.2–60 mc of the permittivity, permeability, and loss angles of oxifer ferrites (Shol'ts and Piskarev, *Bull. Acad. Sci. U.R.S.S., Sér. Phys.*, vol. 16, p. 6; 1952) with initial permeabilities of 200, 400, and 2000 G/oersted are presented graphically. Specimens with various dimensions were used; dispersion effects are practically independent of the dimensions.

538.221:621.318.134 3121
Influence of Alkali and Alkaline-Earth Ions on the Initial Permeability of Manganese-Zinc Ferrites—C. Guillaud, B. Zega, and G. Villers. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2312–2315; May 7, 1956.) Results of measurements are presented as curves for $\mu0/\mu$ as a function of impurity content, where $\mu0$ is the initial permeability of the pure material and μ that of the impure material. The relation between the effectiveness of the impurity and its ionic radius is studied.

538.221:621.318.134 3122
Initial Permeability and Grain Size of Manganese-Zinc Ferrites—C. Guillaud and M. Paulus. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2525–2528; May 23, 1956.) A graph shows the relation between initial permeability and mean grain size, derived on the basis of a careful analysis of the distribution of grain size in 100 specimens. The results are consistent with a mechanism involving rotation of the direction of spontaneous magnetization for grains whose mean dimension is <5.5 μ, and domain-wall displacements for larger grains.

538.23 3123
A Relation between Hysteresis Coefficient and Permeability: Part 3—Ferrite Cores with Rectangular Loop. Part 4—Influence of Coercive Force—M. Kornetzki. (*Z. Angew. Phys.*, vol. 8, pp. 127–135; March, 1956.) Continuation of the investigation noted earlier (1714 of 1955). Anomalies due to the large magnetic crystal energy of several materials are noted and experimental results obtained by various workers are discussed.

539.23:537/538 3124
International Colloquium on the Present State of Knowledge of the Electric and Magnetic Properties of Thin Metal Films in Relation to their Structure—(*J. Phys. Radium*, vol. 17, pp. 169–306; March, 1956.) French text and English abstracts are presented of 27 papers given at a colloquium held at Algiers in April, 1955.

539.23:546.561–31 3125
Electron Interference at Electrolytically Polished Surfaces after Cathode Sputtering—A. Ladage. (*Z. Phys.*, vol. 144, pp. 354–372; February 7, 1956.) Apparatus is described by means of which thin Cu_2O films were detected on the surface of cleaned Cu exposed to air for 30 minutes.

549.514.5:534.21–16 3126
Propagation of Longitudinal Waves and Shear Waves in Cylindrical Rods at High Frequencies—Mc Skimin. (See 2933.)

621.3.049.75 3127
Silver Migration in Electric Circuits—O. A. Short. (*Tele-Tech and Electronic Ind.*, vol. 15, pp. 64–65, 113; February, 1956.) Electrolytic migration of silver used in components and printed circuits may be reduced by covering the silver completely with solder, or by Cr plating. An organic coating is effective if soluble salts are first removed from the surface covered.

621.315.61:621.317.335.029.64 3128
Temperature Dependence of Loss Angle and Dielectric Constant of Solid Insulating Materials in the 4-kMc/s Range—Gross. (See 3139.)

621.315.612.6 3129
Electrical Resistivity of Vitreous Ternary Lithium-Sodium Silicates—S. W. Strauss. (*J. Res. Nat. Bur. Stand.*, vol. 56, pp.1 83–185; April, 1956.) Glasses with compositions in the system xLi$_2$O:(1−x) Na$_2$O:2SiO$_2$ have been investigated over the temperature range 150°–230°C. Resistivity/composition characteristics are presented.

621.315.615.9:621.319.4 3130
Polychloronaphthalene—Impregnated—Paper Capacitors—Coquillion. (See 2980.).

MATHEMATICS

517.9 3131
The Asymptotic Solution of Linear Differential Equations of the Second Order in a Domain containing One Transition Point—F. W. J. Olver. (*Phil. Trans. A*, vol. 249, pp. 65–97; April 19, 1956.)

517.941.91 3132
The Interrelation between the Phase Planes of Rayleigh's Equation and van der Pol's Equation—V. V. Kazakevich. (*Compt. Rend. Acad. Sci., U.R.S.S.*, vol. 107, pp. 521–523; April 1, 1956. In Russian.) The equations

considered are: $\ddot{y}-\mu f(\dot{y})+y=0$ and $\ddot{y}-\mu F(y)\dot{y}+y=0$.

517 3133

Spheroidal Wave Functions [Book Review] —J. A. Stratton, *et al.* Publishers: Technology Press of Massachusetts Institute of Technology, and John Wiley and Sons, New York, 611 pp.; 1956. (PROC. IRE, vol. 44, pp. 951–952; July, 1956.) Contains numerical tables and an introduction, together with a reprint of a paper on elliptic and spheroidal wave functions [1594 of 1942 (Chu and Stratton)].

MEASUREMENTS AND TEST GEAR

621.3.011.3(083.74):621.318.42 3134

The Calibration of Inductance Standards at Radio Frequencies—L. Hartshorn and J. J. Denton. (*Proc. IEE*, part B, vol. 103, pp. 429–438; July, 1956. Discussion, p. 438.) The practice adopted at the National Physical Laboratory for calibrating laboratory standards is described. An accuracy within about 1 part in 10^4 is obtained over a considerable range of inductance values. The accuracy associated with such standards is determined partly by the definition of inductance used; this aspect as well as the experimental technique is discussed.

621.317.3:551.594.6 3135

Measurement of the Amplitude Probability Distribution of Atmospheric Noise—H. Yuhara, T. Ishida, and M. Higashimura. (*J. Radio Res. Labs., Japan*, vol. 3, pp. 101–108; January, 1956.) Noise picked up on a 2-m vertical antenna is amplified at an IF of 100 kc, the output is sliced and the resulting groups of 100-kc pulses are counted. Results obtained during the summer of 1955, on a frequency of 3.5 mc, using a bandwidth of 2.4 kc, show that the noise includes random and impulsive components.

621.317.3:621.319.4:681.142 3136

Industrial Measurement of the Temperature Coefficient of Ceramic-Dielectric Capacitors—J. Peyssou and J. Ladefroux. (*Ann. Radioélect.*, vol. 10, pp. 355–371; October, 1955.) Known beat-frequency and self-synchronizing techniques are reviewed. The accuracy and speed of measurements is increased by using an automatic machine incorporating an analog computer. The construction of a temperature-coefficient distribution curve for a batch of 4000 capacitors is described. For a shorter version, in English, see *Tele-Tech and Electronic Ind.*, vol. 15, pp. 70–71, 166; April, 1956.

621.317.3:621.396.822 3137

New Method of measuring the Effective Value of Band-Limited Radio Noise Voltage—K. Kawakami and H. Aikma. (*J. Radio Res. Labs., Japan*, vol. 3, pp. 109–113; January, 1956.) The noise voltage is passed through a pentode frequency-doubling stage and the output is linearly rectified and smoothed. The resulting voltage is the mean square of the input voltage. Equipment is described for measurements on a center frequency of 455 kc, giving accurate results for an input dynamic range of 30 db.

621.317.33:546.28 3138

The Measurement of the Electrical Resistivity of Silicon—R. H. Creamer. (*Brit. J. Appl. Phys.*, vol. 7, pp. 149–150; April, 1956.) The method described by Valdes (1502 of 1954) was modified by using probes made from wires containing a donor or acceptor impurity for measurements on *n* or *p*-type Si respectively. Potentials were measured with a standard potentiometer, giving an accuracy within ±7 per cent for resistivities up to several hundred Ω. cm.

621.317.335.029.64:621.315.61 3139

Temperature Dependence of Loss Angle and Dielectric Constant of Solid Insulating Materials in the 4-kMc/s Range—F. Gross. (*Nachrichtentech. Z.*, vol. 9, pp. 124–128; March, 1956.) Measurements were made on rod specimens of ceramics, glass, and plastics used in the manufacture of tubes and other equipment, over the temperature range 20°–350°C, using an E_{010}-mode resonator. Theory based on that of Horner, *et al.* (966 of 1964) is outlined; results are presented in tables and graphs.

621.317.335.3.029.64:621.315.614.6 3140

Birefringence and Rectilinear Dichroism of Paper at 9350 Mc/s—R. Servant and J. Gougeon. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2318–2320; May 7, 1956.) The complex dielectric constant of a pile of sheets of paper has been determined by a waveguide method using a swr meter within which the material under test is located. Measurement results are evaluated as absorption coefficients and refractive indices; very considerable differences are observed for the cases of the electric vector a) parallel to and b) perpendicular to the plane of the paper sheets. Some results obtained with kraft paper are shown graphically.

621.317.337:621.372.413 3141

Measurement of the *Q*-Factor of Cavity Resonators, using a Straight Test Line—H. Urbarz. (*Nachrichtentech. Z.*, vol. 9, pp. 112–118; March, 1956.) Methods appropriate for measurements on resonators with only one coupling point, such as those associated with klystrons, are based on determination of the swr and the shift of the minimum along a test line terminated by the resonator. The effect of loading on the *Q*-factor is discussed. Measurements are reported indicating the variation of the resonator input admittance with the area of the coupling loop.

621.317.34:621.3.018.7 3142

An Approximate Method for Investigating Distortion of Test Pulses Transmitted over Coaxial Cables—H. Larsen and H. E. Martin. (*Frequenz*, vol. 10, pp. 65–76; March, 1956.) In practice, the waveforms of pulses used for testing may deviate considerably from ideal forms such as rectangular or cos². The Fourier components of the actual initial waveform can be determined with sufficient accuracy by analyzing its oscillogram. The waveform of the transmitted pulse can then be determined as usual by multiplying together the pulse spectral function and the system transfer function and transforming the product. Application of the theory is described in relation to tests on wide-band cables several km long.

621.317.4 3143

A Rapid Method for Measuring Coercive Force and other Ferromagnetic Properties of Very Small Samples—G. W. van Oosterhout. (*Appl. Sci. Res.*, vol. B6, pp. 101–104; 1956.) The method is based on measurement of the alternating emf generated when the sample is caused to vibrate within a search coil.

621.317.443 3144

Description of a Balance for the Measurement of Magnetization from 1.4°K to Room Temperature—R. Conte. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2528–2531; May 23, 1956.)

621.317.6:621.385.5:621.376.22 3145

Study of Amplitude Modulation applied via a Pentode Suppressor Grid—Loeckx. (See 3237.)

621.317.7:537.54:621.396.822.029.6 3146

On the Effective Noise Temperature of Gas-Discharge Noise Generators—W. D. White and J. G. Greene. (PROC. IRE, vol. 44, p. 939; July, 1956.) A method of calculating the noise temperature is indicated.

621.317.7:537.54:621.396.822.029.6 3147

Wide-Band Noise Sources using Cylindrical Gas-Discharge Tubes in Two-Conductor Lines—R. I. Skinner. (*Proc. IEE*, part B, vol. 103, pp. 491–496; July, 1956.) Noise sources for the dm-λ band are discussed. A noise output which is level over several octaves can be obtained by matching a cylindrical gas-discharge tube directly to a two-conductor line. The matching can be achieved by using conductors of various shapes. Practical design procedure is outlined.

621.317.72+621.317.772 3148

An A.C. Potentiometer for Measurement of Amplitude and Phase—M. J. Somerville. (*Electronic Engng.*, vol. 28, pp. 308–309; July, 1956.) A simple circuit using ac coupled amplifiers permits generation of quadrature components whose phase relation remains unchanged when substantial phase shifts occur in the couplings.

621.317.73+534.64 3149

An Impedance Measuring Set for Electrical, Acoustical and Mechanical Impedances—E. W. Ayers, E. Aspinall, and J. Y. Morton. (*Acustica*, vol. 6, pp. 11–16; 1956.) "An impedance to be measured is compared with a reference impedance of similar nature by connecting each in turn to a source of adjustable strength. If the internal impedance of the source is constant, the vector ratio of the unknown and reference is the ratio of the changes in stimulus required to restore the source to short-circuit conditions, or the reciprocal of this ratio if the source is restored to open-circuit conditions."

621.317.733.029.4:621.375.2 3150

A Tuned Differential Amplifier for Low-Frequency Bridges—W. K. Clothier and F. C. Hawes. (*Aust. J. Appl. Sci.*, vol. 7, pp. 38–44; March, 1956.) The amplifier described is suitable for use as a balance detector where there is high impedance between both detector points and ground. Rejection factors greater than 30,000 are obtained for in-phase input voltages up to 10 v. The amplifier is tunable over the frequency range 15–20,000 cps by means of ladder-type feedback networks. The discrimination against third harmonics of the selected frequency is 130. Maximum gain is 150,000.

621.317.734 3151

Extending the Limits of Resistance Measurement using Electronic Techniques—G. Hitchcox. (*J. Brit. IRE*, vol. 16, pp. 299–309; June, 1956.) Methods for measuring resistance are surveyed with special attention to those for very low and very high resistance. In one method for dealing with very low resistance, test currents with triangular waveform are used to reduce thermal dissipation. A commercial general-purpose megohmmeter is described in some detail.

621.317.734 3152

A Logarithmic Megohmmeter—P. Hariharan and M. S. Bhalla. (*J. Sci. Instrum.*, vol. 33, pp. 158–159; April, 1956.) An ohmmeter based on the logarithmic grid-current /anode-current characteristic of a triode tube covers the range from 1 to 10^6 MΩ on a single approximately logarithmic scale.

621.317.75:621.396.3 3153

The Response of Radio Spectrometers—J. Marique. (*Rev. MF, Brussels*, vol. 3, pp. 167–177; 1956.) The spectrum of repeated signals such as the pulses in on-off telegraphy systems is a function of two factors, one depending on the waveform of the individual signals and the other on the repetition process. The operation of a spectrometer comprising a cascaded-tuned-circuit filter (813 of 1955) is discussed, taking as criterion the time interval $T=2/B_F$, where B_F is the filter bandwidth. See also 1900 of 1955.

621.317.755:531.76 3154

Four-Place Timer Codes Oscillograph Recordings—S. E. Dorsey. (*Electronics*, vol. 29, pp. 154–156; July, 1956.) A 1-kc signal from a tuning-fork oscillator is fed through a trigger circuit into a chain of four decade counters which have additional "staircase" outputs. Differentiation and combination of these outputs provides a cro trace indicating time in increments of 0.001 second up to 9.999 seconds with markers for tenths, hundredths, and thousandths of a second. A simple calibration method is described.

621.317.79:538.632:537.311.33 3155

A Simple Apparatus for recording the Variation of Hall Coefficient with Temperature—E. H. Putley. (*J. Sci. Instrum.*, vol. 33, p. 164; April, 1956.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

536.52:621.385.029.6.032.21 3156

A New Method for the Measurement of Rapid Fluctuations of Temperature—Dehn. (See 3258.)

550.837 3157

Geophysical Prospection of Underground Water in the Desert by means of Electromagnetic Interference Fringes—G. L. Brown: M. A. H. El-Said. (Proc. IRE, vol. 44, p. 940; July, 1956.) Comment on 1171 of 1956 and author's reply.

620.179.1:621–52 3158

An Electronic Position-Tracking Instrument—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 40, pp. 68–69; May, 1956.) The motion of a metal object in a nonconducting medium is automatically followed by a mutual-inductance probe associated with a servomechanism.

621.317.39:531.71 3159

Mechanic-Electric Transducer—K. S. Lion. (*Rev. Sci. Instrum.*, vol. 27, pp. 222–225; April, 1956.) A system for converting mechanical displacement into a voltage is based on the local variations of the voltage between a pair of electrodes in a luminous low-pressure discharge excited by a rf field.

621.317.39:621.383 3160

A Wide-Range Photoelectric Automatic Gain Control—C. Riddle. (*Electronic Engng.*, vol. 28, pp. 288–292; July, 1956.) "A photocell and tube are arranged in such a way that the output voltage is proportional to the light modulation, and independent of the value of the steady light flux. The circuit is extremely simple, and the range over which the light flux may vary is very large (100,000:1)."

621.383:77:522.61 3161

Obtaining the Spectra of Faint Stars by Electronic Photography—A. Lallemand and M. Duchesne. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 2624–2626; May 28, 1956.)

621.385.5:531.745:621.396.934 3162

Photoelectric Angular Error-Sensors—R. A. Nidey and D. S. Stacey. (*Rev. Sci. Instrum.*, vol. 27, pp. 216–218; April, 1956.) A device is described in which Ge-junction photocells are used to produce a voltage dependent on the orientation of a research rocket relative to the sun. See also 3182 below.

621.384.611 3163

Improving the Characteristics of the Cyclotron Beam—W. B. Powell. (*Nature, London.*, vol. 177, p. 1045; June 2, 1956.) Brief preliminary note of a technique involving the use of beam-defining slits on the dee interface.

621.384.612 3164

Excitation of Synchrotron Oscillations due to Electron Radiation Fluctuations in a Strong-Focusing Accelerator—A. A. Kolomenski. (*Zh. Eksp. Teor. Fiz.*, vol. 30, pp. 207–209; January, 1956.) Theoretical note. If $H_{max} \cong 10^4$ oersted and $E \pm 10$ kmev, then the radial rms deviation of the orbit is of the order of a fraction of a centimeter.

621.384.612 3165

Influence of Radiation on Betatron Oscillations of Electrons in Synchrotrons with Strong [alternating gradient] Focusing—A. N. Matveev. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 107, pp. 671–674; April 11, 1956. In Russian.)

621.384.612:681.142 3166

Analog Computer for the Differential Equation $y''+f(x)y+g(x)=0$—Bodenstedt. (See 2973.)

621.385.833 3167

Electrostatic Fields Permitting Rigorous Calculation of the Electron Paths—H. Grümm. (*Ann. Phys., Lpz.*, vol. 17, pp. 269–280; February 29, 1956.) Analysis is given separately for two-dimensional fields (pp. 269–274; and for rotationally symmetrical fields (pp. 275–280).

621.385.833 3168

Calculation of Electrostatic [electron] Lenses—U. Timm. (*Z. Naturf.*, vol. 10a, pp. 593–602; August, 1955.) The use of matrix methods is described and illustrated.

621.385.833 3169

Construction of Magnetic Electron Lenses—P. Durandeau. (*J. Phys. Radium*, vol. 17, Supplement to No. 3, *Phys. Appl.*, pp. 18A–25A; March, 1956.) Design of short-focus lenses for very-high-velocity electrons is based on measurements of the field along the axis by the method described previously (1743 of 1953).

621.385.833 3170

Stereoscopic Reflection Electron Microscopy—D. E. Bradley, J. S. Halliday, and W. Hirst. (*Proc. phys. Soc.*, vol. 69, pp. 484–485; April 1, 1956. plate.) The technique is briefly described, with some practical examples.

621.385.833 3171

Aperture Aberration of 5th Order in Spherically Corrected Electron Microscopes—W. E. Meyer. (*Optik, Stuttgart*, vol. 13, pp. 86–91; 1956.)

621.385.833 3172

The Lower Limit of Aperture Aberration in Magnetic Electron Lenses—H. Grümm. (*Optik, Stuttgart*, vol. 13, pp. 92–93; 1956.)

621.385.833:621.383.2 3173

Area Sources of Low-Energy Electrons for Electron-Optic Studies—R. J. Schneeberger (*Rev. Sci. Instrum.*, vol. 27, pp. 212–215; April, 1956.) If the final stages of the design of electron-optical systems for image tubes are carried out with a demountable tube containing a photocathode, the latter requires repeated cleaning and reprocessing. Three sources suitable as substitutes for the photocathode are discussed, *viz.*, a) a thermionic source which sprays electrons through a perforated large-area electrode at about cathode potential, b) a secondary-emission arrangement using a perforated plate with baffles associated with individual holes, and c) a secondary-emission transmission arrangement.

621.386:621.383.2 3174

Cineroentgenography with Image Intensification—F. J. Euler and P. A. Virbal. (*Elect. Engng., N.Y.*, vol. 75, pp. 238–242; March, 1956.) Intensification of the X-ray image by means of a special form of image-intensifier tube permits shortening of exposure time and increase in thickness of material examined in studies of objects in notion.

621.387.4:621.314.7 3175

The Application of Transistors to the Trigger, Ratemeter and Power-Supply Circuits of Radiation Monitors—E. Franklin and J. B. James. (*Proc. IEE*, Part B, vol. 103, pp. 497–504; July, 1956. Discussion, pp. 516–518.) General requirements and conditions of use of radiation monitors for γ and β-ray survey in connection with geological prospecting are outlined. Discussion indicates that junction transistors are preferable to either filament or cold-cathode tubes or point-contact transistors for these applications.

621.389 3176

An Electronic Machine for Statistical Particle Analysis—H. N. Coates. (*Proc. IEE*, Part B, vol. 103, pp. 479–484; July, 1956.) "A system is described for associating and collecting the intercepts of individual particles in a particle scanning system, where the information is presented as a function of the scanning voltages. A series of stores is used to segregate the intercepts, each store having its own memory system and provision for re-use on completion of the scanning of the particle with which it is associated; the stores can thus be used many times during a single frame scan. A method of adding the intercepts of each particle to obtain measure of the area of the particle is described, but this must be regarded as only one of the possibilities of extracting information from the series of intercepts collected."

621.396.934 3177

Missile Guidance by Three-Dimensional Proportional Navigation—F. P. Adler. (*J. Appl. Phys.*, vol. 27, pp. 500–507; May, 1956.)

621.398:621.376 3178

Telemetering Demodulator for Wide-Band F.M. Data—T. D. Warzecha. (*Electronics*, vol. 29, pp. 157–159; July, 1956.) Demodulation of 12 subcarrier signals is effected by a pulse-averaging technique after recording the signals at a reduced tape speed and converting fm to pfm.

621.398:621.396.93 3179

Remote Radio Control of a Train—(*Elect. J.*, vol. 156, pp. 998–999; March, 30, 1956. Brief account of a system which has been successfully operated in the U.S.A.

621.398:621.396.934 3180

Shipboard Telemetering for Terrier Missiles—W. S. Bell and C. W. Schultz. (*Electronics*, vol. 29, pp. 134–137; June, 1956.) Description of equipment for a six-channel fm/fm system providing magnetic-tape recordings of missile data.

621.398:621.396.934 3181

Transistor Modulator for Airborne Recording—J. L. Upham, Jr., and A. I. Dranetz. (*Electronics*, vol. 29, pp. 166–169; June, 1956.) Description of a ppm telemetry system for indicating pressure or acceleration, based on the displacement of the core of a differential transformer.

621.398:621.396.934 3182

Transistors Telemeter Small Missiles—C. M. Kortman. (*Electronics*, vol. 29, pp. 145–147; July, 1956.) Rate of spin of a missile 2 inches in diameter is determined from the cyclic frequency shift produced by the rotation of a Ge photocell exposed to the ambient light and connected across the coil of a junction-transistor Hartley oscillator. Curves showing oscillator frequency plotted against light intensity, temperature, etc. are given.

621.396.934 3183

Guidance [Book Review]—A. S. Locke and collaborators. Publishers: Van Nostrand, Princeton, N. J., and Macmillan, London, 1955, 729 pp. (*Nature, London*, vol. 177, pp. 1003–1004; June 2, 1956.) A general introduction and reference book, constituting the first of a projected series of five books on the principles of guided-missile design. The subjects involved include servomechanism theory, aerodynamics, radar, navigation, communications, and the application of computers.

PROPAGATION OF WAVES

538.566.029.43:551.594.6 3184

Influence of the Horizontal Geomagnetic Field on Electric Waves between the Earth and the Ionosphere Travelling Obliquely to the Meridian—Schumann. (See 3060.)

621.396 3185

Symposium on Communications by Scatter Techniques—(IRE Trans., vol. CS-4, pp. 1–122; March, 1956.) The text is given of papers presented at a symposium held in Washington in November, 1955. These include the following:

Some Practical Aspects of Auroral Propagation—H. G. Booker (p. 5).

Progress of Tropospheric Propagation Research related to Communications beyond the Horizon—J. H. Chisholm (pp. 6–16).

Practical Considerations for Forward Scatter Applications—J. R. McNitt (pp. 28–31).

Some Meteorological Effects on Scattered V.H.F. Radio Waves—B. R. Bean (pp. 32–38).

Point-to-Point Radio Relaying via the Scatter Mode of Tropospheric Propagation—K. A. Norton (pp. 39–49).

A Simplified Diversity Communication System for Beyond-the-Horizon Links—F. J. Altman and W. Sichak (pp. 50–55).

VHF Trans-horizon Communication System Design—R. M. Ringoen (pp. 77–86).

System Parameters using Tropospheric Scatter Propagation—H. H. Beverage, E. A. Laport, and L. C. Simpson (pp. 87–96).

A Simple Picture of Tropospheric Radio Scattering—W. E. Gordon (pp. 97–101).

Some Ionosphere Scatter Techniques—D. A. Hedlund, L. C. Edwards and W. A. Whitcraft, Jr. (pp. 112–117).

Signal Fluctuations in Long-Range Overwater Propagation—W. S. Ament and M. Katzin (pp. 118–122).

Abstracts of some of these are given in Proc. IRE, vol. 44, p. 831; June, 1956.

621.396.11 3186

Field Strength in the Vicinity of the Line of Sight in Diffraction by a Spherical Surface—K. Furutsu. (*J. Radio Res. Labs., Japan*, vol. 3, pp. 55–76; January, 1956.) The convergency of the formula for diffraction by a spherical earth is improved by using the expression for a flat earth, with an appropriate correction in the form of an integral.

621.396.11:551.510.535 3187

Observations of Ionospheric Absorption at the K.N.M.I. [Royal Netherlands Meteorological Institute]—van Daatselaar. (See 3054.)

621.396.11:551.510.535 3188

On the Existence of a "Q.L."—"Q.T." "Transition-Level" in the Ionosphere and its Experimental Evidence and Effect—D. Lepechinsky. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 297–304; June, 1956.) See 1767 of 1955 (Lepechinsky and Durand).

621.396.11:551.594.6 3189

The Propagation of a Radio Atmospheric—C. M. Srivastava. (*Proc. IEE*, Part B, vol 103, pp. 542–546; July, 1956.) Analysis is presented assuming that the original disturbance is a rectangular pulse of duration 100 μs and that propagation takes place by multiple reflections in the waveguide constituted by the earth and the ionosphere. The theory provides an explanation of the smooth oscillating waveform of atmospherics received from a distance.

621.396.11:621.396.674.3 3190

Radiation from a Vertical Antenna over a Curved Stratified Ground—J. R. Wait. (*J. Res. Nat. Bur. Stand.*, vol. 56, pp. 237–244; April, 1956.) Analysis is presented on the basis of a specified surface impedance at the earth's surface.

621.396.11.001.57 3191

Multipath Simulator Tests Communications—A. F. Deuth, H. C. Ressler, J. W. Smith, and G. M. Stamps—(*Electronics*, vol. 29, pp. 171–173; July, 1956.) A system designed for laboratory testing of long-range communication equipment is described. Two signal paths are provided by acoustic transducers operating at 150 kc in air which is disturbed by heat or fans to effect frequency-selective random fading.

621.396.11.029.4:551.594.6 3192

Propagation of Audio-Frequency Radio Waves to Great Distances—F. W. Chapman and R. C. V. Macario. (*Nature, London*, vol. 177, pp. 930–933; May 19, 1956.) Observations of atmospherics waveforms have been supplemented by simultaneously recording the relative amplitudes of the frequency components in the waveform spectrum. Magnetic recording techniques were used to obtain permanent records of all disturbances reaching a vertical rod antenna. A second channel on the magnetic tape provided information as to the source of individual disturbances. The spectrometer was a modified form of that used previously [419 of 1954 (Chapman and Matthews)]. The results described were obtained from observations of cloud-to-ground discharges at known distances up to about 4000 km. In all cases marked absorption was found at frequencies around 1–2 kc. An attenuation /frequency curve is presented linking the results with those obtained by Eckersley (*J. IEE*, vol. 71, pp. 405–454; September, 1932.) on long-distance radio transmissions at frequencies up to about 30 mc. For a range of frequencies below 200 or 300 cps the attenuation is no greater than for short waves.

621.396.11.029.45 3193

Long-Distance Propagation of 16-kc/s Waves—N. M. Rust. (*Marconi Rev.*, vol. 19, pp. 47–52; 1st Quarter, 1956.) Discussion of papers by Budden (2772 of 1953) and Pierce (2404 of 1955) suggests that the experimental results can be explained qualitatively in terms of simple ionosphere/ground-reflection propagation, taking into account up to four hops, without invoking more elaborate theories. The need for further experimental work is emphasized.

621.396.11.029.51 3194

Change of Phase with Distance of a Low-Frequency Ground Wave propagated across a Coast-Line—B. G. Pressey, G. E. Ashwell, and C. S. Fowler. (*Proc. IEE*, Part B, vol. 103, pp. 527–534; July, 1956.) Continuation of work described previously (1782 of 1953). Observations were made on a frequency of 127.5 kc along a number of paths of lengths up to 22 km radiating from a transmitter near Lewes, England, and crossing the coast between Pevensey and Littlehampton; some paths tangential to the coast-line and some at right angles to the radials were also studied. The results confirm the existence of the phase-recovery effect on passing from low-conductivity ground to sea water. They also indicate systematic phase variations whose magnitudes decay from about 4° near the coast to a negligible amount at 6λ out to sea. A very marked phase disturbance within $\lambda/2$ of the coast on the landward side is also evident; this is similar to that previously observed over geological boundaries on land.

621.396.11.029.51 3195

The Deviation of Low-Frequency Ground Waves at a Coast-Line—B. G. Pressey and G. E. Ashwell. (*Proc. IEE*, Part B, vol. 103, pp. 535–541; July, 1956.) "After consideration of the methods which have been suggested for computing the deviation of ground waves at a coastline, the phenomenon is reexamined in the light of recent experimental and theoretical work on the phase disturbances at such a boundary. It is shown that the deviation may be calculated from the rate of change of phase with distance along the path of propagation. The changes in this rate which occur at the boundary give rise to a considerable increase in the magnitude of the deviation as the receiving point is brought within a few wavelengths of that boundary. This increase near the coast seems to provide an explanation of the unexpectedly large deviations previously observed at medium frequencies. A series of simultaneous measurements of the phase change and the deviation at 127.5 kc along a number of paths crossing the south coast of England are described. Although general agreement between the measured deviations and those derived from the phase curves was obtained on some paths, there were appreciable discrepancies on others. These discrepancies are attributed to the irregularities in the phase surface which were evident over the area and which the method of derivation did not take into account."

621.396.11.029.55:551.510.535 3196

The Prediction of Maximum Usable Frequencies for Radiocommunication over a Transequatorial Path—G. McK. Allcock. (*Proc. IEE*, Part B, vol. 103, pp. 547–552; July, 1956.) "Times of reception of 15 mc radio waves over a transequatorial path of 7500 km have been recorded throughout the recent period of declining solar activity (1950–1954). The analysis of these times has shown that predictions of muf made by the usual control-point method were, in general, too high by about 4 mc, and at times by as much as 7 mc or more. This is contrary to the normal experience for long transmission paths lying within a single hemisphere. When a transmission mechanism involving multiple geometrical reflections is assumed instead of the forward-scattering mechanism implied by the control-point method, it is found that the path can be considered, for the purpose of predicting mufs, to consist of three reflections. The discrepancies between prediction and observations, which still remain after a 3-reflection mechanism has been invoked, are attributed mainly to reflections from the sporadic-E region at the southernmost reflection point, although it is possible that lateral deviation of the radio waves is also a contributing factor."

621.396.11.029.55:551.510.535 3197

Back-Scatter Ionospheric Sounder—E. D. R. Shearman and L. T. J. Martin. (*Wireless Engr.*, vol. 33, pp. 190–201; August, 1956.) Equipment is described for studying waves reflected from irregularities on the earth's surface and propagated back to the source via the ionosphere. The design of a suitable 150-kw pulse transmitter which can be simply tuned to any frequency in the band 10–27 mc is discussed. The same 3-wire rhombic antennas, are used for transmission and reception, with a tunable transmit-receive switch. A receiver of the type described by Piggott (2301 of 1955) provides an output suitable for presentation of the received echoes on a normal timebase display. A photographic record is made of this display, and continuous range/time ($p't$) records are also obtained. The same transmitter and receiver are also used with a continuously rotating Yagi antenna, and ppi. By using speeded-up kinematography. the changes occurring over 24 h may be shown in a few minutes. See also 1854 and 1855 of 1956 (Shearman).

621.396.11.029.6:551.510.52 3198

Some Considerations for the Field Strength of Ultra-short Waves at Night—K. Tao. (*J.

Radio Res. Labs., Japan, vol. 3, pp. 77–99; January, 1956.) The high level of field strength found locally at night is caused by reflection at a tropospheric inversion layer. The formation of such layers is discussed and related to the prevailing meteorological conditions.

621.396.11.029.62:551.510.52 3199
Investigations of the Propagation of Ultra-short Waves—R. Schünemann. (*Hochfrequenztech. u. Elektroakust.*, vol. 64, pp. 107–123; January, 1956.) Expressions are derived relating received field strength to atmospheric pressure, temperature, and humidity and their height gradients, while taking account of diffraction at the earth's surface. Verifying experiments were made over a 76-km path, using a frequency of 68 mc, with the transmitter antenna at a height of 90 m and the receiver antenna at a height of 30 m. The measured field strengths were correlated with meteorological observations; results are shown graphically for eight months, first for the main refracted and diffracted wave only, and then taking account of the reflected wave, which makes an effective contribution for 15–30 per cent of the time.

621.396.812.3:551.510.535 3200
A Correlation Treatment of Fading Signals—N. F. Barber. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 318–330; June, 1956.) An examination in terms of the complete correlogram is made of the fading signals observed at three receivers located at the apices of a right-angled isosceles triangle with equal arms of length 91 m. Methods based on three different sets of assumptions are used to interpret the correlograms in relation to ionospheric drifts. Discussion indicates that a quadratic method of analysis is not affected by decay in the correlogram.

621.396.11.029.62 3201
Atlas of Ground-Wave Propagation Curves for Frequencies between 30 Mc/s and 300 Mc/s [Book Review]—B. van der Pol. Publishers: International Telecommunication Union, Geneva, 1955, 35 pp. + 174 diagrams. (PROC. IRE, vol. 44, p. 952; July, 1956.) Information prepared at the request of the CCIR is presented regarding propagation over a spherical earth allowing for standard atmospheric refraction. The curves are preceded by an outline of the theory.

RECEPTION

621.376.23:621.396.822 3202
Interaction of Signal and Noise in an Inertial Detector—L. S. Gutkin. (*Radiotekhnika, Moscow*, vol. 11, pp. 43–53, February and pp. 51–62; March, 1956.) The detection by a linear inertial detector of a signal in the presence of noise is analyzed for the case when the signal is a) unmodulated, and b) amplitude modulated. The results are compared with the corresponding relations for a noninertial detector. The detector arrangement considered is a diode with *RC* circuit.

621.376.33:621.396.82 3203
Fourier Representation of a Demodulated Beat Oscillation—R. Leisterer. (*Elektronische Rundschau*, vol. 10, pp. 19–20; January, 1956.) The analysis presented shows that, if two sinusoidal signals, slightly differing in frequency, are applied via an ideal amplitude limiter to a linear wide-band fm discriminator, then the lf output voltage due to interference will increase with the signal frequency separation, and the waveform will depend on the amplitude ratio of the signals.

621.314.7:[621.396.62+621.395.625.3 +621.395.92 3204
Transistor Circuitry in Japan—(*Electronics*, vol. 29, pp. 120–124; July, 1956.) Circuits and characteristics of four types of broadcast receiver, a battery-operated tape recorder, and a hearing aid are given.

621.396.621+621.397.62 3205
Preventing Fires from Electrical Causes in the Design and Manufacture of Radio and Television Receivers—H. T. Heaton. (IRE TRANS., vol. BTR-1, pp. 28–36; April, 1955.)

621.396.621:621.396.828 3206
The Compensation of Interference in Carrier-Frequency Receivers by means of an Opposing Receiver connected in Parallel—H. Kaden. (*Frequenz*, vol. 10, pp. 76–82; March, 1956.) Rigorous analysis is presented for the nonideal case, *i.e.*, for circuits with arbitrary response characteristics over the pass band, assuming a sinusoidal signal of frequency within the pass band of the main receiver but outside that of the compensating receiver, and short interfering pulses. Rectifiers with square-law and broken-line characteristics are considered as demodulators; the broken-line characteristic leads to more effective elimination of the interference. For pulses occurring over a certain signal-phase range, the effect of the parallel receiver may be to increase the interference.

3207
621.396.621.029.62:621.396.662:621.314.63
Junction Diode A.F.C. Circuit—G. G. Johnstone. (*Wireless World*, vol. 62, pp. 354–355; August, 1956.) A circuit intended primarily for an fm receiver uses a junction diode biased to cut-off; in this condition the diode capacitance varies with the applied voltage.

621.396.8 3208
Asymmetry in the Performance of High-Frequency Radiotelegraph Circuits—A. M. Humby and C. M. Minnis. (*Proc. IEE*, Part B, vol. 103, pp. 553–558; July, 1956.) A further study has been made of the systematic differences which have been observed previously in the performance of radiotelegraph circuits for transmission in the two opposite directions [3394 of 1955 (Humby *et al.*)]. Measurements on transequatorial circuits suggest that the asymmetry is due at least partly to the combined effects of using directive receiving antennas, and the diurnal and seasonal changes in the sources of atmospheric noise.

621.396.82:621.327.43 3209
Evaluation of Radio Influence Voltages in Fluorescent Lighting Systems—F. H. Wright and S. A. Zimmermann. (*Elect. Engng., N.Y.*, vol. 75, pp. 272–274; March, 1956.) Interference with radio reception is caused mainly by supply-line radiation and by direct conduction. Elimination of interference by a low-impedance earth on the lighting system is unreliable; the connection of capacitors across individual lamps is most effective. In evaluating the efficiency of any filtering system a reference standard obtained by putting 0.01-μF capacitors across each lamp is recommended.

STATIONS AND COMMUNICATION SYSTEMS

621.376.56 3210
Coding of Signals by Damped-Oscillation Method—B. Carniol. (*Slab. Obz., Prague*, vol. 17, pp. 129–134; March, 1956.) A system of pulse coding which obviates the use of a coding tube is described. Voltages pulses of amplitudes proportional to the instantaneous amplitudes of the speech voltage, produced at intervals of 125 μs, excite an *LCR* circuit tuned to 500 kc. The resultant modulated voltage is passed through an amplitude limiter to a binary coder. Basic circuit diagrams of a simple coder and one with symmetrical logarithmic compression are given.

621.39:534.78 3211
The Vobanc—a Two-to-One Speech Band-width Reduction System—(See 2943.)

621.39.01:512.831 3212
Topological Properties of Telecommunication Networks—Z. Prihar. (PROC. IRE, vol. 44, pp. 927–933; July, 1956.) A method of matrix analysis developed in connection with sociological studies is applied to investigate problems relating to the connections between a number of points. Numerical examples are given.

621.396 3213
Symposium on Communications by Scatter Techniques—(See 3185.)

621.396.41+621.395.43]:621.376.3 3214
An Extended Analysis of Echo Distortion in the F.M. Transmission of Frequency Division Multiplex—R. G. Medhurst and G. F. Small. (*Proc. IEE*, Part B, vol. 103, pp. 447–448; July, 1956.) Discussion on 1867 of 1956.

621.396.41:621.376.3 3215
Multiprogram F.M. Broadcast System—W. N. Hershfield. (*Electronics*, vol. 29, pp. 130–133; June, 1956.) A system is described in which three additional programs with band-width 10 kc are transmitted by fm on subcarriers 28, 49, and 67 kc above the main broadcast carrier. Detailed circuit diagrams are given of the subcarrier generator with serrasoid modulator, the transmitter exciter stage, the main-channel receiver, and a subcarrier demodulator unit.

621.396.41.029.6:621.376.3:621.396.82 3216
Nonlinear Distortion in Multichannel Communication Systems with Frequency Modulation—V. A. Smirnov. (*Radiotekhnika*, Moscow, vol. 11, pp. 14–28; February, 1956.) Noise due to multipath propagation and wave-guide mismatch is considered theoretically. The results are more general than those obtained by Borovich (*ibid*, vol. 10, pp. 3–14; October, 1955) and by Bennett et al. (3089 of 1955).

621.396.5:621.396.4 3217
The Copenhagen-Thorshavn Radiotelephony Link—S. Gregersen. (*Teleteknik, Copenhagen*, vol. 7, pp. 15–34; February, 1956.) Detailed description of this hf multichannel system.

621.396.65 3218
V.H.F. Radio Link in the West Indies—R. McSweeny. (*Elect. Engng, N.Y.*, vol. 75, p. 271; March, 1956.) Digest of paper published in *Trans. Amer. IEE*, Part I, *Communication and Electronics*, vol. 74, pp. 781–785; January, 1956. Details are given of two radio links over 69 miles and 45 miles respectively, using fm transmissions on frequencies of 150–160 mc.

621.396.7+621.397.7](47) 3219
Broadcasting in the U.S.S.R.—(*Wireless World*, vol. 62, pp. 379–381; August, 1956.) Some technical details of the sound and vision services are given, with a note on the television standards.

621.396.7(492):621.376.3]+621.397.7(492) 3220
A Survey of the TV and F.M. Projects in the Netherlands—J. L. Bordewijk. (*PTT-Bedrijf*, vol. 7, pp. 1–12; March, 1956. In English.)

621.396.71(489) 3221
Coast Stations in Denmark—K. Svenningsen. (*Teleteknik, Copenhagen*, vol. 7, pp. 1–14; February, 1956.) The radio stations at Thorshavn, Skagen (The Skaw), and Rønne are described; telegraphy and telephony services are handled.

SUBSIDIARY APPARATUS

621-526 3222
An On-Off Servomechanism with Predicted Change-Over—J. F. Coales and A. R. M. Noton. (*Proc. IEE*, Part B, vol. 103, pp. 449–460; July, 1956. Discussion, pp. 460–462.) "A general method has been devised for

achieving optimum switching with an on-off control system. The practicability of predicting the ideal switching time has been demonstrated with a model experiment for which responses to step, ramp, and parabolic input functions have been found to compare favorably with those of an orthodox system."

621-526 3223
The Dual-Input Describing Function and its Use in the Analysis of Nonlinear Feedback Systems—J. C. West, J. L. Douce and R. K. Livesley. (*Proc. IEE*, Part B, vol. 103, pp. 463–473; July, 1956. Discussion, pp. 473–474.)

621.3-71:537.32:537.311.33 3224
Thermoelectric Cooling—L. S. Stil'bans, E. K. Iordanishvili, and T. S. Stavitskaya. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 20, pp. 81–88; January, 1956.) A brief account is given of A. F. Ioffe's theory of thermoelectric cooling (*Energetical Bases of Semiconductor Thermo-Batteries*, published by the U.S.S.R. Academy of Sciences, Moscow, 1956) and of experimental results. Temperature differences up to 70°C have been obtained. Applications being investigated include cooling of components in radio and electronic equipment.

621.314.63:546.28 3225
Diffused p-n Junction Silicon Rectifiers—M. B. Prince. (*Bell Syst. Tech. J.*, vol. 35, pp. 661–684; May, 1956.) Development types with current ratings up to 100 a for reverse peak voltages of 200 v or over are described. Operation is satisfactory at temperatures up to 200°C.

621.314.63:546.28 3226
The Forward Characteristic of the P-I-N Diode—D. A. Kleinman. (*Bell Syst. Tech. J.*, vol. 35, pp. 685–706; May, 1956.) Theory for the *p-i-n* Si diffused-junction diode indicates that the forward characteristic should be similar to that of the simple *p-n* diode until the current density approaches 200 a/cm^2; anomalies in the characteristic at low current densities are unrelated to the presence of the weakly *p* middle region. See also 3225 above.

621.362:537.311.33:537.32 3227
Thermoelectric Generators—A. F. Ioffe. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 20, pp. 76–80; January, 1956. In Russian.) Basic design formulas are given and discussed. Using a semiconductor layer 0.5 cm thick, with thermoelectric coefficient 170×10^{-6} v/deg, a temperature difference of 300°C across it, and a heat input of 11.6 w/cm^2, and assuming a specific mass of 5 and an efficiency of 8 per cent, an output of 0.2 kw/kg may be obtained.

TELEVISION AND PHOTOTELEGRAPHY

621.397.611.2:525.623:621.397.7 3228
The 'Vitascan'—New Color TV Scanner—C. E. Spicer. (*Tele-Tech & Electronic Ind.*, vol. 15, pp. 60–61, 117; February, 1956.) A spot of white light, generated on the screen of a cr tube by a beam deflected at the standard television rate, is projected on the scene and the reflected light is picked up by fixed photomultiplier tubes, associated with color filters, which generate the video signal. General studio lighting is provided by pulsing xenon lamps to be on during the vertical retrace time of the television signal.

621.397.62+621.396.621 3229
Preventing Fires from Electrical Causes in the Design and Manufacture of Radio and Television Receivers—H. T. Heaton. (IRE Trans., vol. BTR-1, pp. 28–36; April, 1955.)

621.397.62 3230
A Television Receiver Suitable for Four Standards—H. L. Berkhout. (*Philips tech. Rev.*, vol. 17, pp. 161–170; December, 1955.) A model suitable for receiving the Belgian 625- and 819-line, the European 625-line, and the French 819-line standards is described. A common vision IF amplifier is used, the frequency being 38.9 mc and the bandwidth 4 mc. The video signal is applied to the picture-tube control grid for positive modulation and to the cathode for negative modulation. Different sound intermediate frequencies are again converted to a common second IF of 7 mc. Flywheel synchronization is used for the horizontal deflection.

621.397.62:525.623 3231
Chrominance Circuits for Colour-Television Receivers—B. W. Osborne. (*Electronic Engng.*, vol. 28, pp. 240–246, 293–297; June /July, 1956.) "A survey of current practice and recent developments in phase synchronization, chrominance demodulator and matrix circuits for use in color-television receivers."

621.397.621:535.623:621.385.832 3232
Television Receiver uses One-Gun Color C.R.T.—(*Electronics*, vol. 29, pp. 150–153; June, 1956.) A description is given of the "apple" tube. An electron beam sequentially strikes vertical phosphor stripes arranged in triplets of red, blue and green on an aluminized screen, with interstices filled with nonluminescent material. Applied behind each red stripe and covering about 40 per cent of the triplet width is an "indexing" stripe of MgO with high secondary-emission characteristic. An intensity-modulated pilot beam from the same electron gun is aligned so that it strikes the same color stripe as the main beam, and the secondary-emission current is used to derive an indexing signal controlling the amplitude and phase modulation of the main signal to produce a color display. Block diagrams and some details of the associated receiver circuit are given.

621.397.7 3233
Optical Multiplexing in Television Film Equipment—A. H. Lind and B. F. Melchionni. (*J. Soc. Mot. Pict. Telev. Engrs*, vol. 65, pp. 140–145; March, 1956. Dsicussion, p. 145.)

621.397.7+621.396.7](47) 3234
Broadcasting in the U.S.S.R.—(See 3219.)

621.397.7(492)+[621.396.7(492):621.376.3 3235
A Survey of the TV and F.M. Projects in the Netherlands—J. L. Bordewijk. (*PTT-Bedrijf*, vol. 7, pp. 1–12; March, 1956. In English.)

621.397.8:621.372:621.3.018.752 3236
The Effect upon Pulse Response of Delay Variation at Low and Middle Frequencies—Callendar. (See 2984.)

TRANSMISSION

621.376.22:621.317.6:621.385.5 3237
Study of Amplitude Modulation applied via a Pentode Suppressor Grid—J. Loeckx. (*Rev. HF*, Brussels, vol. 3, pp. 183–190; 1956.) With this method of modulation, the pentode screen grid is maintained at fixed potential. The relation between the anode current and the grid and anode voltage is derived, and the equation of the modulation characteristic is hence determined explicitly. A measurement method particularly suitable for obtaining the characteristics of power tubes is outlined.

621.396.61:621.396.662 3238
Automatic Tuning for High-Power Transmitter—V. R. DeLong. (*Electronics*, vol. 29, pp. 134–137; July, 1956.)

TUBES AND THERMIONICS

621.314.63(47):546.289 3239
Germanium Diodes—A. N. Puzhai. (*Avtomatika i Telemekhanika*, vol. 17, pp. 140–146; February, 1956.) Discussion of the characteristics of point-contact and junction-type Ge diodes available in Russia.

621.314.632:546.289 3240
Effect of Vacuum Heating and Ion Bombardment of Germanium on Point Contact Rectification—R. B. Allen and H. E. Farnsworth. (*J. Appl. Phys.*, vol. 27, pp. 525–529; May, 1956.) Measurements were made of the characteristics of diodes comprising a Ge crystal with a tungsten or columbium point contact, to determine whether an adsorbed gas layer on the Ge surface is a prerequisite for rectification. Ge surfaces free from such layers are obtained by vacuum heating and argon-ion bombardment. The best rectification characteristics were obtained after the Ge had been subjected to a long anneal, argon-ion bombardment, and a short anneal, in that order. The diode activation potential does not appear to be dependent on the metallic work function.

621.314.7(083.7) 3241
IRE Standards on Letter Symbols for Semiconductor Devices, 1956—(Proc. IRE, vol. 44, pp. 934–937; July, 1956.) Standard 56 IRE 28. S1 on transistors.

621.314.7.002.2 3242
Automatic Etching of Transistor Pellets—(*Electronics*, vol. 29, pp. 226, 236; July, 1956.) A description of the etching, washing, and indium plating of concentric holes in Ge or Si pellets for surface-barrier transistors. The precision electrochemical etching is controlled by a light beam and photocell.

621.314.7:537.311.33 3243
Propagation of a Short Pulse in a Semiconductor bounded by Two Electron-Hole Transistions—E. I. Adirovich and V. G. Kolotilova. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 770–777; December, 1955.) The propagation of a short pulse in a *p-n-p* transistor is considered theoretically. Using the continuity equation for holes, an expression is derived for the concentration of nonequilibrium carriers at an arbitrary cross section due to application of the pulse at the emitter. The collector current is calculated for various values of lifetime of the nonequilibrium carriers, and the effect of the boundary conditions on the electron processes in the body of the semiconductor is discussed.

621.314.7:621.318.57 3244
A Switching Transistor with Short Transition Times—H. Salow and W. v. Münch. (*Z. Angew. Phys.*, vol. 8, pp. 114–119; March, 1956.) A characteristic with an unstable region is obtained by adding an auxiliary collector adjacent to the usual collector of a junction transistor. In an experimental *n-p-n* unit with base thickness of 50μ, a change of emitter/base resistance from 1MΩ to 20Ω was achieved in 2×10^{-7}s. The theory and the characteristics are discussed.

621.314.7:621.387.4 3245
The Application of Transistors to the Trigger, Ratemeter and Power-Supply Circuits of Radiation Monitors—Franklin and James. (See 3175.)

621.314.7:621.396.822 3246
Microphonism due to Transistor Leads—C. W. Durieux and T. A. Prugh. (Proc. IRE, vol. 44, pp. 938–939; July, 1956.) A brief note of observations of voltages generated by the vibrations of transistor leads in a magnetic field.

621.38.004.6 3247
Reliability as a Design and Maintenance Problem—R. Matthews. (*Electronic Engng.*, vol. 28, pp. 310–312; July, 1956.) The subject is discussed particularly in relation to tube performance.

621.383.27:621.387.464 3248
Study of the First-Stage Focusing of a Phomultiplier Tube for Scintillation Counting—G. Wendt. (*Ann. Radioélect.*, vol. 10, pp. 372–386; October, 1955.)

621.383.4 3249
The Photo-effect in Lead Sulphide and Related Materials—R. Stein and B. Reuter. (*Z. Naturf.*, vol. 10a, pp. 655–665; August, 1955.) Discussion of photoelectric inertia effects which have been traced to the presence of excess sulphur. Experiments are reported which indicate that these effects are probably related to the sensitization of the PbS cell by the usual method involving oxidation.

621.383.4/.5:546.817.221 3250
***p-n* Junctions in Photosensitive PbS Layers**—J. Bloem. (*Appl. Sci. Res.*, vol. B6, pp. 92–100; 1956.) PbS layers containing sharp *p-n* junctions can be produced by precipitation from an aqueous solution on to a glass plate partially coated with a thin layer of a trivalent metal; immediately after deposition, the whole of the PbS layer is of *n* type, but the portion on the uncoated glass is converted to *p* type soon after coming into contact with the air. Measurements of the photo-emf and resistance of such cells are reported; variations with storage time were investigated. The influence of oxygen in the ambient gas is discussed.

621.383.5 3251
The Photo-Electromotive Force of Lead Sulphide Photocells—R. Ya. Berlaga, M. A. Rumsh, and L. P. Strakhov. (*Zh. tekh. Fiz.*, vol. 25, pp. 1878–1882; October, 1955.) Layers of PbS were obtained in which an emf appeared during illumination, although no voltage was applied during their preparation. The photo-emf of freshly prepared specimens was of the order of a few mv. When the specimens were heated to temperatures between 500° and 600°C, the photo-emf increased to 3 v. The experimental investigation is described, electron-diffraction diagrams are reproduced, and a theoretical interpretation of the results is given.

621.385.029.6 3252
Theory of the Transverse-Current Traveling-Wave Tube—D. A. Dunn, W. A. Harman, L. M. Field, and G. S. Kino. (Proc. IRE, vol. 44, pp. 879–887; July, 1956.) Tubes are discussed in which an extended beam approaches the helix from the side, either normally or at an angle; each electron, instead of traveling the length of the helix, cuts across it and interacts with it for only a fraction of its length. Three forward waves are produced in such a system. Expressions are derived for the over-all gain. The power output reaches saturation for a given value of input and stays at this value with further increase of input.

621.385.029.6 3253
An Experimental Transverse-Current Traveling-Wave Tube—D. A. Dunn and W. A. Harman. (Proc. IRE, vol. 44, pp. 888–896; July, 1956.) Details are given of the construction and performance of a tube of the class discussed by Dunn et al (3252 above) using a flat helix and a skew beam. The tube operates as an amplifier over the frequency range 1–2 kmc with a power output of the order of 30 mw. The gain/voltage characteristic is markedly different from that of a conventional traveling-wave tube; high attenuation is observed over a wide range of current and voltage values. Gain/current, gain/frequency and saturation-power/frequency characteristics are as predicted by the theory. Experiments are described in which two input signals of different frequencies were applied simultaneously.

621.385.029.6 3254
Some Effects of Magnetic Field Strength on Space-Charge-Wave Propagation—G. R. Brewer. (Proc. IRE vol. 44, pp. 896–903; July, 1956.) General analysis is presented for the propagation of space-charge waves in magnetically focused electron beams. The propagation characteristics for the fundamental radial mode are expressed in terms of the plasma-frequency reduction factor, graphs of which are shown. The case of a beam within a helix, as in the traveling-wave tube, is examined particularly.

621.385.029.6 3255
Study of the Oscillation Modes of the M-Type Carcinotron: Part 1—M. de Bennetot. (*Ann. Radioélect.* vol. 10, pp. 328–343; October, 1955.) The starting current and oscillation frequency are determined theoretically, taking account of space-charge effects. The field of the space harmonic interacting with the electron beam in this case is constituted by the sum of three traveling waves.

621.316.726:621.385.029.6:621.396.96 3256
Klystron Control System—Reeves. (See 3068.)

621.385.029.6:621.396.822 3257
A Dip in the Minimum Noise Figure of Beam-Type Microwave Amplifiers—P. K. Tien. (Proc. IRE, vol. 44, p. 938; July, 1956.) A detailed computation has been made of the fluctuations of electron current and velocity at the potential minimum of a particular tube. The results indicate that the velocity fluctuation is not smoothed and the fluctuations of current and velocity are not correlated. A physical explanation is given of the resulting shape of the cumulative autocorrelation curve. The minimum noise figure for a typical traveling-wave tube as calculated from this autocorrelation curve shows a dip at about 2.5 kmc and a peak at about 4 kmc.

621.385.029.6.032.21:536.52 3258
A New Method for the Measurement of Rapid Fluctuations of Temperature—R. Dehn. (*Brit. J. Appl. Phys.*, vol. 7, pp. 144–148; April, 1956.) Instantaneous changes in cathode surface temperature in an oscillating magnetron are displayed and measured as pulses on a cro screen by means of an infrared-image converter and photomultiplier. The instrument is calibrated against an optical pyrometer; changes of 2°C at 900°C have been detected.

621.385.032.21:537.58 3259
Thermionic Emission Properties of Thin Films of Thorium Oxide and Thorium on Metallic Bases—A. R. Shul'man and A. P. Rumyantsev. (*Zh. tekh. Fiz.*, vol. 25, pp. 1898–1909; October, 1955.) Report on an experimental investigation of thin films of ThO_2 and Th deposited on Mo and Pt bases. The deposition of the films is described in detail and a large number of experimental curves is given. The results are discussed and various suggestions regarding the mechanism of thermionic emission are made.

621.385.032.216 3260
Radioactive Isotope Study of the Dissociation of Barium Oxide under Electron Bombardment—S. Yoshida, N. Shibata, Y. Igarashi, and H. Arata. (*J. Appl. Phys.*, vol. 27, pp. 497–500; May, 1956.) Measurements are reported of the rate of evolution of Ba; the number of Ba atoms produced per bombarding electron is plotted as a function of bombarding-electron voltage and of oxide temperature. The results are qualitatively similar to those for SrO (*J. Phys. Soc. Japan*, vol. 9, pp. 640–641; July/August, 1954). Discussion indicates that they can be reconciled with those of Leverton and Shepherd (3601 of 1952).

621.385.132:681.142 3261
Binary Adder uses Gas-Discharge Triode—F. B. Maynard. (*Electronics*, vol. 29, pp. 196, 202; June, 1956.) The elementary triode cell has a large-area cathode and closely overlaid anode element of fine wire. A probe element in the upper part of the cathode glow, common to a number of cells, acquires a positive charge. The voltage excursion at the probe can be as much as 30 v without causing a discharge in cells other than that actuated. Experimental tubes with a matrix of 30 of these cells have been tested.

621.385.5.032.24:621.374.3 3262
A New High-Slope Multigrid Valve and its Application in Pulse and Switching Circuits—K. Gosslau and W. Guber. (*Frequenz*, vol. 10, pp. 83–89; March, 1956.) An experimental heptode Type-V108 with three frame grids had slopes of 13 and 7.5 ma/v respectively at the two control grids, high pulse current intensity, and adequate loading capacity at the first screen grid. A pulse distributor using this tube is described.

621.385.832:621.397.621:535.623 3263
Television Receiver uses One-Gun Color C.R.T.—(See 3232.)

MISCELLANEOUS

061.6:621.396 3264
International Cooperation in Radio Research—URSI and IRE—J. H. Dellinger. (Proc. IRE, vol. 44, p. 866–872; July, 1956.) The internal structure of the International Scientific Radio Union is described, and its relations with the CCIR and the IRE are explained.

621.3:537 3265
Advances in Electronics and Electron Physics, Vol. VII [Book Review]—L. Marton (Ed.). Publishers: Academic Press, New York, 1956, 503 pp., Proc. IRE, vol. 44, Part 1, pp. 828–829; June, 1956.) Review articles are presented on the physics of semiconductor materials, theory of electrical properties of Ge and Si, energy losses of electrons in solids, sputtering by ion bombardment, observational radio astronomy, analog computers, and electrical discharge in gases and modern electronics.