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# PROCEEDINGS OF THE IRE®

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## CONTENTS




THE COVER—The newly opened New York Coliseum will be the scene this month of the largest technical exhibition ever held, as the IRE National Convention gets under way on March 18. All four floors of the huge building will be used to house the latest products and most important developments of some 840 electronics firms. Technical sessions will be held at both the Coliseum and the nearby Waldorf-Astoria Hotel. Full details concerning the technical program start on page 371, and a list of exhibitors and their products appears in the advertising section of this issue.

# Raymond A. Heising

WINNER OF THE 1957 FOUNDERS AWARD

Raymond A. Heising was born August 10, 1888, at Albert Lea, Minn. He received a degree in electrical engineering from the University of North Dakota in 1912 and the master's degree from the University of Wisconsin in 1914.

From 1914 until his retirement in 1953, Dr. Heising was associated with the Western Electric Company and Bell Telephone Laboratories. He played a major role in the development of radiotelephone systems for military use in World War I, and for transoceanic and ship-to-shore public service use. He also conducted much research work on ultra-short waves, electronics and piezoelectric crystal devices. He invented several modulation systems which are still in wide use today: the constant potential system, the grid modulation system for radio, the rectifier modulation system used in carrier telephony, and in particular, the constant-current or Heising modulation system. He has over one hundred United States patents, including the patents on the Class C amplifier and diode-triode detector amplifier circuit, and has published numerous technical papers in engineering journals.

Since 1953 Dr. Heising has been engaged as an independent consulting engineer and patent agent.

He is a Fellow of the American Institute of Electrical Engineers and American Physical Society. He was awarded the Modern Pioneer Award from the National Association of Manufacturers in 1940. He received an honorary Doctor of Science degree from the University of North Dakota in 1947, and the Armstrong Medal from the Radio Club of America in 1954.

An Associate of the IRE since 1920 and winner of the Morris Liebmann Memorial Prize in 1921, he became a Fellow in 1923. He served as IRE President in 1939, Treasurer from 1943 to 1945, and as an elected member of the Board of Directors for seventeen years. He was chairman of numerous IRE committees, among them the committees on Sections, Professional Groups and Office Quarters.

Dr. Heising will receive the IRE Founders Award, an award which is given only on special occasions to outstanding leaders in the radio industry, at the banquet of the annual IRE National Convention this month. The award is being bestowed upon him "for his leadership in Institute affairs, for his contributions to the establishment of the permanent IRE Headquarters, and for originating the Professional Group system."

# Julius A. Stratton

WINNER OF THE 1957 MEDAL OF HONOR

Born at Seattle, Wash. in 1901, Julius A. Stratton attended the University of Washington for one year until he came to the Massachusetts Institute of Technology to obtain his bachelor's degree in 1923 and master's degree in 1925. He then did graduate study in Grenoble and Toulouse, France, and the Technische Hochschule of Zurich, Switzerland, awarded him the degree of Doctor of Science in 1927. During the following year he studied under a traveling scholarship from M.I.T., chiefly at Munich and Leipzig, Germany.

Dr. Stratton joined the staff of M.I.T. in 1928 to serve in the electrical engineering and physics departments for twenty years. In 1945 he was appointed Director of the Research Laboratory of Electronics. He was named Provost of M.I.T. in 1949, and in 1951 he became vice-president as well. Last year he was named to the newly-created post of Chancellor. As Chancellor, Dr. Stratton administers the academic program of the institution, acts as deputy to the president, and serves as general executive officer.

During World War II he served as Expert Consultant in the Office of the Secretary of War, and was awarded its Medal for Merit in 1946.

He is a Fellow of the American Institute of Physics and the American Academy of Arts and Sciences, and a member of the American Philosophical Society, the National Academy of Sciences, Tau Beta Pi, and Sigma Xi. He is an appointed trustee of the Ford Foundation, and one of the nine-member National Science Board of the National Science Foundation.

Dr. Stratton is the author of numerous technical papers and books on theoretical physics, especially electromagnetic theory. His many activities include membership on the Defense Science Board, the National Science Foundation Advisory Committee on Government-University Relations, and the Naval Research Advisory Committee.

He became an IRE Member in 1942, a Senior Member in 1943, and a Fellow in 1945. He served on the Board of Directors from 1948–1951, and again in 1954. He headed the Radio Wave Propagation and Utilization Technical Committee from 1945 to 1948.

Dr. Stratton will receive the Medal of Honor, IRE's highest technical award at the convention banquet this month "for his inspiring leadership and outstanding contributions to the development of radio engineering as a teacher, physicist, engineer, author and administrator."

# Scanning the Issue

**On the Statistics of Individual Variations of Productivity in Research Laboratories** (Shockley, p. 279)—A novel statistical study has been made of the large differences in scientific creativity that exist between individual research workers. The author uses as his yardstick the number of publications that an individual has made, showing first that for the purposes of this study this is a surprisingly accurate method of measuring scientific productivity. A study of the publication rates of individuals in a number of research laboratories reveals that the variations in productivity are exponential in character. The author explains this phenomenon on the basis of simplified models of the mental processes concerned. For example, the man who can grasp and relate five ideas at the same time is probably capable of producing a great many more inventions than the man who can grasp only four ideas. Thus a small change in certain factors of the mental process can result in a very large change in output. The study closes with an examination of the relationship between salary and productivity which leads to some important conclusions regarding the effect which the present limits on government salaries is having on the originality and leadership available in government laboratories.

**Molecular Amplification and Generation of Microwaves** (Wittke, p. 291)—Continuing its series of invited review papers, the PROCEEDINGS presents this month a discussion of a radically different method of generating and amplifying microwaves which has greatly excited the interest of scientists during the last two or three years. This new method, which is still in the early stages of development, makes use of the internal energy of molecules rather than the movement of charged particles as a source of rf energy, and opens the door to a whole new class of gaseous and solid-state microwave devices. The operation of these devices is tied to the fact that the internal energy systems of molecules exist in several distinct energy states. Upon exposure to a microwave field of the right frequency, a molecule in a lower energy state may jump to a higher energy state, absorbing energy from the field in the process. More important, a high-level molecule may drop to a lower energy level and give up some of its internal energy. Under normal circumstances, all molecular systems have a predominance of low-level molecules, and thus are absorptive. Hence the problem of building a molecular amplifier is one of first creating an "unnatural" condition in which the high-level molecules predominate and then inducing them to give up some of their energy. The several methods thus far proposed for accomplishing this, and the underlying physical principles involved, are the subject of this thorough and well-written discussion. The timeliness of this article is attested to by the fact that just one month ago a leading laboratory announced the first successful operation of a solid-state microwave oscillator utilizing these new principles. As further progress is made, we can look forward to new devices which, among other things, will provide us with extremely low noise amplification for communications and radio astronomy applications, super-stable oscillators for frequency standards, and a promising method of generating frequencies in the millimeter and submillimeter range.

**The Spacistor, A New Class of High-Frequency Semiconductor Devices** (Statz and Pucel, p. 317)—A new approach is suggested for extending the useful frequency range of semiconductor devices into the microwave region. The relatively slow diffusion of minority carriers through the almost field-free regions of the base is an important frequency-limiting factor in present devices. In the new class of devices proposed here, electrons or holes are injected directly into space-charge regions, and the slow-diffusion areas are thus avoided. The spacistor is characterized by junctions which are reverse-biased at voltages high enough for the injected carriers to be multiplied by the avalanche process. The principles, problems and suggested solutions presented here lay important groundwork for the development of higher frequency semiconductor devices.

**Ferrite Apertured Plate for Random Access Memory** (Rajchman, p. 325)—A simple method of fabricating a plate containing a large array of memory cells has been developed that promises to greatly increase the storage capacity of even the largest present-day digital memory systems. The plate, which is made of nonconducting ferrite material, is molded with an array of holes in it and is then coated with a conductive pattern to form a printed winding linking all holes in series. Large stacks of these plates can be readily wired together to form an easily assembled large-scale memory. This is a major improvement over current laborious methods of assembling memory cells individually, and opens possibilities of memories with capacities of millions of bits.

**Stability and Power Gain of Transistor Amplifiers** (Stern, p. 335)—This paper offers the practicing engineer a new and practical measure of the maximum realizable stable power gain of a transistor amplifier that operates at frequencies where it could become unstable. The clarity of the presentation, as well as the importance of the contribution, will make this work of general interest to practically all people who design any but audio frequency types of transistor amplifiers. The results and point of view are interesting also for the cases of very-high-frequency triode tube amplifiers, where the same sort of problems arise and where the analysis presented applies equally well.

**Ferriod Radiator System** (Reggia, et al, p. 344)—The dielectric, magnetic and polarization characteristics of ferrites have been brought together in this paper to perform a variety of useful functions in a microwave antenna system. The radiating elements consist of dielectric rods made of ferrite, which are shaped to give maximum radiation along the axis of the rods and minimum radiated power in the side lobes. Ferrite-loaded cavities and waveguides are used for a number of switching and coupling schemes, including devices for beam lobing and array scanning. The small size of these systems will make them very useful in reducing the size of microwave antenna systems, a matter of considerable importance, for example, in portable radars.

**IRE Standards on Piezoelectric Crystals—The Piezoelectric Vibrator: Definitions and Methods of Measurement** (p. 353)—The IRE Committee on Piezoelectric Crystals has produced a timely and useful Standard that specifies the nomenclature and a practical method of measuring the various quantities associated with piezoelectric vibrators.

**IRE National Convention Program** (p. 371)—A program is given of the Convention activities and abstracts of the 284 papers which will be presented on the 18th through the 21st of this month at the New York Coliseum and Waldorf-Astoria Hotel.

**Whom and What to See at the Radio Engineering Show** (p. 110A)—A full list of the 840 Convention exhibitors, booth numbers and personnel, and products that you will meet and see on March 18–21 at the New York Coliseum is given, with floor plans showing booth locations.

# Poles and Zeros

**Big Step.** On January 4th last, the Board of Directors by unanimous vote set in motion the IRE Group Affiliate Plan. This plan, originally proposed by Editor Emeritus A. N. Goldsmith, permits members of other professional societies to participate as individuals in Professional Group activities and to receive PG TRANSACTIONS without prior admission to IRE membership. The details of the Affiliate Plan are presented on page 278 of this issue by W. R. G. Baker, Chairman of the Professional Groups Committee.

This is a big step for the IRE, certainly its most important action since the founding of the Professional Group system. The proposal was debated for months before its approval by the Board because, as Dr. Baker points out, the Affiliate Plan carries with it the risk that Affiliate status may be elected by some specialists who, in good conscience, should become IRE members.

The Plan has been worked out to minimize this risk. For example, no one who has been an IRE member during the five years prior to his application will be admitted as an Affiliate. The risk is thus confined to the future, but it is possible that the growth of IRE membership may be diminished if Affiliate status is more attractive than IRE membership to any substantial number of potential members.

We predict that no such trend will develop, that, in fact, more members will be attracted to full membership in the IRE through initial exposure as Group Affiliates than will be lost. Such a counter-trend can be expected to appear if three prerequisites are met. First, it depends on the continued conviction of IRE members, particularly those active in PG affairs, that the parent IRE serves a purpose and provides services which transcend individual specialties—a conviction that the IRE is, and must continue to be, more than a mere holding company for Professional Group assets. Second, it depends on recognition by technical workers generally that over-specialization is dangerous, that lack of interest in other fields marks a sterile intellect, and is injurious to the man and his profession.

Third, as a matter of hard practicality, it depends on value received. According to the present rules, Group Affiliate status in the majority of the Groups costs $6.50 per year. IRE membership costs $10.00 the first three years and $15.00 thereafter. What does the IRE member get for the difference? He gets the privilege of holding office but this affects few members since there are less than 500 elective and appointive offices in the Sections, Groups and national organization. The privilege of holding office is real, but it is not a compelling influence.

The big difference is the PROCEEDINGS. Members receive it. Affiliates do not. Affiliates will apply for full membership, by and large, because they want to read the PROCEEDINGS. This is the challenge which the Affiliate Plan presents to the Editorial Board, its heirs and assigns. With the help of all concerned, we aim to meet it.

**Back Issues.** One of the many incidental useful services performed by the editorial staff is bringing together those who have back issues of IRE publications to dispose of and those who wish to acquire same. If you have back issues to offer, two courses of action are open: (1) we will send you a list of dealers in second-hand technical periodicals who may, if the collection is complete and goes back a number of years, offer you a modest sum and pay the freight into the bargain; or (2) if such negotiations come to naught, the Institute may take your back issues off your hands if they are in short supply, but will limit its outlay to the transportation charges to IRE Headquarters in New York. If you elect the latter option, please communicate your intentions to IRE before shipping, describing your collection and its condition. HQ will respond promptly with shipping instructions or will suggest other means of disposal.

If you are on the acquisition side of the brokerage, write stating your needs. Back issues from our regular stock will be sold to you at the established prices. If you prefer, or if we are out of stock, we will send the list of second-hand dealers, and you take it from there.

Incidental intelligence: the back-issue demand displays a pronounced peak 17 years after the date of issue. From this we deduce that patent searchers, hoping to find an anticipation of an issued patent (which expires in 17 years), are among our best customers.

**Meetings.** Members who wonder whether the two mansions the IRE occupies as Headquarters are useful as well as ornamental will be reassured by the fact that, during the 260 working days of 1956, no fewer than 391 committee meetings were held at Numbers 1 and 5, East 79th Street.

# The IRE "Affiliate" Plan—A New Venture in Engineering Society Structure and Service

W. R. G. BAKER, *Chairman*
*IRE Professional Groups Committee*

On January 4, 1957 the IRE Board of Directors arrived at a decision which may in time prove to be one of the most far-reaching in its 45-year history. On that date the Board adopted a plan which will enable non-IRE members whose main professional interests lie outside the sphere of IRE activities to become affiliated with certain of the IRE Professional Groups *without* first having to join the IRE itself.

This plan is aimed at those specialists in other fields of science and technology whose work touches upon our own electronics and communications field only in specialized areas. In effect, the IRE is extending the specialized services of its Professional Groups to every field of science and engineering.

An outstanding example of where these services are needed may be found in the case of the medical and biological sciences. At the present time some 1400 IRE members enjoy the privileges of membership in the Professional Group on Medical Electronics. And yet there are hundreds, perhaps thousands, of medical doctors, biologists, and others to whom the activities of this Group would be of interest and value. Both they and the Group would benefit from their participation. To require these persons, who have no interest in radio engineering, to join the IRE in order to join the Group is unreasonable, and probably futile as well. In fact, it was largely to provide an answer to this particular problem that the Affiliate Plan was first conceived, although it pertains to other fields as well, such as Computers, *etc.*

The Affiliate Plan is admittedly an experiment. So far as is known, no other society has ever tried a similar scheme. The Board of Directors feels strongly that the benefits afforded by the plan justify the risk that some persons who should join the IRE will instead become Affiliates. To minimize this risk, the plan has been carefully worked out along the following lines:

1) Participation in the plan is at the option of each Professional Group. It is not expected that all Groups will adopt it; only those which feel it serves a need in their particular field.

2) Each Group interested in initiating the Affiliate Plan must submit to the Chairman of the Professional Groups Committee a list of accredited organizations which has been selected and approved by its Administrative Committee, for official approval by the IRE Executive Committee.

3) To be an Affiliate of a Professional Group, a person must belong to an accredited organization approved by that Group and the IRE Executive Committee. Moreover, he shall not have been an IRE member during the five years prior to his application. He may affiliate with more than one Group, provided the accredited organization to which he belongs is recognized by the Groups concerned.

4) The fee for Affiliates shall be the assessment fee of the Group, plus $4.50. The latter covers IRE subsidies to the Group, Professional Group overhead expenses borne by IRE Headquarters, and 50 cents which is to be rebated to IRE Sections for mailing and meeting costs.

5) An Affiliate will be entitled to receive the TRANSACTIONS of his Group and that part of the IRE National Convention Record pertaining to his Group. He will be eligible for a Group award, and may attend local or national meetings of the Group by payment of charges assessed Group members.

6) An Affiliate cannot serve in an elective office in the Group or Group Chapter, nor vote for candidates for these offices.

7) An Affiliate may hold an appointive office in the Group or Group Chapter.

8) An Affiliate may not receive any IRE benefits that are derived through IRE membership.

The Affiliate Plan is a bold and farsighted venture; one that recognizes and provides for the rapidly spreading influence of electronics in every walk of scientific and technological life, and one that enables the IRE to further its aims as a professional engineering society—the advancement of radio engineering and related fields of engineering and science.

# On the Statistics of Individual Variations of Productivity in Research Laboratories*

WILLIAM SHOCKLEY†, FELLOW, IRE

In the following pages a co-winner of the 1956 Nobel Prize in Physics presents a novel study of one of today's most precious commodities—scientific productivity. The author not only measures the variations that exist between different research workers, he also explains these differences and draws some specific conclusions about the relationship of salary to productivity. PROCEEDINGS readers will find this an especially timely and significant discussion, particularly in view of the present widespread concern about manpower shortages and proper utilization of scientific personnel.—*The Editor*

***Summary*—It is well-known that some workers in scientific research laboratories are enormously more creative than others. If the number of scientific publications is used as a measure of productivity, it is found that some individuals create new science at a rate at least fifty times greater than others. Thus differences in rates of scientific production are much bigger than differences in the rates of performing simpler acts, such as the rate of running the mile, or the number of words a man can speak per minute.**

**On the basis of statistical studies of rates of publication, it is found that it is more appropriate to consider not simply the rate of publication but its logarithm. The logarithm appears to have a normal distribution over the population of typical research laboratories. The existence of a "log-normal distribution" suggests that the logarithm of the rate of production is a manifestation of some fairly fundamental mental attribute. The great variation in rate of production from one individual to another can be explained on the basis of simplified models of the mental processes concerned. The common feature in the models is that a large number of factors are involved so that small changes in each, all in the same direction, may result in a very large change in output. For example, the number of ideas a scientist can bring into awareness at one time may control his ability to make an invention and his rate of invention may increase very rapidly with this number.**

**A study of the relationship of salary to productivity shows that rewards do not keep pace with increasing production. To win a 10 per cent raise a research worker must increase his output between 30 and 50 per cent. This fact may account for the difficulty of obtaining efficient operation in many government laboratories in which top pay is low compared to industry with the result that very few highly creative individuals are retained.**

## I. INTRODUCTION

EVERYONE who has been associated with scientific research knows that between one research worker and another there are very large differences in the rate of production of new scientific material. Scientific productivity is difficult to study quantitatively, however, and relatively little has been established about its statistics. In this article, the measure of scientific production I have used is the number of publications that an individual has made.

The use of the number of publications as a measure of production requires some justification. Most scientists know individuals who publish large numbers of trivial findings as rapidly as possible. Conversely, a few outstanding contributors publish very little. The existence of such wide variations tends to raise a doubt about the appropriateness of quantity of publication as a measure of true scientific productivity. Actually, studies quoted below demonstrate a surprisingly close correlation between quantity of scientific production and the achievement of eminence as a contributor to the scientific field.

The relationship between quantity of production and scientific recognition has been studied recently by Dennis,[1] who considered a number of scientists who have been recognized as outstanding. As a criterion of eminence for American scientists, he has used election to the National Academy of Sciences; his study is based on 71 members of the National Academy of Sciences who lived to an age of 70 or greater and whose biographies are contained in the Biographical Memoirs of the Academy. He finds that all of these people have been substantial contributors to literature with the range of publications extending from 768 to 27, the median value being 145. (Based on a productive life of approximately 30 years, this corresponds to an average rate of publication of about 5 per year, a number to which I shall refer in later parts of this discussion.) Dennis concludes that relatively high numbers of publications are characteristic of members of the National Academy of Sciences. He conjectures that of those who have achieved the lesser eminence of being listed in American Men of Science, only about 10 per cent will have a

* Original manuscript received by the IRE, December 3, 1956. Presented first as the invited lecture, Operations Res. Soc. of Amer., Washington, D. C., November 19, 1954; also at the Washington Phil. Soc., late spring, 1955; and at the 1955 fall meeting of the Natl. Acad. of Science. It has been reported briefly in *Newsweek*, December 6, 1954; *Chem. Week*, November 26, 1955; abstracted in *Science*, December 10, 1955; and in *Science Digest;* February, 1955.

† Shockley Semiconductor Lab. of Beckman Instr., Inc., Mountain View, Calif. This material was prepared while the author was Deputy Director and Res. Director of the Weapons Systems Evaluation Group, Dept. of Defense, on leave from Bell Telephone Labs., Inc.

[1] Wayne Dennis, "Bibliography of eminent scientists," *Sci. Monthly*, vol. 79, pp. 180–183; September, 1954.

publication record exceeding the 27; which represents the minimum publisher of the 71 listed in Biographical Memoirs of the National Academy of Sciences. He has also studied eminent European scientists and comes to essentially the same conclusion. In fact his study goes further and shows that almost without exception heavy scientific publishers have also achieved eminence by being listed in the Encyclopedia Britannica or in histories of important developments of the sciences to which they contributed.

It should be remarked that in Dennis' work, he includes more routine types of contributions (such as popular articles) than are generally associated with scientific eminence. However, it may still be appropriate to quote a few of the statistics obtained by Dennis for people who certainly classify in the genius class of the scientific publishers. Among these Dennis refers to: Pasteur with 172 publications, Faraday with 161, Poisson with 158, Agassiz with 153, Gay-Lussac with 134, Gauss with 123, Kelvin with 114, Maxwell with 90, Joule with 89, Davy with 86, Helmholtz with 86, Lyell with 76, Hamilton with 71, Darwin with 61, and Riemann with 19. Riemann, who was the least productive, died at the age of 40. At his rate of publication, he would probably have contributed at least another 10 or 20 publications had he lived to the age of 70. Even with 19, he was in the top 25 per cent of the 19th century scientists referred to in Dennis' study.

The chief conclusion reached in this article is that in any large and reasonably homogeneous laboratory, such as, for example, the Los Alamos Scientific Laboratory and the research staff of the Brookhaven National Laboratory, which are included in this study, there are great variations in the output of publication between one individual and another. The most straightforward way to study these variations is to list the number of individuals with zero, one, two, etc., numbers of publications in the period studied. This compilation may then be plotted as a distribution graph [see Fig. 2(b) for an example]. In some cases, however, the data are too meager for a smooth trend to be seen easily and another form of presenting the data is more convenient.

The form used for most of the data presented in this paper is the *cumulative distribution graph.*

Such a graph can be illustrated in terms of the distribution of the height of a regiment of men. If the men are lined up in order of increasing height at a uniform spacing, then, as shown in Fig. 1(a), there will be a steady increase in height from the shortest man to the tallest man. There will usually be a few men who are exceptionally short, a few men who are exceptionally tall. For the majority of the men the height will vary relatively uniformly along the line of the men. In general, one should thus expect an S-shaped curve with an inflection point near the middle of the distribution.

Such a curve is closely related to the distribution in height shown in Fig. 1(b), which represents the number of men whose height lies in any particular interval of height. This can be obtained from Fig. 1(a), as is represented there, by drawing two lines bracketing a certain interval in height and counting the number of men lying in this range. Fig. 1(b) represents a smooth curve drawn through such a distribution. It can, in fact, be obtained from Fig. 1(a) by drawing a smooth curve through the distribution in height and differentiating the number of men as a function of the height.

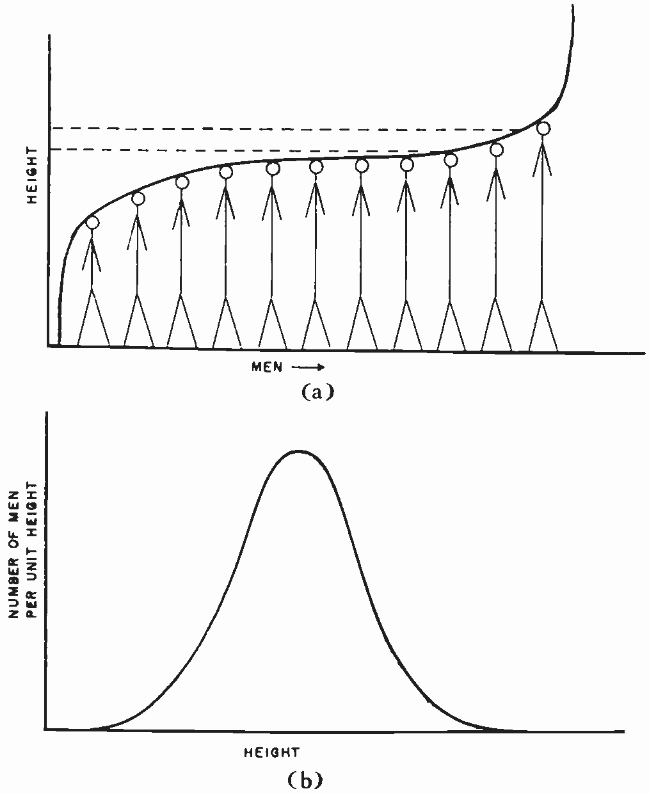



Fig. 1—The cumulative-distribution graph and the normal-distribution curve. (a) The cumulative-distribution graph represented by men arranged in order of height at uniform spacing. (b) A "smoothed" distribution curve, of normal form, such as might be obtained from (a) by finding the number of men in each small increment of height.

For many natural phenomena and in particular for those in which the measured quantity varies due to the additive effects of a large number of independently varying factors of comparable importance, a Gaussian or normal distribution, like that of Fig. 1(b), is obtained. Conversely, if distribution is normal, then the cumulative distribution graph will have the symmetrical S-shaped characteristic in Fig. 1(a), the middle flat portion corresponding to large numbers of cases in the central range, and the rapid convergence of the extremes to their asymptotes corresponding to the scarcity of cases which deviate much from the mean value.

One of the new results of this study, presented below in Section IV, is that the data on rates of publication can be well represented by a normal distribution when treated in a certain fashion. Some possible explanations for this observation are discussed in Section VI.

## II. A Study of Publication Records

As a first example, I shall discuss the statistics of the publications of a group of people in the Los Alamos Scientific Laboratory. This sample of approximately 160 people was selected on the basis that the individuals were professionally mature and located in laboratories whose activities are of such a nature that there is some probability that workers in them might contribute to a physical or electrical engineering publication. Such publications are abstracted in *Science Abstracts A* and *B*, respectively. The publication record for each individual was ascertained by looking through the author index of *Science Abstracts* for the years 1950 to 1953, inclusive.

From these data, a cumulative-distribution graph constructed like that shown in Fig. 1(a) is obtained by listing the men in order of their publications. It is found that approximately half of the individuals have no publications at all. Then there are about 30 individuals with one publication, 20 individuals with two publications and so on. The cumulative-distribution curve shown in Fig. 2(a) has little resemblance to the simple S-shaped curve shown in Fig. 1(a). For one thing it is concave upwards throughout. For another it shows too many individuals with publication rates higher than seven in four years compared to the shape of the curve up to that rate. The distribution curve, shown in Fig. 2(b), is not normal, but instead is essentially hyperbolic in form.

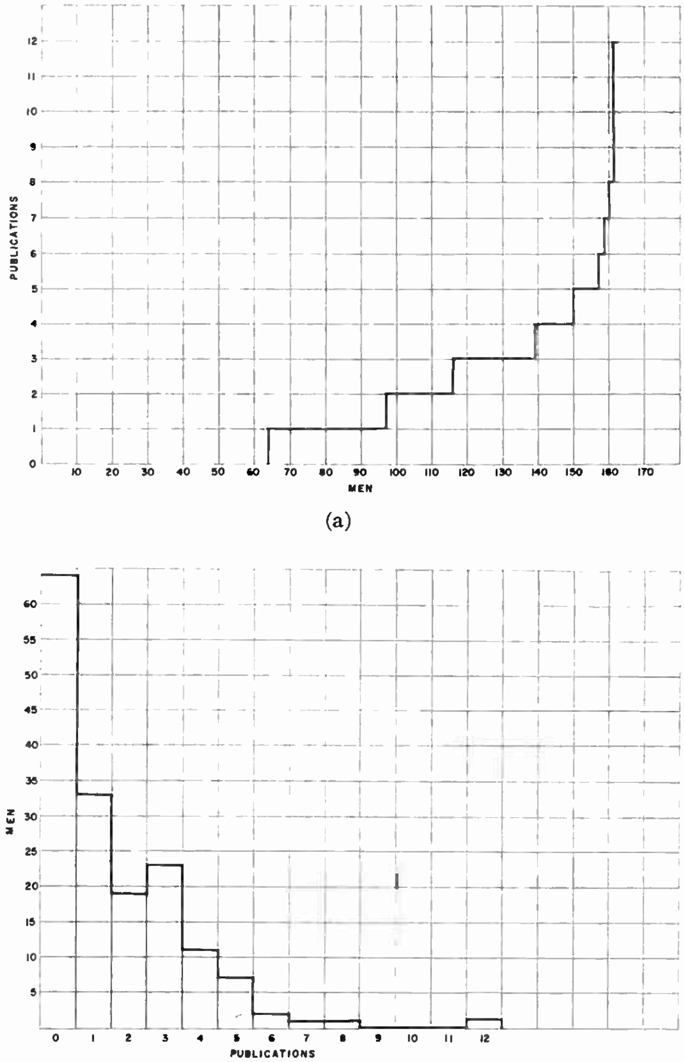
Replotting this same data in Fig. 3 on a logarithmic scale for the number of publications results in a line which does look much more like a portion of the cumulative-distribution graph for a normal distribution. The line is not a smooth curve, of course, but rises in steps. However, a smooth curve drawn through the steps has an approximately linear portion, corresponding to the linear portion of Fig. 1(a), followed by an abrupt turn up at the high end corresponding to the relatively small number of people who on the logarithmic scale have exceptionally large rates of publication.

It is one of the chief conclusions of this study that the more or less normal distribution of the logarithm of rate of publication is characteristic of the statistics of the scientific creative process. Perhaps the most important feature of this conclusion is that the rate of publication increases approximately exponentially from individual to individual, taken in order of increasing rate, and that the differences in rate between low and high producers are very large. The conclusion that the exponential character of the distribution is fundamental to the creative process gains support from the fact that certain other hypotheses intended to explain it as some sort of artifact can be examined and rejected.

In subsequent sections we shall refer to the normal distribution of the logarithm as *log-normal distribution*.

## III. Some Basic Data on Rates of Publication

One of the first hypotheses called the "organization hypotheses" put forward to explain how the log-normal




(a)

(b)

Fig. 2—Distribution of rate of publication (number of entries in *Science Abstracts A and B* in four years) at Los Alamos. (a) Cumulative distribution. (b) Distribution (number of men with each rate of publication).

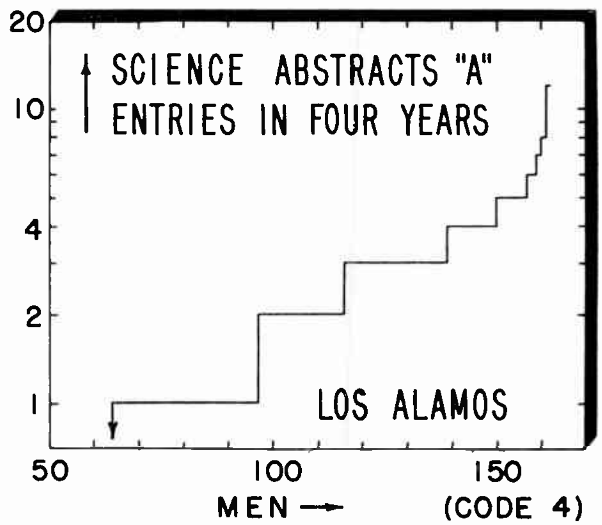



Fig. 3—Cumulative distribution on logarithmic scale for number of publications at Los Alamos.

distribution arises was that it is a consequence of the organization of research activities in large, modern laboratories. In such laboratories, physical scientists frequently make use of very complicated apparatus and large nuclear machines. As a result of this collaborative effort they frequently publish papers jointly, the number of authors varying from two to five or six in ordinary cases. The "organization hypothesis" endeavors to use joint authorship to explain the exponential character as follows: As a consequence of the size of the teams who work together, an individual who has some supervisory or organizational responsibility may contribute to the activities of many men and be listed as a co-author on many papers. As a result, a relatively few people will appear as co-authors of a very large number of papers and this group can be better included in a log-normal distribution than in a normal distribution.

This "organizational hypothesis" can be disposed of by several arguments, some of which are quite instructive. One of these arguments is based on the observation that the exponential aspects of the cumulative-distribution graph is independent of the particular organizational features of the laboratory considered and is a general characteristic of all laboratories studied in this article. For example, the organizational situation in some of the laboratories of the National Bureau of Standards would not lead to large numbers of publications by supervisors. For one Division of the National Bureau of Standards, records were available of the total number of publications and patents made by the individuals in this Division during a period of several years. These data are shown in Fig. 4. It is seen from this figure that the data lie on a relatively smooth exponentially increasing trend followed by a rapid turn-up corresponding again to a few individuals with exceptionally high publication records. Since the organization of activities is quite different in the Bureau of Standards from what it is at Los Alamos while the distribution curve is the same, the "organizational hypothesis" can be discarded.

The "organizational hypothesis" can also be rejected by studying the effect of joint authorship on the distribution of rate of publication. We shall illustrate this argument using data from the Brookhaven National Laboratory. There are approximately 180 members of the research staff of the Brookhaven National Laboratory. The "total" number of entries plotted as a cumulative distribution for these people is shown as the line marked "total" in Fig. 5. Since Brookhaven operates in a fashion rather similar to Los Alamos, it might be expected that the "organizational hypothesis" would apply equally well here. In order to test this, two other lines have been constructed on Fig. 5.

The bottom line, marked "solo," has been obtained by discarding all publications having more than one author. It is seen that a relatively small fraction of the people have made "solo" publications. However, it should be noted that the most prolific publishers of these

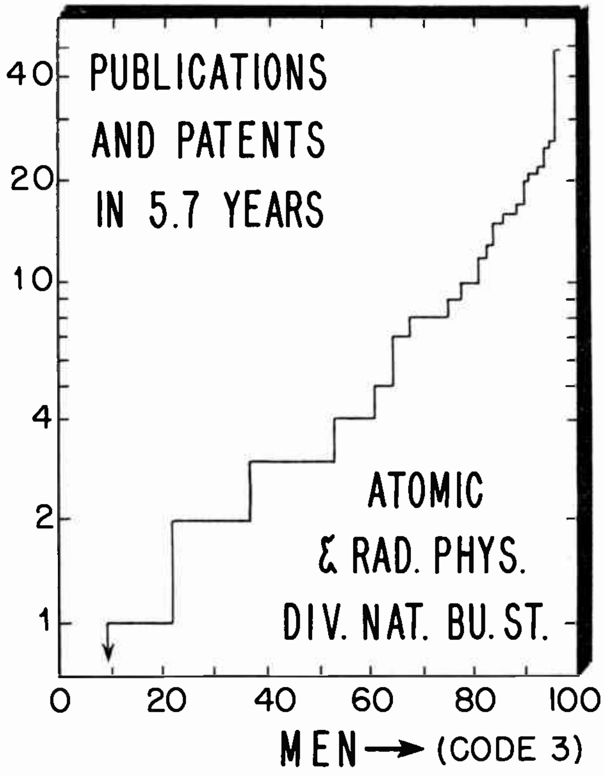



Fig. 4—Cumulative distribution on logarithmic scale for publications and patents for Atomic and Radiation Physics Div., National Bureau of Standards, for a period of 5.7 years.

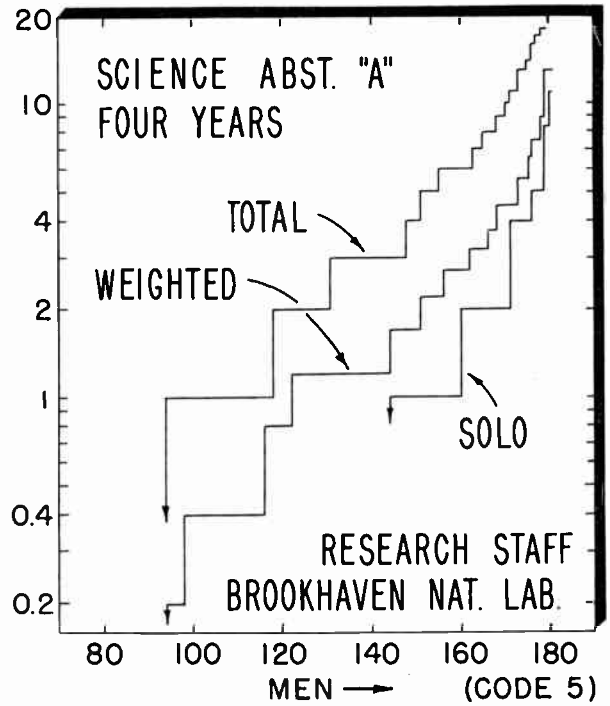



Fig. 5—Cumulative distributions on logarithmic scale for 3 cases at Brookhaven National Lab.

have published at nearly half the maximum rate for the "total" line. On the other hand, a large number of people who appear as co-authors in the "total" distribution have no "solo" publications whatever. This fact shows that the rapidly rising part of the line is due largely to people who are capable of producing "solo" publications, a conclusion contrary to the expectation based on the "organizational hypothesis." In fact, the evidence is that publication of about half of the people is supported by the more productive ones who would be capable of publishing at relatively high rates strictly on their own.

The middle line marked "weighted" is obtained by dividing the credit for multiple-author publications equally among the various authors. For example, each man on a four-author publication receives a contribution of 0.25 publication. The "weighted" line again shows the steadily increasing trend and does not permit an undue credit to be given to people who, through organizational position, may appear as a joint author on a large number of publications. This furnishes further support for the thesis that the exponential trend of the cumulative distribution is a fundamental characteristic of the distribution of productivity among the members of the laboratory rather than some organizational artifact.

Another possible explanation which can also be discarded is that the distribution of degree of publication from one person to another is a consequence of the distribution in age of the population considered. In principle, some such distribution might be obtained as a result of distribution in age since people on the average have a maximum in their publication rate at an age of about 35. The distribution of publication in age has been studied by Lehman.[2] Some of Lehman's results for rate of publication as a function of age are shown in Fig. 6. Very similar results are obtained for other geographical samples. Actually, what Lehman has studied is not simply publication record but "creative production." He judges creative production by references found in histories of science and other similar sources. Since the distribution of workers in the laboratories considered in this study shows a fairly uniform distribution from age 25 to age 50, it is difficult to see how the variation in productivity with age as shown in Fig. 6 could result in a very small fraction of people with exceptionally high publication rates: from Fig. 6, we would estimate that the maximum publication rate would be perhaps twice the publication rate of the median man. In contrast to this, the studies shown for Figs. 2, 3, and 4 correspond to maximum publication rates substantially more than ten times that of the median man.

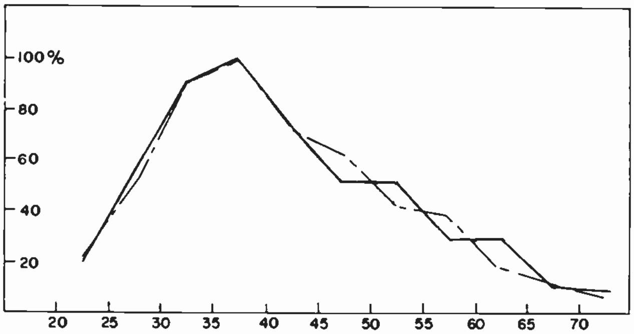



Fig. 6—Creative production rate in science and mathematics vs age for (solid line) nationals of 14 different countries other than Russia, England, France, Italy, Germany, and the U.S.A. and for (broken line) U.S.A.

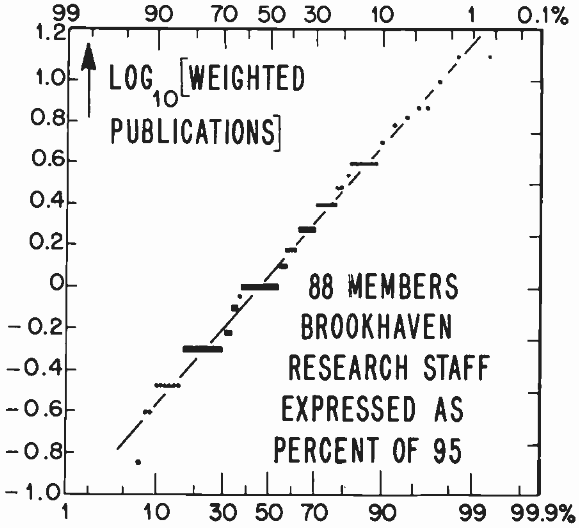



Fig. 7—Cumulative distribution of logarithm of "weighted" rate of publication at Brookhaven National Lab. plotted on probability paper.

## IV. The Log-Normal Distribution of the Rate of Publication

The conclusion is thus reached that the exponential variation of productivity in the cumulative distribution graph is a characteristic feature of the statistics of productivity in a research laboratory. This conclusion receives further support from an additional analysis of the data which show that the logarithm of the rate of publication can be well represented as a normal distribution in the cases studied.

The validity of the assumption of a normal distribution can be tested by making use of so-called "probability paper." On such paper, the cumulative number of men is expressed on a percentage scale. This percentage scale is so distorted as to increase the spread on the scale at percentages near the extreme distribution. This results in stretching out the ends of the cumulative-distribution graph of Fig. 1(a) so that it becomes a straight line, provided the distribution itself is normal.

Such a test has been applied to the weighted rate of publication for the Brookhaven Laboratory shown in Fig. 5. The result is shown in Fig. 7. It is seen that a straight line can be drawn in a very satisfactory way through the data with the exception of the two extreme men. It should be noted that in many cases so many men were assigned the same publication number that they have been represented as solid blocks on the diagram rather than as individual points. This grouping together is a genuine effect in the case of people who

[2] H. G. Lehman, "Men's creative production rate at different ages and in different countries," *Sci. Monthly*, vol. 78, pp. 321–326; May, 1954.

published one "solo" publication during the period studied and thus have a logarithm of zero and those who have appeared on two publications or as a co-author of a single two-author publication and appear at logarithms of 0.3 and −0.3. Some of the other groupings have resulted artificially from the means of handling the statistics: for simplicity in listing the people, the scale of possible publications was divided into intervals and those whose publication rates fell in these intervals were grouped together. If this had not been done, the data would fall more closely along a straight line, *i.e.*, the "fit" to the normal distribution would be better.

Fig. 7 illustrates strikingly the range of variation in rate of publication—a factor of 40-fold between lowest 10 per cent and highest 5 per cent.

The fit shown on Fig. 7 is based on the assumption that the research staff of Brookhaven may be divided into two parts, one part containing 95 members who have some likelihood of publishing physics papers referenced in *Science Abstracts A* and 85 others with negligible likelihood of making such publications. The number 95 was found by trial and error to give the best straight line in Fig. 7. This arbitrary procedure does have justification in terms of the distribution of activities in the Brookhaven Research Staff. In fact if the list of members of the Research Staff at Brookhaven is examined name by name, it is found that many are biologists, medical physicists, and the like whose fields are not covered by *Science Abstracts*. The final conclusion is that all but 101 names are considered extremely unlikely to make publications abstracted in *Science Abstracts A*. Since the difference between 101 and 95 is negligible in respect to other uncertainties in the study, we may conclude that for the publishing part of the population the rate of publication is well represented by a normal distribution on the logarithmic scale, or for brevity, a *log-normal* distribution.

Generally similar fits are obtained for the Los Alamos data and for the National Bureau of Standards data. Furthermore, the data on "total" and "solo" entries in *Science Abstracts A* can be fairly well fitted by log-normal distributions. The fit is very "jumpy," however, since the only possible values for publication rates are integers. On the basis of the rather limited investigation that I have carried out to date in regard to the distributions for "solo" and "total" rates of publication, it appears that these also have log-normal distributions except that the rates of publication differ from the "weighted" rates by factors of 0.6 and 1.6, respectively.

It would be interesting to compare the statistics of science departments in universities with those of the large laboratories studied above. This has not yet been done except for the limited data on the Physics Department of Columbia University shown in Fig. 8. In spite of the smallness of the sample, the general trend of the data is such as to give confidence that the log-normal distribution will also hold in such cases.

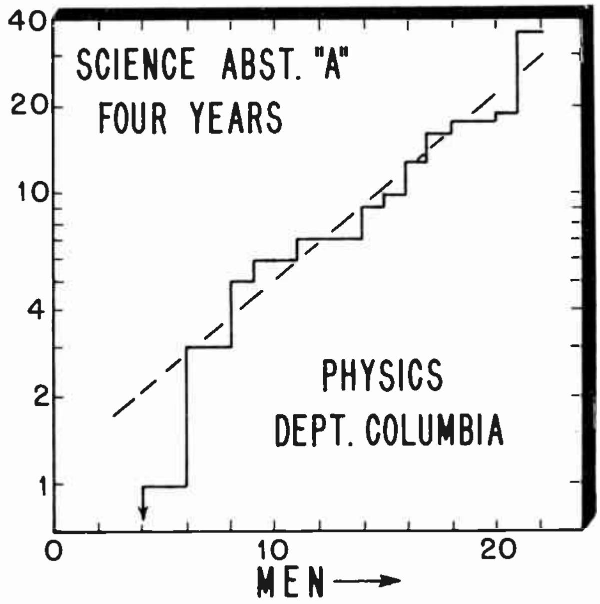


Fig. 8—Cumulative distribution on logarithmic scale for publications of the Physics Dept., Columbia Univ., for 4 years.

## V. A Study of Patent Activity

Another measure of creative technical production, which is relatively readily available for study, is patent activity. Shown in Figs. 9 and 10 (opposite) are cumulative-distribution curves for patents for two large laboratories in the fields of electrical apparatus and communications. All of the data correspond essentially to "solo" publications since the number of joint patents is very small compared to individual patents.

It is instructive to compare patents with publications. Such a comparison is presented in Fig. 11 for a selected group of 60 men from one of the laboratories considered in Figs. 9 and 10. The most significant factor to note is that on the logarithmic scale, the patent distribution is markedly steeper.

## VI. Speculations on the Origin of the Log-Normal Distribution

The very large spreads in productivity, for example the variation by nearly one hundred fold between extreme individuals in Fig. 7, are provocative of speculation. Most rates of human activity vary over much narrower limits, for example, pulse rates outside the two to one range from 50 to 100 per minute are extremely rare. Very few individuals walk at speeds outside the range of 2 to 5 miles per hour. In competitive activities involving trained and selected people, such as running the mile, the variation is much smaller, the ratio of speed for the mile between world's record and good high school performance being probably less than 1.5.

In the study presented here the individuals are presumably specially selected by natural ability and specially trained to accomplish scientific production. Yet the spread in rates is enormously greater than it is for the more physical activities discussed above. I believe

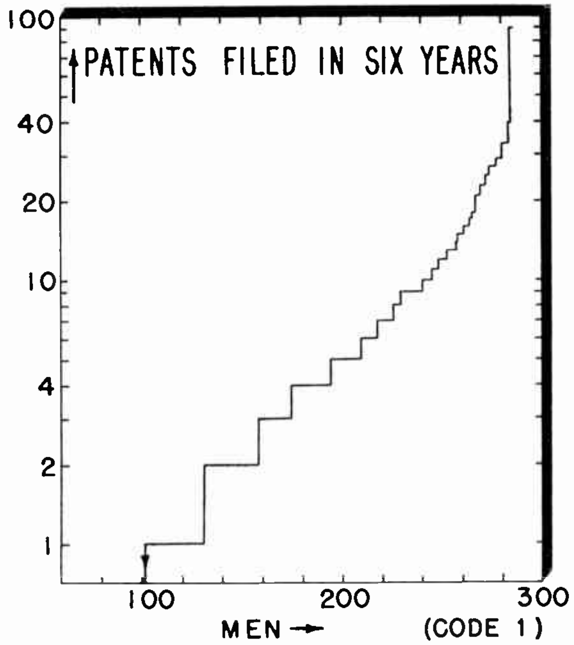



Fig. 9—Cumulative distribution on logarithmic scale for patents at a large industrial laboratory.

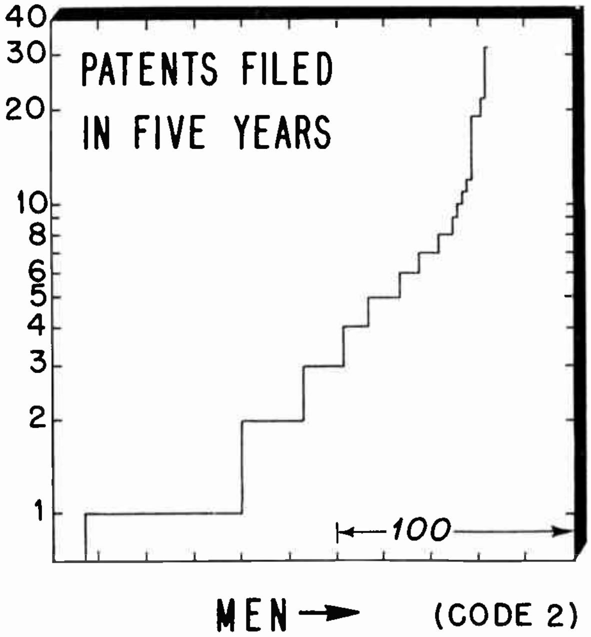



Fig. 10—Cumulative distribution on logarithmic scale for patents at another large industrial laboratory.

that it is possible to explain to some degree how such large variations in rate may occur in terms of certain characteristics of the creative scientific process. The basis of the explanation is that the large changes in rate of production may be explained in terms of much smaller changes in certain attributes. I shall illustrate this in terms of a simplified example of the inventing process.

In order to make an invention for which the United States Patent Office will issue a patent, it is, in general, necessary to conceive a new combination of features and

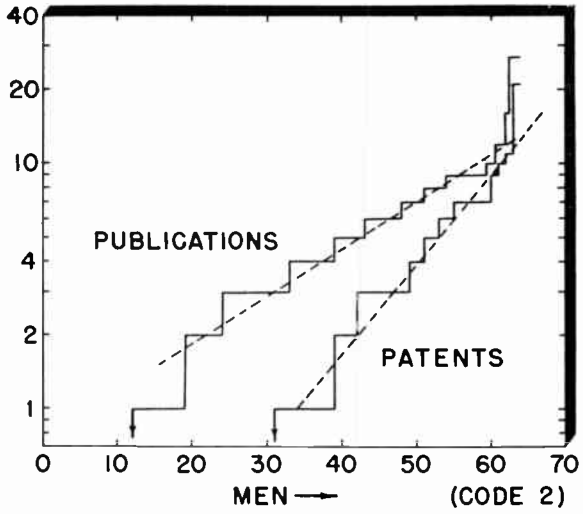



Fig. 11—Comparison of patent and publication activity for a group of research workers at a large industrial laboratory.

to appreciate how this combination may be useful. Let us suppose that the inventor perceives that he has made an invention when he appreciates the relationship between some number of ideas. For example, the automobile self-starter might have been conceived by recognizing the relationship of the following 4 ideas: the idea that a means of starting the engine without using human, muscular strength would be useful, the idea that the necessary energy could be held in reserve in a storage battery, the idea that a relatively small high speed electric motor could be used to turn the larger gasoline engine at starting speed, and the idea that the electric motor could be subsequently disengaged in order to avoid rotating it at excessive speeds.

Now let us suppose that there is some attribute of the human brain which allows an individual to be aware of "$m$" ideas and their relationships.[3] Then it follows that a man with $m=3$ will never invent the self-starter in the form discussed above whereas a man with $m=4$ can do so. A man with a higher value of $m$ is much more likely to make the invention than a man with $m=4$. In fact, it may be established, by use of the formulas for permutations and combinations, that men with $m=5$, 6, and 7 can hold the 4 essential ideas in awareness (together with 1, 2, or 3 irrelevant ideas) in 5, 15, and 30 times as many ways as the man with $m=4$. This shows that a variation of 50 per cent in "brain capacity" ($m=4$ to $m=6$) can produce an increase in invention rate of 15-fold for inventions requiring the interaction of 4 ideas.

It may be instructive to illustrate the considerations presented above by an example which can be shown in detail. Suppose out of realm of idea associated with some field of endeavor an invention can be made by

[3] N. Rashevsky, "Mathematical Biophysics," University of Chicago Press, Chicago, Ill., ch. 29; 1938, presents very similar reasoning. His results are expressed in the form of equations rather than by numerical examples and lead to somewhat more general conclusions than those presented here.

holding ideas "1" and "2" in mind and seeing the relationship between them. Then a man with $m=2$ can make the invention in two ways as represented below:

$$(1, 2) \text{ and } (2, 1).$$

But a man with $m=3$ can think of these two ideas and some irrelevant idea $x$ in six ways:

$$(1, 2, x),\ (2, 1, x),\ (1, x, 2),\ (2, x, 1),\ (x, 1, 2),\ (x, 2, 1).$$

Thus for every case in which the $m=2$ man can think of the idea, there are 3 ways in which the $m=3$ man can do it. Thus the $m=3$ man has 3 times as many chances to make the invention.

Evidently this advantage increases rapidly with the increasing complexity of the problem. For a 10-idea invention an 11-idea man has an 11-fold advantage over a 10-idea man; that is a 10 per cent increase in "mental capacity" produces a 1100 per cent increase in output. It is my impression that this sensitivity to the interaction of many factors in mental creativity is the key to the large variations in output found in this study. According to this explanation, the log-normal distribution in productivity then results from a normal distribution, over a relatively small range (say $m=8$ to $m=12$ in the model considered), of some attribute which controls productivity in a very sensitive way.

Still another way of rationalizing the log-normal distribution may be based upon the hypothesis that the interacting mental factors are of several different kinds rather than several of one kind, as in the case of several ideas as discussed above. For example, consider the factors that may be involved in publishing a scientific paper. A partial listing, not in order of importance, might be: 1) ability to think of a good problem, 2) ability to work on it, 3) ability to recognize a worthwhile result, 4) ability to make a decision as to when to stop and write up the results, 5) ability to write adequately, 6) ability to profit constructively from criticism, 7) determination to submit the paper to a journal, 8) persistence in making changes (if necessary as a result of journal action). To some approximation, the probability that a worker will produce a paper in a given period of time will be the product of a set of factors $F_1$, $F_2$, etc. related to the personal attributes discussed above. The productivity of the individual would then be given by a formula such as

$$P = F_1F_2F_3F_4F_5F_6F_7F_8. \tag{1}$$

Now if one man exceeds another by 50 per cent in each one of the eight factors, his productivity will be larger by a factor of 25. On the basis of this reasoning we see that relatively small variation of specific attributes can again produce the large variation in productivity.

The factor explanation discussed above also has an appeal from the point of view of the log-normal distribution. According to the formula, the logarithm of the product is the sum of the logarithms of the several factors. If we suppose that these factors vary independently, then to a good approximation their sum will have a normal distribution, and so, consequently, will the logarithm of the productivity. It seems to me that this is at present the most attractive explanation for the log-normal distribution.

In closing this section mention should be made of an attempt to fit the data by assigning to each individual a single parameter describing his creative potential. This parameter was referred to as "mental temperature" when the original lecture was given. It was introduced in analogy with the quantity $\beta$ or $1/kT$ which occurs in the equation for rates of chemical reaction or thermionic emission. According to this hypothesis an individual $i$ is characterized by a value $\beta_i$. In a situation $s$ his rate of production is determined by a rate constant $P_s$ and a barrier $U_s$, so that his rate of production is

$$P(i, s) = P_s \exp(-U_s\beta_i). \tag{2}$$

The rate constant $P_s$ probably depends on $\beta_i$ but in a relatively insensitive way, so that to a first approximation this dependence can be neglected.

On the basis of this equation, the difference between the two curves of Fig. 11 is to be attributed to a $U$ value 1.7 higher for patents than for publications.

There appears to be a tantalizing possibility of establishing scales for $U$ and $\beta$ by comparing publications and patents and one laboratory with another. One might, for example, assume that the distribution of $\beta$ values is the same in two laboratories having the same pay scales and similar working conditions. Then if $U$ is chosen as unity for one activity in one of these, the scale of $U$ can be chosen for the other cases in terms of the ratio of slopes like those of Fig. 11. Approximate values of $P_s$ can be chosen by assuming that $\beta=0$ represents a situation in which the worker never lacks an idea to publish or an invention to patent so that his rate of production is limited by the mechanics of the situation. Such cases might correspond approximately to the most outstanding publishers in Dennis' study. On this basis $P_s$ values of the order of 10 per year for either publications or patents might be chosen. I have made some attempts to establish scales of this sort but they are not well enough developed to warrant inclusion here.

## VII. The Relationship Between Salary and Productivity

From the point of view of the economics of running a research laboratory, it is important to know the relationship between salary and productivity. For example, if the better paid men are more productive than their fellows in greater proportion than the increase in pay, then they are a sound investment. On the other hand, if they are less productive per salary dollar, then it may be wiser to hire relatively fewer of these outstanding people.

The question just posed is to some degree academic—anyone who has had to do with managing research knows that progress depends largely on a relatively small number of exceptionally able individuals. He also knows that these people are usually substantially better paid than their fellows. How much better one can afford to pay outstanding people and still find them profitable is a quantitative question faced by many organizations during periods of rapid build up. The findings in this section throw some light on this question, the conclusion being that, in general, scientific productivity is so much greater for the outstanding people that in the current scientific labor market, it is unlikely that they will be overpaid.

It is clear, of course, that increasing salary of an individual will usually not increase his productivity much, if at all. In some cases it may even have the opposite effect by reducing incentive. What is studied here is the statistical relationship between salary and productivity as established by existing pay roll procedures. If any causal relationship is important in this connection, it is that high productivity of an individual causes the management to give him high rewards.

Before considering the method of investigating the statistical relationship between salary and productivity, it may be worth-while to say something about salary in general. In determining the salary of an individual in a research laboratory, the management takes into account many factors. Only one of these factors is considered in the previous parts of this study, namely, the rate of scientific production as measured by total numbers of publications or patents. This factor is probably rarely considered in a quantitative way. Instead, the usual procedure is for a group of people charged with supervising research workers to gather together and discuss the relative merit of the individuals. In such considerations, quantitative measures of the individual's contributions are seldom referred to. There probably does not exist at the present time any valid analysis of the various factors that are considered and their relative importance. Among them may be mentioned, however, the originality and importance of publications which are made. Thus quality as well as quantity is brought into account. Other factors which are certainly considered are the ability of an individual to carry out the techniques of his work, whether these be of a theoretical nature involving pencil and paper or the manipulation of apparatus; the ability to contribute to the solutions of problems of other workers in the organization; the ability to produce cooperation among other workers; the ability to attract productive candidates to the organization; the ability to influence the activities of other workers along lines which are more wisely chosen than they would choose themselves with respect to the goals of the organization as a whole; the ability to carry out activities which enhance the prestige of the organization. These and many other factors are generally considered in determining a man's "merit" and thus deciding what salary he should receive.

The assumption of this article is that merit and salary are somehow determined by the combination of such factors as those which we have discussed above. These factors are not closely correlated with each other, although it is probable that there is a tendency for outstanding ability in any one to be coupled with a probability of higher abilities in the others as well. The only attribute which has been studied here is simple quantitative productivity in the sense of publications and patents. If it is found that this attribute, which was studied purely for purposes of convenience, is strongly correlated with increasing salary, then it seems likely that the other attributes are also strongly correlated with salary.

It is not appropriate to consider simply the relationship of salary to productivity. The reason for this is that there is a general tendency of salary to increase with age, this being a recognition of increasing general judgment and experience with age as well as a socially acceptable procedure. Thus, in order to get a truly representative comparison of merit with productivity, it is necessary to correct for age. This procedure can be done in various ways; the one selected for this article being that associated with the concept of "merit quartiles."

The division of the population of a laboratory into "merit quartiles" may be illustrated with the aid of Fig. 12. This figure represents the salaries of a group of

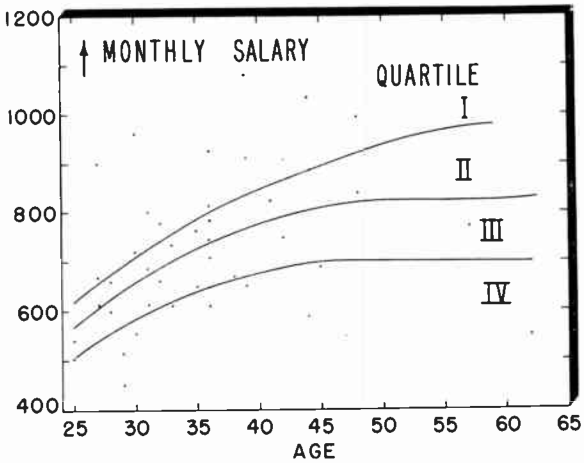



Fig. 12—Salary vs age (for a representative sample only of individuals) in a laboratory considered in this study with lines dividing the distribution into "merit quartiles"). Effective about October, 1954.)

individuals in a laboratory covered in this study. Each individual is represented by a point on the figure which shows his salary and his age. Three lines have been drawn on the figure dividing it into four groups of individuals, called quartiles. The procedure for drawing these lines is as follows: in each relatively small age interval the population of the laboratory is divided into halves such that half of the group gets more than the median salary and half less. Then the upper and lower halves are similarly divided into 2 equal parts so that each age interval is divided into 4 quartiles. This procedure is carried out for the various age intervals and then a smooth curve is drawn. These smooth curves are drawn in such a way that at each age interval, approximately $\frac{1}{4}$ of the population of the laboratory lies in quartile I and approximately $\frac{1}{4}$ in each of II, III, and IV. Thus the people in the first or top quartile have approximately the same age distribution as those in the second, etc. Furthermore, all of the people in the top quartile obtain higher salaries than those in the second quartile at the same age.

These merit quartiles furnish a basis for dividing the laboratory into parts in accordance with salary but chosen in such a way that the age distribution in each part is similar. Thus any effect of varying productivity with age affects all the quartiles about the same way.

Fig. 13 shows a similar plot for the individuals in a U. S. government laboratory operated under Civil Service. It is to be noted that the highest salaries at any age range are substantially lower than those in the other non-Civil Service laboratory. The difference would be even more striking if the higher paid executive types of an industrial laboratory could also be shown.

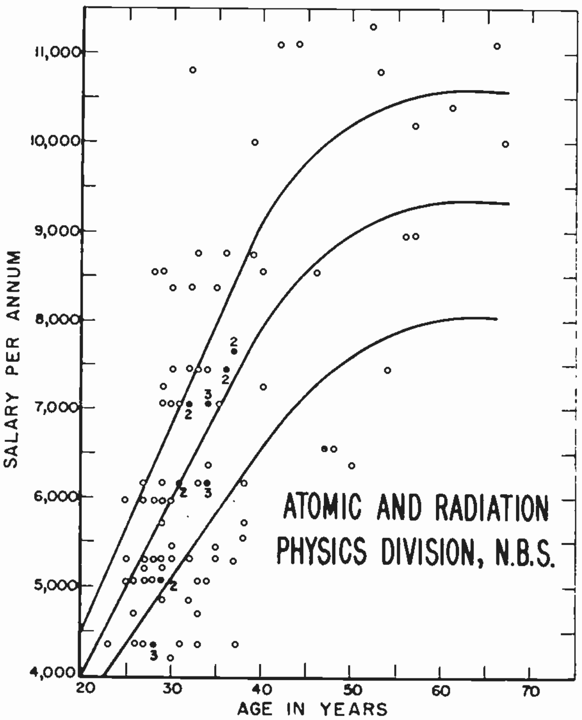



Fig. 13—Salary vs age for Atomic and Radiation Physics Div., National Bureau of Standards, together with "merit quartiles" divisions. (Effective about October, 1954.)

The use of merit quartiles, deciles, or similar divisions is playing a progressively more important role in salary administration.[4] One of the great advantages of the merit scale is that it provides an intuitively satisfactory way of ranking the individuals in an organization. The same would not be true if the men were ranked simply according to salary; thus a very able young man at a relatively low salary would be obviously out of place in company with an older group of average ability (but with more experience) at the same salary and it would be difficult to get any sense of order from such listing. On the other hand, a group of supervisors can come to agreement and reach decisions surprisingly easily about merit rank between people whose ages and salaries may differ by large amounts. I do not believe that it is evident in any *a priori* sense that such agreement would be expected; it appears rather an interesting and useful experimental result. In a sense, it is a surprising result since, as discussed above in this section, such diverse factors are considered in making the judgment. The agreement as to merit ranking by a supervisory group does not, of course, imply that the worth of the individual is truly assessed in any absolute sense. However, the large degree of consistency does imply that a useful and impartial tool for salary administration exists.

In principle, an organization can establish a family of merit curves at each raise period (allowing for cost of living adjustments, changing competition, etc.). The new salary for a man whose merit rank is correctly appraised can then be simply read off his location on the new curves. It sometimes happens, due perhaps to accidents of recruiting or due to changing skills on the part of the worker, that a revision of merit rating occurs. It is generally felt that only a fraction, say 50 per cent, of the correction should be made in any one raise since this will tend to smooth out fluctuations in the salary system.

A set of quartiles like those shown in the two previous figures have been prepared for the research staff of the Brookhaven National Laboratory. For each one of these quartiles, which contain about 46 men each, the publication records have been compiled as cumulative-distributions. These are not presented as graphs with steps since there are so many cases of overlap that the lines for different quartiles are very hard to separate. Consequently, smoothed distribution curves have been drawn through the steps in the manner illustrated in Fig. 1(a). The resulting curves are shown as Fig. 14.

From Fig. 14 it is readily seen that approximately the same numbers of people in quartiles I and II published, but that the amount of publication of the high publishing members of quartile I was larger by almost a factor of 2 than for the corresponding people of quartile II. Quartile III contains some individuals having high rates of publication and a smaller fraction of people publishing. The total amount of publication in quartile IV was substantially less than quartile III.

Similar diagrams have been made for other laboratories but there is no great uniformity in their characteristics. However, there is a very general trend which holds for all cases considered. This trend is for the average rate of publication per individual to increase steadily from quartile to quartile, being highest for the first or best paid quartile.

From the type of spread which is observed in Fig. 14,

[4] Employee interest is also high. For example, merit curves have been deduced from polls of employees of Bell Tel. Labs. by the Conf. of Prof. Tech. Personnel Inc., P.O. Box 625, Summit, N. J.

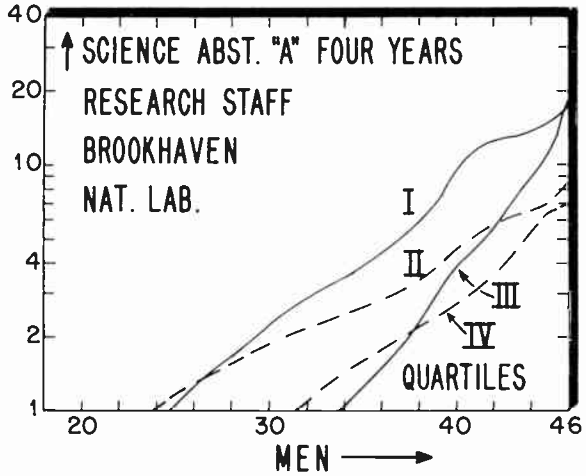



Fig. 14—Cumulative distributions (shown as smooth lines) for the four "merit quartiles" of the research staff of Brookhaven National Lab.

it is evident that publication per se is not given heavy weight in determining merit in terms of salary at Brookhaven. It is evident that something like 10 or 15 per cent of the individuals in quartiles III and IV exceed the publication records of about 50 per cent of the people in quartiles I and II. However, this is not sufficient to give them in terms of salary a recognition equal to those of quartiles I and II. Thus it follows that other factors certainly are being considered in determining salary.

From the general shape of the curves shown in Fig. 14, a very crude sort of an estimate can be made of the number of additional factors which must be taken into account in determining salary provided these factors are assumed to have importance approximately equal to amount of publication. For example, if we compare quartiles I and II we see that only 10 per cent, approximately, in quartile I exceed the maximum production of people in quartile II. This suggests that there might be something of the order of 10 other factors involved in weighting the people of quartile I, each one of these 10 other factors contributing to a group of about 10 per cent who exceed the performance of individuals in quartile II. Evidently this type of reasoning does not apply in the same way to quartiles I and III, but the fact that something between $\frac{1}{10}$ and $\frac{1}{4}$ of quartiles III and IV exceed most of the people in quartile I in terms of amount of publication suggests an analysis might lead to the conclusion that in determining subjectively the merit rating of an individual, salary reviewers act as if there were something like 4 to 10 factors of comparable importance to amount of publication.

I shall now return to the question taken up in the beginning of this section, namely, the quantitative relationship between salary and productivity. For the various laboratories considered in the study, sets of quartiles have been drawn and the average amount of production determined for each quartile. This information is gathered together in Fig. 15. The data have been expressed in terms of rate of activity in publication or patents per man-year. For the publications the total number of publications was used (not "solo" or "weighted"). It is observed that in all cases there is a monotonic increase in rate of activity with quartiles, increasing towards the highest paid quartile, quartile I. The actual spread in amount varies by a factor of about 9 for the most rapidly varying case and by a factor of a little over 3 for the most slowly varying case.

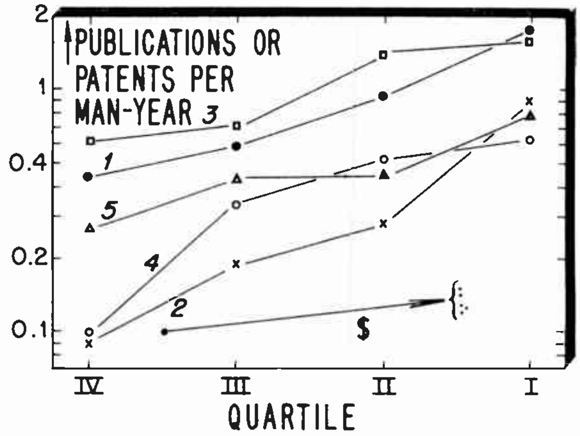



Fig. 15—Relationship between productivity and quartile number and salary and quartile number for several laboratories.

A comparison with salary is also indicated in the figure. The ratio of salary at the dividing line between quartiles III and IV to that between quartiles I and II at age 35 is also shown on the figure. Five cases have been considered and there are somewhat different spreads in salary for these. The line represents a sort of weighted average of the change in salary.

It is clear from inspection of the figure that in progression from quartile to quartile there is much less increase in salary than in productivity, in fact productivity lines rise 3 to 5 times as steeply as the salary lines. In other words, statistically an increase of 30 to 50 per cent in productivity is necessary for an individual to obtain an increase in salary of 10 per cent. However, as the reasoning given in connection with Fig. 14 shows, increase in scientific productivity alone is not sufficient to produce the increase in merit rating. In fact, coupled with the 30 to 50 per cent increase in productivity, there probably must be comparable increases in other kinds of contribution. In other words, the individual probably must become 30 to 50 per cent better in all respects in order to receive a recognition corresponding to a 10 per cent increase in salary.

## VIII. Relevance to Civil Service Salary Scales

I should like next to discuss the relevance of these findings to the problem of Civil Service scientists in government laboratories. In addition to relatively low salaries positions in government laboratories are less attractive than those in industry or in universities. This is especially true in laboratories in the military establishments where periodically changing direction by officers who are not experienced in directing research frequently leads to morale problems. These problems have been thoroughly explored and reported in detail in the recent report[5] of the Riehlman committee of Congress. Clearly it is important to retain in these laboratories some highly-qualified, strong-minded, inspired leadership in order to prevent research effort from becoming thoroughly second grade.

This brings us to the most important conclusion in the study, and one which might possibly furnish a basis for action. The top salaries in government laboratories are substantially lower than both those in industry and in universities, at least for people in the latter whose line of work involves undertaking summer assignments and doing consulting. Even if there were no disadvantages aside from salary, the limits of salary set by Civil Service scales probably have a most severe effect on the leadership and originality available in government laboratories. Although these attributes have not been studied quantitatively, all of the findings in this article are consistent with the idea that leadership and originality increase very rapidly with salary just as do rate of publication and rate of invention. Cutting off the top of the salary scale at, say, $12,000 per year as compared to $18,000, will mean a reduction of productivity of 3 to 8 fold, according to the statistics deduced in connection with Fig. 15. Statistically, for the higher salaried man the return per dollar of salary is two to five times as great so far as individual productivity is concerned. If leadership qualities vary in a way similar to productivity, the return from increased salary will be enormously greater since an effective leader may substantially improve the output of many men.

In closing, I should emphasize that there are outstanding exceptions to most statistical results. Government laboratories do succeed in retaining a few outstanding individuals. These are unfortunately exceptions rather than the rule. Because of the present top limits on Civil Service salaries for scientists, the taxpayer's dollar is buying less research value than it should. A policy of having more highly paid positions might well double the return per dollar. It might also contribute significantly to offsetting the lead which the U.S.S.R. has currently gained in numbers of technical degrees granted in universities per year.

## IX. Acknowledgment

The preparation of this paper has been made possible by the cooperation of a number of individuals and organizations. In particular, I would like to thank Dr. Lauriston S. Taylor and Mrs. Shea Kruegel of the National Bureau of Standards, Dr. L. J. Haworth of Brookhaven National Laboratory, Dr. John K. Herzog of Los Alamos Scientific Laboratory, and Dr. C. H. Townes of Columbia University. I would like also to acknowledge certain anonymous help. The appearance of this article in the Proceedings of the IRE results from a suggestion by E. W. Herold.

[5] Organization and Administration of the Military Res. and Dev. Programs, Twenty-fourth Intermediate Rep. of the Committee on Government Operations; August 4, 1954.

# Correction

J. R. Wait and H. H. Howe, authors of "The Waveguide Mode Theory of VLF Ionospheric Propagation," which appeared on page 95 of the January, 1957, issue of Proceedings of the IRE, have brought the following corrections to the attention of the editors.

In (2), $(h/\lambda)^{1/2}$ should be replaced by $(h/\lambda)$ and $S_n^{3/2}$ should be replaced by $\epsilon_n S_n^{3/2}$ where $\epsilon_0=1$, $\epsilon_n=2(n\neq 0)$.

In (3), $(\lambda/n)$ should be replaced by $(\lambda/h)$.

In Fig. 1, the abscissa labeled $h$ should be $L$.

# Molecular Amplification and Generation of Microwaves*

JAMES P. WITTKE†

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

**Summary—This paper is a general introduction to the field of amplification and generation of microwaves using molecular rather than electronic processes. The basic physical properties of molecular systems as related to amplification are reviewed. The properties of molecular amplifiers, such as gain, bandwidth, saturation power, and noise figure are discussed, and several specific types of amplifiers are described. These include the molecular-beam Maser, the "hot-grid cell," amplifiers excited by rf pulses and by "adiabatic fast passage," and amplifiers based on multilevel molecular internal energy systems, including "optically pumped" amplifiers.**

**Molecular amplifiers may add very little noise to the signal to be amplified: noise figures under one db can be obtained. With suitable feedback, such amplifiers become oscillators of extremely high spectral purity. High gains can be achieved using regeneration, but bandwidths are relatively small. These range from the order of tens of kilocycles for amplifiers using a gaseous molecular system, to megacycles, using solids. Molecular amplifiers saturate at low input powers, of the order of microwatts. Variations of the devices discussed may provide a means of generating millimeter and submillimeter waves.**

## I. Introduction

In familiar microwave devices, such as klystrons, magnetrons, and traveling-wave tubes, dc power is converted to microwave power by the interaction of moving charged particles (electrons) with a microwave field. Microwave fields can also interact with *uncharged* matter (molecules). In such "molecular interaction," there is no flow of charge, with its kinetic or positional energy being transformed to rf energy, but instead the *internal* energy of molecules is *directly* converted to microwave energy. This is a familiar process at very short wavelengths, for example, emission of visible light from a gas discharge. This light represents the direct transformation of internal electronic excitation energy of *neutral* atoms to electromagnetic radiation energy. However, there is a difference of profound importance between such optical radiation and the microwave radiation of interest in this paper. Atoms excited in a discharge tube have a very strong tendency to radiate light spontaneously. The presence of similar radiating atoms nearby has negligible effect on the emission, and the atoms emit independently of each other in an *incoherent* fashion. In the microwave frequency region such spontaneous emission of internal energy as radiation is relatively unimportant. Molecules must be *induced* to give up their internal energy by locating them in a microwave field that either is generated by neighboring radiating molecules or is externally applied; this induced emission is *coherent*, that is, phase-related to the general radiation field.

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The motions of electrons in atoms and molecules are such that the associated energies form a discrete set, that is, only those motions leading to certain *allowed* internal energies can exist. Thus, the internal energy associated with the motion of electrons bound in molecules is quantized; an electron can only exist in certain *energy states*. It has become conventional to indicate the scheme of energy states for a molecule by a diagram such as Fig. 1. From this representation the term *energy level* has arisen, and is used interchangeably with *energy state*.

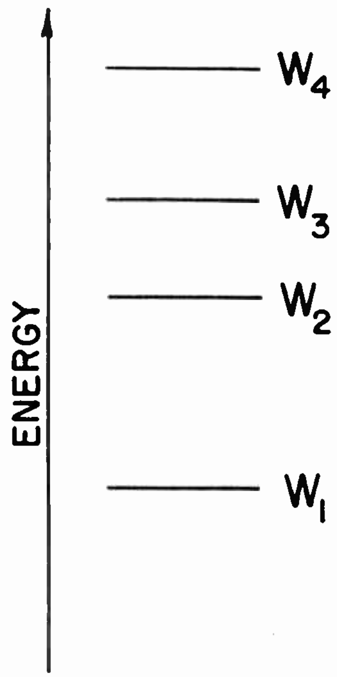


Fig. 1—Schematic representation of energy states (or levels).

The vibrational energies of atoms bound together in molecules and molecular rotational energies are similarly quantized and may also be represented by an energy level diagram. Other sources of internal energy, such as the orientation of molecular (and nuclear) dipole moments in an applied electric or magnetic field, are also quantized.

The various internal energy states corresponding to different excitations of electronic motion generally lie a few electron-volts apart. Molecular vibrational energies are much smaller, typically of the order of $10^{-3}$ electron-volts; rotational energies are even lower, in the $10^{-4}$ to $10^{-5}$ electron-volt range, as are orientational energies of paramagnetic ions in usual laboratory magnetic fields (a few thousand Gauss). Nuclear moments are about three orders of magnitude weaker yet, corresponding to energies of only about $10^{-7}$ electron-volts.

Molecules in certain of their energy states can interact with a microwave radiation field of appropriate frequency and either absorb energy from the radiation field while jumping to a state of greater internal energy, or, under the influence of the radiation field, can give up some of their internal energy and drop to a state of lower energy. The amount of internal energy thus transferred, *i.e.*, the difference in energy of the two energy levels, $W_1 - W_2$, is linearly related to the frequency of the radiation field:

$$W_1 - W_2 = h\nu.$$

Here $h$ is Planck's constant and $\nu$ is the frequency. vhf fields (at about 100 mc) correspond to energy differences of only about $4 \times 10^{-7}$ electron-volts. These can therefore produce nuclear magnetic resonance; that is, they can interact with molecules to change the orientation of nuclear magnetic moments in fields of a few thousand Gauss. Microwave frequencies of about 10 kmc correspond to energies of about $4 \times 10^{-5}$ electron-volts, which are typical of rotational and electronic paramagnetic energies. Visible light, as mentioned earlier, corresponds to the relatively large energy associated with changes of electronic motional energies (a few electron-volts). Because of the very small energy transfers associated with low frequencies, useful molecular amplifiers are more likely to operate at frequencies above 1000 mc.

Consider two energy levels of a molecule whose energy difference corresponds to a microwave frequency. If a molecule in the lower energy state is placed in a microwave field of this frequency, it will absorb energy and hop to the upper state. An atom in the upper state, on the other hand, will give up energy to the microwave field and drop to the lower level. The probability for both transitions is the same. Therefore, whether a system of many molecules exhibits a *net* absorption or emission of energy depends on whether more molecules are in the lower or upper energy state. All molecular systems when allowed to come to thermal equilibrium have more molecules in the lower of each pair of states, and hence are absorptive. However, it is possible to build devices that will disturb the molecular system and put it in an emissive condition where more molecules are in the upper of a pair of states. Such a disturbed, emissive system is a "molecular amplifier": a microwave field incident on it will take up energy from the molecules. By providing suitable feedback, a "molecular oscillator" can be made.

Several methods of producing an emissive condition have been devised. One that has been tested is the molecular-beam Maser,[1,2] in which a beam of ammonia molecules is physically separated into two beams, one consisting of upper-state molecules and the other of lower-state molecules. The upper-state beam is then used in a resonant cavity as the emissive molecular system of a microwave amplifier. As an oscillator the beam Maser has exceedingly high spectral purity; as an amplifier it has only a very narrow amplification bandwidth.

Another device, the "hot-grid cell," also uses ammonia gas. However, unlike the beam Maser, this is a "sealed off" gas system that does not require continuous pumping. An emissive state is maintained by selective reflection of upper-state molecules at one wall. In addition to the advantage of eliminating pumps, the amplification bandwidth is an order of magnitude larger than in the beam-type Maser. It is still only tens of kilocycles, however.

To get greater bandwidths without loss of gain, molecular systems of higher density are needed. It is expected that megacycle bandwidths can be attained using solids. Two methods of exciting such systems by interchanging the populations of two levels have been found. These are: 1) excitation by applying a short controlled pulse of microwave power at the transition frequency, and 2) excitation by "adiabatic fast passage," where a strong exciting microwave field is swept in frequency past the transition frequency. However, these excitation methods are essentially pulsed ones, giving only intermittent amplification. Another method of exciting a solid molecular system that operates continuously has been developed. In this, a strong microwave field equalizes the populations of two energy states and amplification is secured at a lower frequency by inducing transitions between one of these two levels and an intermediate undisturbed level. Whereas conventional tubes require dc power to operate, all of these methods of exciting solid systems entail the supplying of energy to the system at microwave frequencies. In a solid, the lifetime of an excited state during which it can interact with radiation is determined in part by thermal motions of the lattice atoms. Therefore, solid-state devices seem to work best at very low (liquid helium) temperatures; this practical disadvantage will probably be overcome by future developments of materials in which interactions with the lattice are very weak.

[1] J. P. Gordon, H. J. Zeiger, and C. H. Townes, "The Maser—new type of microwave amplifier, frequency standard, and spectrometer," *Phys. Rev.*, vol. 99, pp. 1264–1274; August 15, 1955.

[2] The term "Maser," an acronym for "microwave amplification by stimulated emission of radiation," has been coined to describe the general class of microwave amplifiers based on molecular interaction. It is often used in a more restricted sense to refer to one form of molecular amplifier, that using a beam of excited molecules.

A very narrow band gaseous amplifier that is excited, not with microwave energy but with visible light, is also being worked on. This should provide, as does the molecular beam Maser, a tuned element of exceedingly high $Q$ for an "atomic clock" or frequency standard.

There are two basic features of molecular amplification and generation that make it attractive in comparison to electronic devices. First, molecular amplification does not have many of the sources of noise that plague electronic amplification, such as flicker noise, induced grid noise, shot noise, and partition noise.[3,4] Molecular amplifiers can therefore have low noise figures and molecular oscillators can have high spectral purity. Second, the frequencies are determined by the internal molecular structure. Therefore, the limitations on the generation of very short microwaves imposed by electron transit time effects of conventional tubes no longer restrict the attainable frequencies. Of the disadvantages of molecular amplifiers, the narrow amplification bandwidth is perhaps the most serious.

The next section gives a general discussion of the basic physical processes involved in emissive molecular systems and of the requisite molecular properties. Those who are familiar with the basic concepts in the interaction of microwaves with matter will wish to skip over this section. The specific requirements for amplification and oscillation are discussed in Section III; Section IV discusses in detail the practical approaches to the subject outlined earlier. A discussion of the present state of the art and comments on possible future developments in Section V conclude the paper.

## II. Basic Physical Properties and Processes of Molecular Systems

### A. Description of a Single Molecule

Numerous experiments have shown that an isolated molecule cannot have arbitrary internal energy, but can only have one of a *discrete* number of energies; that is, molecular internal energies are *quantized.* This behavior is in contrast to that of a free electron, whose kinetic energy, or potential energy in an electric field, can assume any of a continuous range of values.[5] Molecules and electrons may therefore be expected to have basically different types of interaction with electromagnetic radiation. To study molecular interactions, the properties of molecules in the various internal energy states must be known.

According to quantum theory, the state of a molecule is completely specified by a *state function.* Any general state of a molecule can be expressed by a linear combination of the state functions that describe the individual quantized energy states. Denote the function corresponding to the state with internal energy $W_i$ by $\psi_i$; then any general state $\psi$ can be expressed by:

$$\psi = \sum_i a_i\psi_i. \tag{1}$$

The $a_i$'s, which may be complex, are analogous to Fourier coefficients in a Fourier decomposition, and are called "probability amplitudes." They have the physical significance that if the internal energy is measured, the probability that the value $W_i$ will be obtained is equal to $|a_i|^2$. Eq. (1) has the interpretation that unless $a_i=0$ for $i\neq j$ and $|a_j|=1$, the energy of the system is *uncertain:* an energy measurement may yield *any* of the values $W_i$ for which $a_i\neq 0$. A state describable by such a state function is called a *superposition* energy state, as it is made up of a linear combination of pure energy states $\psi_i$. In predicting the result of an energy measurement, it is seen that only the *magnitude* of the $a_i$'s is important. Other properties of the system in a superposition state, however, such as its interaction with microwaves, depend on the relative *phases* of the various $a_i$'s, as well as on their magnitudes. There are certain processes that do not change the energy of the molecule, but produce changes in these relative phases; the interaction process between molecules and radiation is critically dependent on such processes.

### B. Description of Many-Molecule Systems

Thus far, only isolated single molecules have been considered. When a number of such molecules are assembled, the interaction between these elements may be weak, or sporadic. In either case, the system of individual-molecule energy levels may still be used, with the aid of statistical methods, to predict the behavior of the many-molecule macrosystem. On the other hand, in cases where the interactions are continuous and strong, the individual-molecule energy level scheme breaks down and must be replaced by one treating the macrosystem as a whole. Systems with weak or intermittent interactions between the elements will be of interest in this paper. Even in a solid, molecular systems that are but weakly coupled to each other and their surroundings may be found. Certain paramagnetic ions, introduced as impurities in a host crystal lattice, are examples.

Consider a system of many molecules. If the internal energy of one of the molecules is measured, it will be found to be one of the discrete allowed values. The question arises: what can be said about the distribution of the molecules among these allowed energy states? The answer given by statistics is the following. If the system is in thermal equilibrium at an (absolute) temperature $T$, in any sample containing many molecules the ratio of the number of molecules with energy $W_1$ to the number of molecules with energy $W_2$ is given by:

$$\frac{N(W_1)}{N(W_2)} = \frac{N_1}{N_2} = \exp\left[-\frac{(W_1 - W_2)}{kT}\right] \tag{2}$$

[3] J. R. Pierce, "Physical sources of noise," Proc. IRE, vol. 44, pp. 601–608; May, 1956.

[4] J. L. Lawson and G. E. Uhlenbeck, "Threshold Signals," McGraw-Hill Book Co., Inc., New York, N. Y.; 1950.

[5] The kinetic energy and any positional potential energy of a molecule are similarly unquantized. However, we are considering only "molecular interactions," where changes of internal, rather than kinetic (or positional), energy are involved.

where $k=1.38\times10^{-16}$ erg/°K ($1.38\times10^{-23}$ joule/°K) is Boltzmann's constant. This Boltzmann distribution has the properties that states of lower energy are more highly populated than states of higher energy and that the population ratio of any two states can be enhanced by lowering the temperature and reduced by raising it.

### C. Relaxation Effects

The Boltzmann distribution of molecules among the various possible energy states is an *equilibrium* one; that is, as long as the temperature of the system is held constant, once this distribution is attained, it will remain the same indefinitely.[6] Any distribution whatever, if it is left undisturbed for a sufficient length of time, will spontaneously change until it is an equilibrium one. The processes that produce such changes in an arbitrary distribution are known as the "relaxation mechanisms." Interaction with a "black-body" radiation field, collisions between molecules, and electric and magnetic dipolar interactions between molecules are among such processes of importance in molecular amplifiers.

For convenience in handling many problems, the relaxation processes are divided into two classes which are not necessarily mutually exclusive. These correspond to the two different ways a system can deviate from equilibrium. Once the total number of molecules in each energy state is known, the total internal energy $W$ of the molecular system is given by

$$W = \sum_i N_i(W_i) \times W_i.$$

If the total energy of the molecular system differs from its thermal equilibrium value $W_{eq}$, spontaneous changes will occur that result in a net transfer of energy between the internal degrees of freedom of the molecular system and its surroundings (a heat reservoir). Those processes that contribute to this exchange of energy are called "longitudinal" relaxation processes. In many cases of practical importance, the displacement of the molecular internal energy $W$ from its equilibrium value $W_{eq}$ decreases exponentially with time:

$$\frac{d(W - W_{eq})}{dt} = -\frac{1}{T_1}(W - W_{eq}).$$

The reciprocal of the decay constant, which has the dimensions of a time, is known as the "longitudinal" (or "spin-lattice") relaxation time. It is conventionally designated $T_1$.[7]

[6] It thus must correspond to a state of maximum entropy, or maximum disorder.

[7] The reader is warned of the possible confusion of the absolute temperature, $T$, and the relaxation times $T_1$ (and $T_2$). This symbology has become standard in the field, and hence is retained here. It arose in the field of nuclear magnetic resonance, where the internal energy is the magnetic dipolar or "spin" orientation energy of nuclei in an applied magnetic field. The term "spin-lattice" relaxation also arose in nuclear magnetic resonance, where the crystal lattice constitutes the heat reservoir.

There is another kind of relaxation process. Since the equilibrium state is one of complete randomness or molecular chaos (maximum entropy), any coherence or definite phase relationship between the various molecules constituting the system corresponds to a (partially) ordered, and hence nonequilibrium, state. In this case, the energy of the system $W$ need not depart from its equilibrium value $W_{eq}$ to produce a nonequilibrium state. Relaxation processes that tend to destroy such order in the system are known as "transverse" processes, and where they can be adequately described by an exponential decay of the order are describable phenomenologically by a "transverse" (or "spin-spin") relaxation time $T_2$. Examples of this type of relaxation process can be found that either can simultaneously transfer energy between the system and surroundings, or can involve only members of the system itself, and hence cannot exchange energy with the heat reservoir. A collision between two gas molecules, where internal energy can be transformed to kinetic energy of the molecules is an example of the former; the magnetic dipolar interaction between two localized spins in a crystal is an example of the latter.

### D. Molecular Interaction with a Radiation Field

A molecule can exchange energy with an oscillating electromagnetic field only if the frequency of the oscillation is related to the allowed internal energies of the molecule by the Bohr frequency condition

$$W_i - W_j = h\nu, \tag{3}$$

where $W_i$, $W_j$ are the internal energies of the $i$th and $j$th quantum states, $\nu$ is the frequency of the radiation, and $h=6.624\times10^{-27}$ erg-sec ($6.624\times10^{-34}$ joule-sec) is Planck's constant. The energy in a monochromatic electromagnetic wave is also quantized; that is, the energy in the wave comes in "packets," or *photons*, and can only change in finite steps corresponding to the energy of a photon. For a wave of frequency $\nu$, the energy of the photons is just $h\nu$. The interaction may then be considered to be either one of two processes: the absorption of a photon by the molecule with a corresponding increase in internal energy, or the emission of a photon with a corresponding drop in internal energy. With this picture the above frequency condition (3) follows immediately from the principle of conservation of energy. The apparent discrepancy between a quantized radiation field of discrete photons and the continuously variable electric and magnetic fields usually dealt with can be resolved by considering the size of a photon. For a microwave frequency of 10 kmc, the photon energy is only $6.6\times10^{-17}$ erg ($6.6\times10^{-24}$ joule); a wave at this frequency carrying a microwatt of power has a flux of $1.5\times10^{17}$ photons/sec. It is because ordinary practice deals with such huge numbers of photons that the discreteness of the wave energies does not generally make itself manifest.[8]

There are three basic types of radiation field-molecule interaction: absorption, induced or stimulated emission, and spontaneous emission. In absorption, as the name implies, a photon is absorbed by the molecule, which is then lifted to a state of higher energy. Induced emission is the reverse process, whereby, under the action of an applied electromagnetic field, a molecule emits a photon at the frequency of the applied field and drops to a lower internal energy level. In spontaneous emission, a molecule in an excited energy state, *i.e.*, an energy state other than the lowest, even though it is undisturbed by an applied field, emits a photon at the characteristic frequency given by the Bohr frequency condition and drops to a lower energy state. These types of interaction were treated on a phenomenological basis by Einstein[9] before an adequate microscopic quantum picture was available. His "$B$-coefficient" gives the probability for either absorption or induced emission; his "$A$-coefficient" gives the spontaneous emission probability. In amplification, we are clearly dealing with induced emission. In the case of a molecular microwave generator, the basic process is *coherent* spontaneous emission. This can alternatively be considered a case of *self-induced* stimulated emission, where the energy supplying the inducing field comes from the molecular system itself.

It is not possible to cause transitions between every two internal energy states by means of electromagnetic radiation. In addition to the conservation of energy requirement given by (3), momentum conservation and other requirements must be met. The effect of this is to impose restrictions on what pairs of states may be coupled by radiation. These restrictions are known as "selection rules," and are usually expressed in terms of allowed changes in the quantum numbers specifying the internal states. For example, the rotational energy of a simple diatomic molecule may be written as

$$W_J = BJ(J+1)$$

where $B$ is a constant depending on the molecule and $J$ is the rotational quantum number. The usual selection rule for transitions between the states of energy $W_J$ is given by

$$\Delta J = J_{\text{final}} - J_{\text{initial}} = \pm 1,$$

*i.e.*, the rotational quantum number must change by one unit when a photon is emitted or absorbed, the plus sign corresponding to absorption, and the minus to emission. This is because each photon carries one unit of angular momentum, which must be conserved.

A very important point about absorption and emission must be made. Both a microscopic quantum mechanical treatment and Einstein's phenomenological treatment give the result that, for a given applied microwave field strength, the probability of a molecule in the lower of two energy states absorbing a photon is exactly equal to the probability of a molecule in the upper of the same two states giving out a photon by induced emission. This has the consequence that any macrosystem that is in a state where two energy levels are equally populated will be completely transparent to radiation at the frequency corresponding to their energy difference. Neglecting the usually insignificant spontaneous emission effects, it is only an imbalance in populations that produces an observable effect. For computational purposes it is therefore possible, when considering transitions of molecules between two energy levels, to "pair off" the molecules as far as possible, and to deal only with the "excess" molecules in the more highly populated state.

To make this somewhat clearer, consider a gas in thermal equilibrium. Since radiation at only one frequency will be of interest, only the two states coupled by radiation at this frequency need be considered. Let the upper energy state be designated 2), the lower 1); let there be a total of $N$ molecules in the two states.

$$N_1 + N_2 = N.$$

From (2),

$$\frac{N_2}{N_1} = \exp\left[-\frac{W_2 - W_1}{kT}\right] = \exp\left[-\frac{h\nu}{kT}\right]. \tag{4}$$

Solving these equations for $N_1$ and $N_2$ gives

$$N_1 = \frac{N}{1+\exp\left(-\dfrac{h\nu}{kT}\right)},$$

$$N_2 = \frac{\exp\left(-\dfrac{h\nu}{kT}\right)N}{1+\exp\left(-\dfrac{h\nu}{kT}\right)}.$$

"Pairing off" molecules, the "excess population" is

$$N_1 - N_2 = N\tanh\left(\frac{h\nu}{2kT}\right). \tag{5a}$$

At microwave frequencies, $h\nu/2kT \ll 1$ generally holds, and (5a) becomes

$$N_1 - N_2 \approx \frac{h\nu}{2kT}N. \tag{5b}$$

[8] The question of the meaning of *phase* when a radiation field is considered to consist of particle-like photons, and other questions dealing with the "wave-particle duality" of radiation are too involved to go into here. The interested reader is referred to texts on quantum theory for further discussion, *e.g.*, D. Bohm, "Quantum Theory," Prentice-Hall, Inc., New York, N. Y. chs. 1 and 2; 1951, or W. Heitler, "The Quantum Theory of Radiation," University Press, Oxford, Eng., 2nd ed.; 1950.

[9] A. Einstein, "On the quanta theory of radiation," *Phys. Zeit.*, vol. 18, pp. 121–128; March 15, 1917. For a brief review of Einstein's argument, see, E. U. Condon and G. H. Shortley, "The Theory of Atomic Spectra," Cambridge Univ. Press Cambridge, Eng., ch. 4; 1951.

(At extremely low temperatures, the exact (5a) must be used.) If radiation at frequency $\nu$ is incident on the gas, it will absorb energy (since $N_1 > N_2$) at the same rate as a gas composed of $(h\nu/2kT)N$ molecules all of which are in the lower of the two energy states.

It has been seen that an electromagnetic field of the proper frequency can induce transitions between the energy levels. It must be remembered, however, that while this is going on, the relaxation mechanisms are operative, tending to restore the system to thermal equilibrium. For weak transition-inducing fields, the relaxation mechanisms are strong enough to maintain the system in an absorptive state not much different from that of thermal equilibrium. However, if the electromagnetic field strength is increased, molecules tend to undergo induced transitions to the upper energy state faster than the relaxation processes can restore them to the lower state. The effective "pairing off" of molecules is then shifted from that at equilibrium, and there are fewer molecules now in the "excess" to absorb. The total absorbed power does not decrease, as the power absorption per molecule is greater at higher fields. It does reach a limit, however, determined by the maximum rate that the relaxation processes can remove the internal energy from the molecular system thus supplied by the microwave field. Under these conditions, the microwave absorption is said to be "saturated." An analogous situation will be encountered in considering amplification.

The above discussion has assumed that the energy levels are perfectly sharp, that is, that only one (monochromatic) frequency couples two energy levels. This is not strictly true: transitions may be induced by a band of frequencies centered at the frequency given by (3). Thus, absorption or emission spectral "lines" have a finite frequency width which can be ascribed to a "width" of the corresponding energy levels. The physical reason for this width is that because of relaxation processes, a molecule interacts with the radiation field for only a finite time before it is "dephased" and (effectively) the interaction stops momentarily. But in an interaction time $\sim T_2$, a molecule will "see" essentially the same number of cycles of the microwave field as long as the frequency lies in a band of frequencies of width $\Delta\nu \approx (1/T_2)$. In other words, a molecule is not able to tell what frequency is acting on it to within a range $\Delta\nu$ in the time $T_2$ before the interaction is interrupted. The spectral width of a molecular resonance should thus be given by $\Delta\nu T_2 \sim 1$, with short relaxation times leading to broad lines and conversely. This is as observed: in a gas, where $T_2$ is inversely proportional to the pressure, the observed line width is also proportional to the pressure. Dephasing processes not describable by a relaxation time $T_2$, such as the Doppler effect, have a similar result. Such effects are commonly met with in the microwave spectroscopy of gases and provide a means of studying molecular interaction processes in detail.[10,11] In a solid, where the molecules are very near to each other, strong interactions, leading to short relaxation times, are frequently encountered. Resonances in solid systems are thus apt to cover a broad band of frequencies. (Resonances of over a hundred megacycles are generally too broad to use for amplification purposes, but widths roughly one megacycle and up can be found.)

In some cases it happens that certain frequencies couple more than one pair of states. This may happen because two or more internal states correspond to the same energy (degeneracy), or because two or more pairs of levels "accidentally" are separated by the same energy gap. The first of these cases is the more common. However, in many systems of interest neither effect occurs. To simplify the discussion, it will be assumed that neither occurs in all the following. It is a good approximation to ignore the effects of all energy states but the two that are directly coupled. This reduces the situation to a relatively simple two-state case.

### E. Properties of a Two-State System

In the case that only two energy levels are of importance, one deals with a system of molecules that interacts with the radiation field only when the frequency of the field is near the resonant frequency $\nu_0$. Such a radiation field induces transitions between the two energy states, molecules in the lower state undergoing transitions that absorb energy from the radiation field and lift them to the upper energy state, and molecules in the upper energy state giving up energy to the field and dropping to the lower state. As mentioned earlier, the transition probability for an atom in the lower state to absorb a photon and jump to the upper state is exactly equal to that for an atom in the upper state to emit a photon and jump to the lower state (neglecting the usually small spontaneous emission effects). A semiquantum mechanical calculation of these transition probabilities is given in the Appendix. The basic result of interest here is that if a molecule is in either one of the energy states at time $t = 0$, at time $t$ the probability that it has undergone a transition and is in the other state is given by:

$$\text{Transition Probability} = \frac{\left(\frac{pE}{h}\right)^2}{(\nu - \nu_0)^2 + \left(\frac{pE}{h}\right)^2} \cdot \sin^2\left\{\pi\left[(\nu - \nu_0)^2 + \left(\frac{pE}{h}\right)^2\right]^{1/2} t\right\}. \quad (6)$$

[10] W. Gordy, W. V. Smith, and R. F. Trambarulo, "Microwave Spectroscopy," John Wiley and Sons, Inc., New York, N. Y.; 1953.

[11] C. H. Townes and A. L. Schawlow, "Microwave Spectroscopy," McGraw-Hill Book Co., Inc., New York, N. Y.; 1955.

A dipole transition has been assumed, with $p$ the molecular electric dipole moment and $E$ the amplitude of the applied microwave field. (Although the electric case has been assumed, all expressions are equally valid if $E$ is replaced by a microwave magnetic field $H$[12] if $p$ is then taken as the *magnetic* dipole moment.) $\nu$ is the frequency of the applied field and $\nu_0=(W_2-W_1)/h$. Eq. (6) is plotted in Fig. 2; as shown there, the transition probability oscillates between zero and

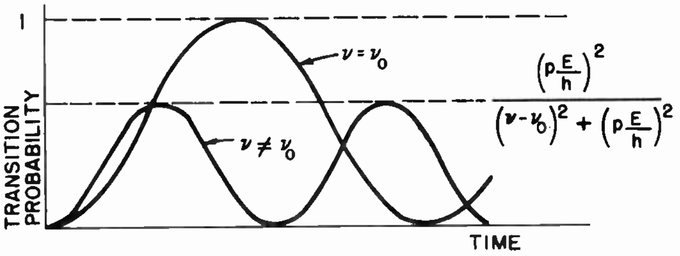


Fig. 2—Transition probabilities between energy states as a function of time for molecules initially in one of two microwave-coupled states when subjected to microwave radiation at or near the resonant frequency.

$$\frac{\left(\frac{pE}{h}\right)^2}{(\nu-\nu_0)^2+\left(\frac{pE}{h}\right)^2}.$$

On resonance, $(\nu=\nu_0)$, a molecule oscillates between the two states, while somewhat off resonance, a molecule that starts in one state never quite gets to the other (with certainty) before the process is reversed.

The average power flow to or from the radiation field due to transitions depends on a suitable averaging of transition probabilities, the averaging being dependent on the physical situation envisioned. One important case is that of collisions randomly distributed in time determining the interaction period. From (6), the instantaneous rate of energy transfer is

$$P(t) = P(t-t_1) = h\nu_0 \times \frac{d}{dt}\,(\text{Transition probability})$$

$$= \frac{\pi h\nu_0\left(\frac{pE}{h}\right)^2}{\left[(\nu-\nu_0)^2+\left(\frac{pE}{h}\right)^2\right]^{1/2}}$$

$$\cdot \sin\left\{2\pi\left[(\nu-\nu_0)^2+\left(\frac{pE}{h}\right)^2\right]^{1/2}(t-t_1)\right\}. \quad (7)$$

Here the last interruptive collision occurred at time $t_1$.

[12] Strictly speaking, $B$ should be used instead of $H$ for the microwave field. Common usage in the field, however, replaces $B$ with $H$; to avoid confusion with the literature, $H$ will be used throughout this paper.

The probability that a molecule had its last collision between times $t_1$ and $t_1+dt_1$ is given by

$$\frac{1}{T_2}\exp\left[-\frac{(t-t_1)}{T_2}\right]dt_1.$$

The average power is found by averaging (7) over this distribution of times $t_1$:

$$P_{\text{avg}} = \int_{-\infty}^{t} P(t-t_1)\,\frac{1}{T_2}\exp\left[-\frac{(t-t_1)}{T_2}\right]dt_1$$

$$P_{\text{avg}} = \frac{h\nu}{2T_2}\left(\frac{pE}{h}\right)^2 \frac{1}{(\nu-\nu_0)^2+\left(\frac{1}{2\pi T_2}\right)^2+\left(\frac{pE}{h}\right)^2}.$$

The above averaging over collision times implicitly assumes that $T_1=T_2$. A calculation made under the assumption of $T_1\neq T_2$ gives the more general result

$$P_{\text{avg}} = \frac{h\nu}{2T_2}\left(\frac{pE}{h}\right)^2 \frac{1}{(\nu-\nu_0)^2+\left(\frac{1}{2\pi T_2}\right)^2+\frac{T_1}{T_2}\left(\frac{pE}{h}\right)^2}. \quad (8)$$

This is the often-observed Lorentz line shape (see Fig. 3). From the last factor in (8) it is seen that the power

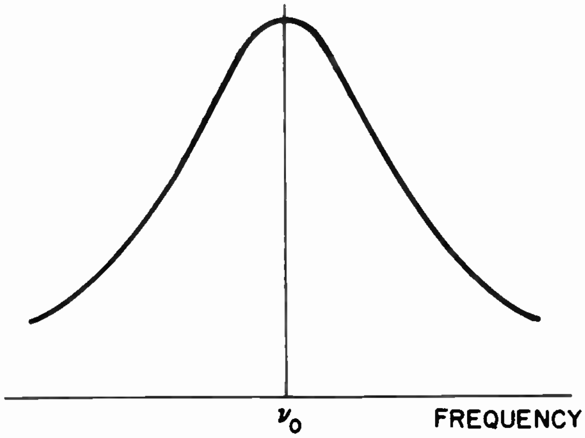


Fig. 3—Average power transferred between molecules and a microwave field at frequencies near resonance, showing the often-observed Lorentz shape.

is roughly independent of the frequency as long as

$$(\nu-\nu_0)^2 \ll \left(\frac{1}{2\pi T_2}\right)^2+\frac{T_1}{T_2}\left(\frac{pE}{h}\right)^2,$$

while it drops off rapidly with increasing departure from resonance in the range

$$(\nu-\nu_0)^2 \gg \left(\frac{1}{2\pi T_2}\right)^2+\frac{T_1}{T_2}\left(\frac{pE}{h}\right)^2.$$

The frequency bandwidth for interaction is thus approximately given by

$$(\nu - \nu_0)^2 \approx \left(\frac{1}{2\pi T_2}\right)^2 + \frac{T_1}{T_2}\left(\frac{pE}{h}\right)^2. \tag{9}$$

This illustrates a general feature common to all physical situations: for short transverse relaxation times $T_2$, the interaction is spread over a broad band of frequencies, and conversely, and for high microwave field strengths, the interaction frequency width is similarly (saturation) broadened.

With a different physical situation, such as in a device employing a beam of molecules where the interaction time is the time for the beam to traverse the device, the time averaging is different from the above and the resulting line shape is non-Lorentz. The basic ideas and general results are the same whatever the proper averaging, however.

### F. Classical Treatment

The foregoing has been a semiquantum-mechanical *microscopic* approach, where the results of experiments with a large number of molecules were deduced on the basis of what happens to individual molecules. In most cases of interest for amplification, only the gross *macroscopic* behavior is important, and all results can be obtained on a phenomenological basis using a purely classical treatment.

The connection between the microscopic and a macroscopic treatment is easily seen in the case of a system of electron spins in a magnetic field, for example. The elementary dipoles can exist in only two energy states, corresponding to alignment roughly along and against the field. As seen from (5b), at thermal equilibrium the excess number of spins in the lower energy state, leading to alignment with the field, is given by

$$N_1 - N_2 = \frac{h\nu}{2kT} N.$$

Now, however,

$$h\nu = 2\mu H,$$

where $\mu$ is the magnitude of the component of an elementary dipole along the field, so that

$$N_1 - N_2 = \frac{N\mu}{kT} H.$$

The macroscopic magnetic moment is given by

$$M = \mu(N_1 - N_2) = \frac{N\mu^2}{kT} H.$$

From this, a static magnetic susceptibility $\chi_0$ can be defined by

$$M = \chi_0 H,$$

with

$$\chi_0 = \frac{N\mu^2}{kT}.$$

(The discrepancy by a factor of 3 between this and the classical Langevin formula is apparent rather than real: $\mu$ is the field-parallel component of the total dipole moment $\mu_0$, and it can be shown quantum mechanically that $\mu_0 = \sqrt{3}\mu$.)

The macroscopic moment $\vec{M}$ is parallel and proportional to the total angular momentum $\vec{A}$ of the system.

$$\vec{M} = \gamma\vec{A}.$$

$\gamma$ is known as the *gyromagnetic ratio* of the system. The classical relation between the torque applied to a system possessing angular momentum and the rate of change of the angular momentum then gives the dynamic behavior of the molecular system.

$$\frac{d\vec{A}}{dt} = \text{Torque} = (\vec{M} \times \vec{H})$$

$$\frac{d\vec{M}}{dt} = \gamma(\vec{M} \times \vec{H}),$$

or

$$\frac{dM_x}{dt} = \gamma(M_y H_z - M_z H_y)$$

$$\frac{dM_y}{dt} = \gamma(M_z H_x - M_x H_z)$$

$$\frac{dM_z}{dt} = \gamma(M_x H_y - M_y H_x).$$

Take the direction of the (assumed large) static magnetic field along the $z$ axis and assume that the microwave magnetic fields are in the $x-y$ plane. The spin energy of the system is then given by

$$W = -M_z H_z,$$

with a thermal equilibrium value of

$$W_{eq} = -M_{eq} H_z.$$

If the effects of relaxation processes are included in the above equations, they become

$$\frac{dM_x}{dt} = \gamma(M_y H_z - M_z H_y) - \frac{1}{T_2} M_x$$

$$\frac{dM_y}{dt} = \gamma(M_z H_x - M_x H_z) - \frac{1}{T_2} M_y$$

$$\frac{dM_z}{dt} = \gamma(M_x H_y - M_y H_x) - \frac{1}{T_1}(M_z - M_{eq}).$$

These are the famous equations first introduced by Bloch[13] to explain the phenomena of nuclear magnetic resonance.

Similar classical treatments can be given for the molecular amplification systems discussed later in this

[13] F. Bloch, "Nuclear induction," *Phys. Rev.*, vol. 70, pp. 460–474; October 1, 1946.

paper by introducing a *complex* electric or magnetic microwave susceptibility, the imaginary part corresponding to the absorption (or emission) of energy by the molecular system and the real part to the molecularly-induced dispersion. In general, however, the *microscopic* approach will be used in this paper as it gives a better detailed picture of the processes occurring.

## III. Basic Amplifier and Oscillator Relations

In the preceding section, a discussion of the basic physical processes of importance in dealing with microwave amplification using the molecular interaction between matter and radiation was given. These ideas will now be applied to the microwave amplification or generation process itself, in particular, to the possible time-dependence of gain, the power handling capabilities, the available gain, the amplification bandwidth, the noise figure, and the efficiency of molecular amplifiers.

### *A. Continuous and Pulse Operation*

The excitation process of the molecular system during which it is put into an emissive condition, *i.e.*, one in which there are more molecules in the upper of two microwave-coupled energy states, inevitably interferes with the amplification or oscillation process. Thus, if all molecules in the system are excited simultaneously only pulse operation[14] is possible, and each useful period of amplification or oscillation must be followed by a "dead" period while the system undergoes excitation. To achieve continuous operation, the various molecules constituting the macrosystem must undergo excitation at different times. The excitation process can occur either in the same region as the amplification process, or the molecules can be excited outside the interaction region and then brought into the amplification region. Each of the above possibilities is used in one or more of the devices to be discussed later.

In all molecular devices, both amplifiers and oscillators, power losses in the microwave structures are of great importance. To have an amplifier (or oscillator), it is essential that more power be transferred to the field from the molecular system than is lost (to resistive heating) in the structure. In a pulsed system, the state of the macrosystem is a function of time, and hence whether or not this condition for amplification is satisfied may also be time-dependent.

### *B. Saturation*

As seen above (8) for absorption (it is also true for emission), the transfer of energy to the radiation field is proportional to the incident power, and hence the amplification is linear, as long as

$$\frac{T_1}{T_2}\left(\frac{pE}{h}\right)^2 \ll \left(\frac{1}{2\pi T_2}\right)^2$$

or

$$E \ll \frac{h}{2\pi p(T_1T_2)^{1/2}}. \qquad (10)$$

Physically, linear amplification is only obtained when but a small fraction of the total available energy in a molecular system is extracted by induced emission. Thus, unless the strength of the transition-inducing microwave field satisfies the above condition, the amplifier will be operating in a nonlinear (saturated) fashion. As seen from (9), the frequency bandwidth for interaction (with no saturation) is related to the relaxation time by

$$(\text{Molecular bandwidth}) \lesssim \frac{1}{\pi T_2}.$$

Therefore, the condition for linearity of amplification is

$$E \ll \frac{h}{2p}\left(\frac{T_2}{T_1}\right)^{1/2} \times (\text{Molecular bandwidth}).$$

For electric dipole transitions, such as are common in a gaseous system, $p$ is of the order of $10^{-18}$ esu[15] ($1.6\times10^{-29}$ coulomb-meters); for magnetic dipole transitions, as would be found in a paramagnetic amplifier, $p$ is about $10^{-20}$ emu ($10^{-23}$ joule-meter$^2$/weber). Thus, for linear amplification, the microwave fields must satisfy the conditions:

$$\begin{aligned} E\ (\text{electric dipole}) &\ll 3\times10^{-9}\left(\frac{T_2}{T_1}\right)^{1/2} \\ &\quad\times (\text{Molecular bandwidth})\ \text{stat-volts/cm} \\ &\ll 10^{-4}\left(\frac{T_2}{T_1}\right)^{1/2} \\ &\quad\times (\text{Molecular bandwidth})\ \text{volts/meter} \\ H\ (\text{magnetic dipole}) & \\ &\ll 3\times10^{-7}\left(\frac{T_2}{T_1}\right)^{1/2} \\ &\quad\times (\text{Molecular bandwidth})\ \text{Gauss} \\ &\ll 3\times10^{-11}\left(\frac{T_2}{T_1}\right)^{1/2} \\ &\quad\times (\text{Molecular bandwidth})\ \text{webers/meter}^2. \end{aligned} \qquad (11)$$

These conditions for linearity can be related to power flow into a cavity amplifier. For a solid-state (paramagnetic) amplifier at 10 kmc, for a bandwidth of 1 mc and $T_1=T_2$, the microwave magnetic field strength in the cavity must be much less than a Gauss. This corre-

[14] A point of terminology arises here. It is always possible to gate a continuous amplifier (or oscillator) off and on, thereby achieving intermittent or "pulsed" operation. In this paper, however, the term "pulse operation" will not refer to such "gated continuous operation" but only to systems where the energy transfer to the field is intrinsically intermittent due to the mode of excitation.

[15] Gaussian units are used throughout this paper, *i.e.*, cgs electrostatic units (esu) for electric quantities and cgs electromagnetic units (emu) for magnetic quantities. This choice is made to conform with conventional usage in the literature of this field, where Gaussian units are almost invariably used.

sponds to a input power much less than a watt. As the example chosen represents a type of molecular amplifier that is relatively difficult to saturate, it is clear that low power handling capabilities should be expected in molecular amplifiers.

*C. Gain*

In considering the gain obtainable from a molecular amplifier, the physical situation must again be considered. A rough estimate for a cw amplifier consisting of a waveguide containing molecules in an emissive condition will first be given. For such an amplifier the molecules must have a proper phase relationship with the signal field or they will absorb rather than emit energy. Proper phasing can be assured by bringing the molecules into interaction with the signal field in the pure energy state $\psi_2$, corresponding to molecules in the upper energy level, rather than in a superposition energy state. Assuming that the signal to be amplified is on resonance, $\nu=\nu_0$, and is weak enough to avoid saturation effects, the average power emitted by each molecule is, from (8),

$$P = \frac{2\pi^2\nu p^2E^2T_2}{h}. \tag{12}$$

If $N$ molecules contribute to the emission process, the total power is

$$P_{\text{total}} = \frac{2\pi^2N\nu p^2E^2T_2}{h}.$$

The incident power in a waveguide is of the order of

$$P_{\text{inc}} = \frac{cAE^2}{8\pi}$$

where $c$ is the velocity of light and $A$ the cross sectional guide area. Since the power gain in an infinitesimal length of amplifier is proportional to the incident power, the wave will grow exponentially as it travels down the guide, past the emitting molecules. The gain can thus be written, neglecting guide attenuation, as

$$\text{Gain} = \exp(\alpha l)$$

where $l$ is the length of the amplifier and the gain coefficient $\alpha$ is given by:

$$\alpha = \frac{P_{\text{emitted/cm}}}{P_{\text{incident}}} \approx \frac{16\pi^3\nu p^2T_2N_0}{hcA}. \tag{13}$$

Here $N_0$ is the number of molecules per unit length contributing to the emission, that is, in the upper state "excess." Under realizable conditions (for example, in the "hot-grid cell" discussed later), $\alpha$ may be of the order of $10^{-3}$ to $10^{-4}$ cm$^{-1}$. Thus, to get a gain of ten, an amplifier length of several meters is required.

Physically, (13) can be understood as follows. The gain increases with $\nu$, as the energy given up by each interacting molecule (the "size" of each photon) is proportional to the frequency. Stronger coupling between the radiation field and the molecules as determined by the dipole moment $p$ also results in higher gain, as does increased "excess population" of emissive molecules $N_0$. Finally, the longer the time $T_2$ a molecule interacts with the radiation field before it is disturbed, the greater the probability of its giving up its available internal energy as a photon and thus contributing to the amplification.

To avoid the practical difficulties associated with a long amplifier, a resonant structure such as a cavity that provides a long effective electrical path in a physically short length may be employed. If the molecular system is located in a resonant cavity, positive feedback is automatically built into the amplifier. It is well-known that the gain of such a regenerative amplifier can be arbitrarily large, becoming infinite at the threshold of oscillation. The actual situation may comprise a transmission cavity or a reaction cavity, with arbitrary couplings. A case particularly easy to analyze that displays many of the features of interest is that of a reaction cavity matched to the transmission line (Fig. 4). The output of the amplifier will be defined as the power leaving the cavity, and the input as the incident power. The behavior of such an amplifier can be seen in the following way.

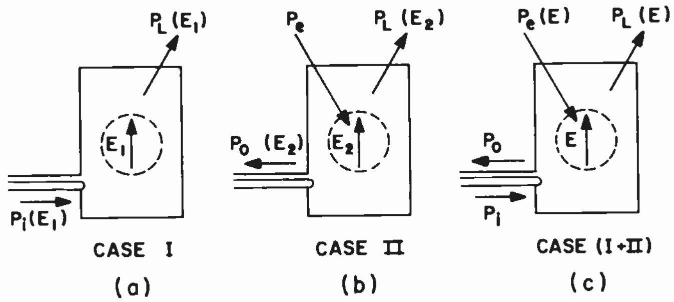


Fig. 4—Power and field relations in a matched reaction cavity amplifier. (a) A wave is incident on the cavity, building up an oscillating field $E_1$ in the cavity and being dissipated in resistive losses at the cavity walls. (b) An oscillating molecular system supplies power $P_e$ in the cavity, building up a field $E_2$ and producing an outgoing wave, $P_0$. (c). The two cases are superposed.

Consider two situations:

1) Power $P_i$ is incident on the cavity, which contains the molecular system, but the molecular interaction is effectively "turned off" by shifting the interaction frequency (*e.g.*, by "detuning" with a Stark[16] or Zeeman field). A resonant field of strength $E_1$ at the molecular system is built up in the cavity, and there are wall losses $P_L$. Since the cavity is matched, there is no outgoing wave. [See Fig. 4(a).] In this case

$$P_i = P_L = \frac{\nu V\overline{E_1^2}}{4Q_0} \tag{14}$$

[16] The shift in the energy of a molecular state due to an applied electric field is known as the Stark effect, after its discoverer. A similar shift of an energy level in a magnetic field is known as the Zeeman effect.

where $\overline{E_1^2}$ is the cavity field squared averaged over the cavity,

$$\overline{E_1^2} = k_0\overline{E_1^2}, \tag{15}$$

$V$ is the cavity volume and $Q_0$ is the unloaded cavity $Q$. $k_0$ is related to the "filling factor."

2) In this situation there is no incident wave, but the molecular system in the cavity is supplying energy to the field, setting up an oscillating field $E_2$. Now there is an outgoing wave $P_0$ and wall losses $P_L(E_2)$. [See Fig. 4(b).] These are related to $E_2$ by

$$P_0 = P_L = \frac{\nu V \overline{E_2^2}}{4Q_0}. \tag{16}$$

Let $P_e$ be the power emitted by the molecular system. Then

$$P_e = P_0 + P_L.$$

Finally, superpose the two configurations so that the fields $E_1$ and $E_2$ are in phase [Fig. 4(c)]. Conservation of energy requires that

$$P_i + P_e = P_0 + P_L.$$

$P_e$ and $P_L$ are functions of the total field $E = E_1 + E_2$ in the cavity, whereas $P_i$ is related only to the part $E_1$ and $P_0$ to the part $E_2$:

$$P_i(E_1) + P_e(E) = P_0(E_2) + P_L(E). \tag{17}$$

But

$$P_i(E_1) = \frac{k_0\nu V}{4Q_0} E_1^2$$

$$P_0(E_2) = \frac{k_0\nu V}{4Q_0} E_2^2$$

from (14), (15), and (16), while

$$P_L(E) = \frac{k_0\nu V}{4Q_0} E^2.$$

To make the discussion definite, assume $P_e$ has the Lorentz form given by (8):

$$P_e(E) = \frac{k_1 E^2}{(\nu - \nu_0)^2 + \left(\frac{1}{2\pi T_2}\right)^2 + \left(\frac{pE}{h}\right)^2}. \tag{18}$$

Defining

$$D \equiv \frac{k_0\nu V}{4Q_0}\left(\frac{h}{p}\right)^2\left[(\nu - \nu_0)^2 + \left(\frac{1}{2\pi T_2}\right)^2\right]$$

$$S \equiv \frac{k_1 h^2}{p^2}$$

$$e_i \equiv \frac{\left(\frac{p}{h}\right)}{\left[(\nu - \nu_0)^2 + \left(\frac{1}{2\pi T_2}\right)^2\right]^{1/2}} E_i,$$

the conservation of energy (17) becomes

$$De_1^2 + \frac{Se^2}{1 + e^2} = De_2^2 + De^2.$$

Defining further an "oscillation parameter" $L$:

$$L \equiv \frac{S}{2D} = \frac{2k_1 Q_0}{k_0\nu V\left[(\nu - \nu_0)^2 + \left(\frac{1}{2\pi T_2}\right)^2\right]}$$

the equation becomes, using $e = e_1 + e_2$,

$$e^3 - e_1 e^2 + (1 - L)e - e_1 = 0. \tag{19}$$

Solutions of this equation, giving $e$, and hence $e_2$, as a function of $e_1$ for a fixed value of $L$ can be easily, if tediously, found. The gain of the amplifier is given by

$$\text{Gain} = \frac{E_2^2}{E_1^2} = \frac{e_2^2}{e_1^2}.$$

Gain as a function of input power (actually $e_1^2$) is plotted in Fig. 5 for various $L$; output power ($e_2^2$) is

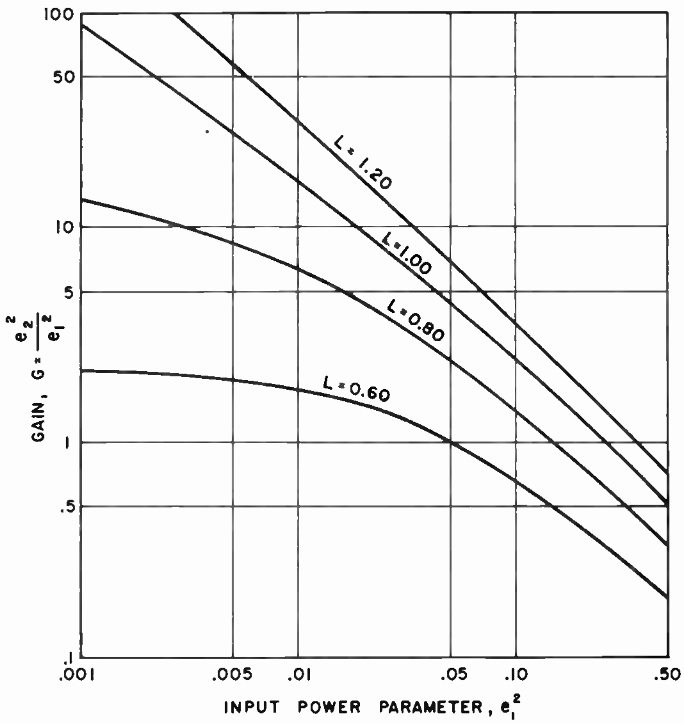


Fig. 5—Gain vs input power parameter $e_1^2$ for various values of oscillation parameter $L$ for a matched reaction cavity amplifier. $L$ is the ratio of power supplied by the (unsaturated) molecular system to twice the power lost resistively at the walls.

plotted against input power ($e_1^2$) in Fig. 6. Consideration of these figures shows that the optimal operating condition is with low $e_1$ (no saturation). In this limit, the gain is

$$\text{Gain (no saturation)} = \frac{L^2}{(1 - L)^2}. \tag{20}$$

When there is no incident power, (19) becomes

$$e^3 + (1 - L)e = 0$$

with the solutions $e=0$ and $e^2=L-1$. For $L<1$, the second solution is meaningless. However, for $L>1$, a finite real field can exist in the cavity, *i.e.*, oscillations can be sustained. Thus it is appropriate to call $L$ the "oscillation parameter." The figures clearly show the saturation behavior at high incident fields where the gain falls off, eventually becoming less than one when the power supplied by the molecular system is unable to compensate for the large wall losses associated with the large cavity fields due to the large input wave. Eq. (20) for low power level gain shows the characteristic increase in gain (to infinity) as the oscillation parameter is increased to one.

It is similarly possible to analyze the case of a transmission cavity with arbitrary coupling, but this general case is mathematically rather unwieldy. Similarly, the details, but not the general picture, will be changed for a physical situation giving a relationship between molecularly emitted power $P_e$ and microwave field strength $E$ differing from (18).

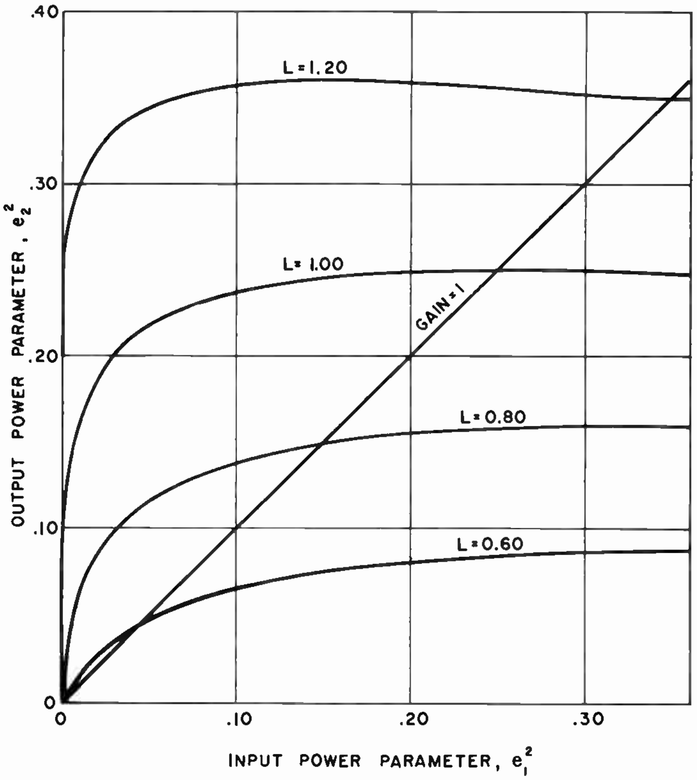


Fig. 6—Output power parameter $e_2^2$ vs input power parameter $e_1^2$ for various values of oscillation parameter $L$ for a matched reaction cavity amplifier.

### D. Bandwidth

Two factors influence the amplification bandwidth: the molecular relaxation processes and the microwave structure. (It is possible, of course, that the latter will also determine the former.) The relaxation processes determine what might be called the molecular bandwidth, which is the width of the resonance line as observed in ordinary absorptive spectroscopy. This width has been discussed earlier in Section II. The actual amplifier will have a somewhat narrower bandwidth, however.

To see this, consider the case of a nonresonant waveguide amplifier with a Lorentz molecular band-shape;

$$\text{Gain} = \exp(\alpha l)$$

and

$$\alpha = \frac{\text{const.}}{(\nu - \nu_0)^2 + \left(\frac{1}{2\pi T_2}\right)^2}.$$

Defining the bandwidth as the frequency interval between the points where the gain is one-half its maximum value, it is readily shown that

(Amplification bandwidth)

$$= \left[\frac{\ln 2}{\ln \frac{(\text{Gain})_{\max}}{2}}\right]^{1/2} \times \frac{1}{\pi T_2}$$

$$= \left[\frac{\ln 2}{\ln \frac{(\text{Gain})_{\max}}{2}}\right]^{1/2} \times (\text{Molecular bandwidth}).$$

Similarly, for the case of a matched-cavity amplifier with a Lorentz-type resonance, from

$$\text{Gain} = \frac{L^2}{(1 - L)^2}$$

and the oscillation parameter $L$,

$$L = \frac{\text{const.}}{(\nu - \nu_0)^2 + \frac{1}{(2\pi T_2)^2}},$$

assuming high gain, *i.e.*, $L \approx 1$, it can be shown that

(Amplification bandwidth)

$$= \frac{(\sqrt{2} - 1)^{1/2}}{(\text{Gain})^{1/4}} \times (\text{Molecular bandwidth})$$

$$= \frac{0.644}{(\text{Gain})^{1/4}} \times (\text{Molecular bandwidth}).$$

The above expressions show that the gain-bandwidth product is not constant, but increases with increasing gain. For reasonable gains ($10 \leqq \text{Gain} \leqq 100$), bandwidths between one-fifth and two-thirds the "molecular width" may be expected.

One of the greatest limitations on molecular amplifiers is the difficulty of getting useful bandwidths. From the discussion of Section II, the molecular bandwidth is inversely proportional to the relaxation time. Eq. (13) shows that the gain increases with frequency, dipole moment, relaxation time, and density of "active" molecules. If the bandwidth is increased by shortening the relaxation (or interaction) time in some way, the gain will fall off unless there are compensating changes in either frequency, dipole moment, or molecular density. The possible range of these parameters must thus be considered.

In an amplifier where the molecular system is gaseous, electric dipole transitions, which are much stronger than magnetic dipole transitions, can be used. However, dipole moments are of the order of $10^{-18}$ esu ($1.6\times10^{-29}$ coulomb-meters) whatever the gas used; varying the gas molecule used will not appreciably alter $p$. It might be thought that if in a gas amplifier gas collisions are the dominant relaxation mechanisms, any desired bandwidth could be achieved by increasing the density, since $T_2N_0$ is independent of pressure.

$$\left(T_2 \sim \frac{1}{N_0},\ \text{so bandwidth} \sim N_0\right).$$

In theory this is so; in practice, excitation mechanisms fail to work at high gas pressures. This at present provides an upper limit to expected bandwidths of under 100 kc. Some improvement can be expected by going to molecular resonances at higher frequencies. However, increased losses in the associated microwave circuits tend to reduce the gain obtained in this way.

Another possible way of increasing bandwidth is to use solid (or liquid) molecular systems, with their huge increases in density over gaseous systems. This can compensate for the gain lost by spreading the molecular emission energy over a greater range of frequencies, and greater bandwidths can be obtained this way; however, there are important counterbalancing effects to be considered. One is that in condensed systems magnetic dipole transitions must be used, as the strong, rapidly fluctuating electric fields encountered in liquids and solids relax electric dipole-coupled systems so rapidly that they cannot be excited with presently-available techniques. The largest magnetic dipoles found are of the order of a Bohr magneton, or

$$10^{-20}\ \text{emu}\left(10^{-23}\ \frac{\text{joule-meter}^2}{\text{weber}}\right).$$

Since the dipole moment enters as the square, whereas the relaxation time enters only as the first power in the gain coefficient expression (13), a density increase of $10^4$ is required just to counterbalance the effects of the weaker dipole moment.

Even with magnetic dipoles, relaxation times may be unusably short. At present, paramagnetic materials seem the most promising, the paramagnetism arising from unpaired electron spins. The energy levels determining the resonance frequency are generally in this case associated with the Zeeman energy of the two possible spin orientations with respect to an external applied magnetic field. The transition frequency is given by

$$\nu = \frac{g\mu_B H}{h} \equiv \gamma H.$$

Here $g$ is the spectroscopic splitting factor and $\mu_B = 9.27\times10^{-21}$ erg/Gauss ($9.27\times10^{-24}$ joule-meter²/weber) is the Bohr magneton. For paramagnetic spin systems with usable relaxation times, $g\approx 2$. This gives a frequency of

$$\nu = 2.8H \text{ mc}$$

where $H$ is in Gauss.[17] To avoid seriously broadening the resonance by magnetic field inhomogeneities, it is necessary to have a field uniform over the sample to a fraction of a per cent of the total field. An additional advantage of a paramagnetic amplifier over a gaseous one is that the frequency is readily controllable by means of the applied magnetic field.

### E. Noise Figure

It was pointed out in the introduction that one of the important properties of molecular amplification is that it may introduce very little additional noise onto a signal. To see this, an expression for the noise figure of a waveguide-type of molecular amplifier will now be derived. Other discussions of molecular amplifier noise have been given in the literature.[1,18]

Referring to Fig. 7, let $P_s$ be the signal power in a wave traveling to the right at position $x$, and $P_N$ the corresponding noise power (in a unit frequency band about the frequency of $P_s$). In a length of amplifier $dx$, the increase in signal power $dP_s$ is given by

$$dP_s = (N_2 - N_1)\Gamma P_s dx - \alpha_g P_s dx. \tag{21}$$

Here $N_2$ is the number of molecules in the upper energy state per unit length, $N_1$ the number in the lower state per unit length, $\alpha_g$ is the waveguide attenuation coefficient, and $\Gamma$ is a constant depending on the dipole matrix element for the transition and the guide geometry and mode. The first term on the right-hand side of (21) represents the molecular amplification (increase in signal power due to excess of stimulated emission over

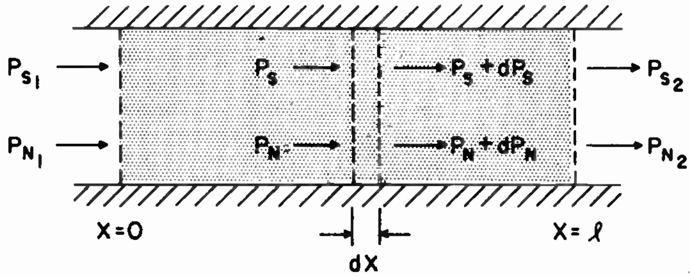


Fig. 7—Signal and noise powers in a waveguide amplifier.

[17] It should be noted that nuclear magnetic moments could in principle be used instead of electronic moments. However, the $\gamma$'s corresponding to nuclear moments are two to three orders of magnitude smaller than the electronic $\gamma = 2.8$ mc/Gauss, and thus magnetic fields in the $10^5$–$10^6$ Gauss range are necessary to put the transition frequency in the microwave range.

[18] K. Shimoda, H. Takahasi, and C. H. Townes, "Fluctuations in amplification of quanta" (to be published).

absorption) and the second the reduction in signal power due to resistive losses in the walls.

A similar expression can be written for the increment of noise power:

$$dP_N = (N_2 - N_1)\Gamma P_N dx - \alpha_g P_N dx + \alpha_g P_{N0} dx + \beta N_2 dx. \qquad (22)$$

The first two terms on the right here are exactly analogous to the terms in the signal power expression. There are two additional terms, however. The first of these, $\alpha_g P_{N0} dx$, represents black-body radiation from the walls. The subscript 0 refers to a condition of thermal equilibrium at the wall temperature $T_0$, taken to be room temperature. The second additional term, $\beta N_2 dx$, represents the effects of spontaneous incoherent emission: it is proportional to the number of molecules in the upper energy state. (The proportionality constant $\beta$, like $\Gamma$, depends on the dipole moment and guide geometry.)

The relationship between $\beta$ and $\Gamma$ can be determined by considering the special condition of thermal equilibrium. Then $dP_N = 0$, and (22) becomes

$$0 = (N_{20} - N_{10})\Gamma P_{N0} dx - \alpha_g P_{N0} dx + \alpha_g P_{N0} dx + \beta N_{20} dx$$

or

$$\beta = \left(\frac{N_{10}}{N_{20}} - 1\right)\Gamma P_{N0}.$$

But

$$\frac{N_{20}}{N_{10}} = \exp\left(-\frac{h\nu}{kT_0}\right),$$

from (4), and the Planck distribution law for black-body radiation gives

$$P_{N0} = \frac{h\nu}{\exp\left(\frac{h\nu}{kT_0}\right) - 1}.$$

From these is obtained the relationship between $\beta$ and $\Gamma$:

$$\beta = h\nu\Gamma.$$

Rewriting (22) using this relation gives

$$dP_N = [(N_2 - N_1)\Gamma - \alpha_g]P_N dx + (h\nu\Gamma N_2 + \alpha_g P_{N0})dx.$$

This can be directly integrated to give

$$\frac{P_{N2} + \dfrac{h\nu\Gamma N_2 + \alpha_g P_{N0}}{(N_2 - N_1)\Gamma - \alpha_g}}{P_{N1} + \dfrac{h\nu\Gamma N_2 + \alpha_g P_{N0}}{(N_2 - N_1)\Gamma - \alpha_g}} = \exp\{[(N_2 - N_1)\Gamma - \alpha_g]l\}$$

where $P_{N2}$ is the noise power out of the amplifier, $P_{N1}$ is the noise power input, and $l$ is the amplifier length. Similarly integrating (21) for the signal power, the power gain is

$$\text{Gain} = \frac{P_{S2}}{P_{S1}} = \exp\{[(N_2 - N_1)\Gamma - \alpha_g]l\}.$$

The gain may be written as

$$\text{Gain} = \exp[(\alpha - \alpha_g)l]$$

where $\alpha$ is the "gain coefficient."

Define

$$M \equiv \frac{h\nu\Gamma N_2 + \alpha_g P_{N0}}{(N_2 - N_1)\Gamma - \alpha_g}.$$

Then

$$\text{Gain} \equiv G = \frac{P_{N2} + M}{P_{N1} + M}$$

or

$$\frac{P_{N2}}{P_{N1}} = G + (G - 1)\frac{M}{P_{N1}}.$$

The noise figure $F$ of the amplifier is defined by

$$F \equiv \frac{P_{N2}}{GP_{N1}} \quad \text{for} \quad P_{N1} = P_{N0}.$$

Thus,

$$F = 1 + \left(1 - \frac{1}{G}\right)\frac{M}{P_{N0}}.$$

$M$ can be put in the form

$$M = \frac{h\nu}{1 - \dfrac{N_1}{N_2}}\left(1 + \frac{\alpha_g}{\alpha - \alpha_g}\right) + \frac{\alpha_g P_{N0}}{\alpha - \alpha_g}$$

and

$$F = 1 + \left(1 - \frac{1}{G}\right)\left[\frac{h\nu}{P_{N0}}\,\frac{\left(1 + \dfrac{\alpha_g}{\alpha - \alpha_g}\right)}{\left(1 - \dfrac{N_1}{N_2}\right)} + \frac{\alpha_g}{\alpha - \alpha_g}\right].$$

In the case $G \gg 1$, this can be factored into two terms:

$$F = \left[1 + \left(\frac{N_2}{N_2 - N_1}\right)\frac{h\nu}{P_{N0}}\right]\left[1 + \frac{\alpha_g}{\alpha - \alpha_g}\right]. \qquad (23)$$

The first term on the right is the contribution to the noise figure due to spontaneous incoherent emission, the second, the contribution of the effects of black-body radiation in the guide. The two terms are closely related, however as the gain coefficient $\alpha$ and the term

$$\left(\frac{N_2}{N_2 - N_1}\right)$$

both depend on the "excess population" in the upper state produced by the excitation process. This expression also shows clearly that losses are vitally important: unless $\alpha \gg \alpha_g$, the amplifier will have a poor noise figure.

It is also clear that a large class of "paired off" molecules adversely affects the noise figure by increasing $N_2$ without increasing $N_2-N_1$. However, if the gain coefficient $\alpha$ greatly exceeds the attenuation coefficient $\alpha_0$ and there are few molecules $N_1$ in the lower state,

$$F \approx 1 + \frac{h\nu}{P_{N0}} \approx 1 + \frac{h\nu}{kT_0}$$

which is of the order of 1.004, or a noise figure of 0.02 db.

### F. Efficiency

In general, the efficiency of molecular amplifiers is low. If the over-all efficiency is defined as the ratio of the increase of microwave power in the amplifier to the total power expended in associated electronic circuits, etc., as well as in the molecular system, the efficiency is invariably low (well under one per cent). The reasons for this vary with the different means of exciting the molecular systems, and hence can only be properly discussed in connection with an excitation method. However, a few general considerations will indicate that high efficiency is not to be looked for in molecular amplifiers.

Roughly speaking, energy must be supplied to each molecule at a rate given by

$$P_{\text{in}} \approx \frac{h\nu}{T_1}$$

where $\nu$ is the amplification frequency. The power transferred to the radiation field from the molecular system is, from (12),

$$P_{\text{out}} \approx \frac{2\pi^2\nu p^2 E^2 T_2}{h}.$$

The efficiency is thus of the order of

$$\text{Efficiency} = \frac{P_{\text{out}}}{P_{\text{in}}} \approx 2\pi^2 T_1 T_2 \left(\frac{pE}{h}\right)^2.$$

The condition for linear amplification is that (10)

$$\left(\frac{pE}{h}\right)^2 \ll \frac{1}{4\pi^2 T_1 T_2}.$$

So

$$\text{Efficiency} \ll 1/2.$$

From this it is seen that the requirement of linear amplification severely limits the efficiency that can be expected.

## IV. Practical Approaches

Having discussed molecular amplifiers in general, various devices that apply the idea of emissive molecular interaction to solve the practical problems of low-noise amplification, the generation of microwaves, and production of frequency standards of high constancy ("atomic clocks") will now be considered.

### A. Molecular-Beam Maser

The various devices developed as molecular amplifiers differ mainly in the way in which an emissive condition is obtained. A discussion of amplifiers in a sense thus becomes a discussion of excitation methods. One well-known type of molecular amplifier is the ammonia-beam Maser.[1,19] Here the excitation method consists of physically separating a beam of gas molecules into two beams, one containing the lower state, and one the upper state emissive molecules. After the separation, the emissive beam is used in an amplifier cavity.

The Maser is shown schematically in Fig. 8(a). The gas molecules issue from the source (a chamber maintained at an ammonia pressure of a few millimeters of mercury) as a well-collimated beam, the directivity being obtained by the means of a source opening consisting of channels which are long compared to their diameter and short compared to the mean free path of the gas. This beam travels first through a focuser [Fig. 8(b)] and finally into the resonant cavity. The focuser removes from the beam the molecules in the lower of the two energy states of interest, leaving in the beam entering the cavity only upper state molecules.

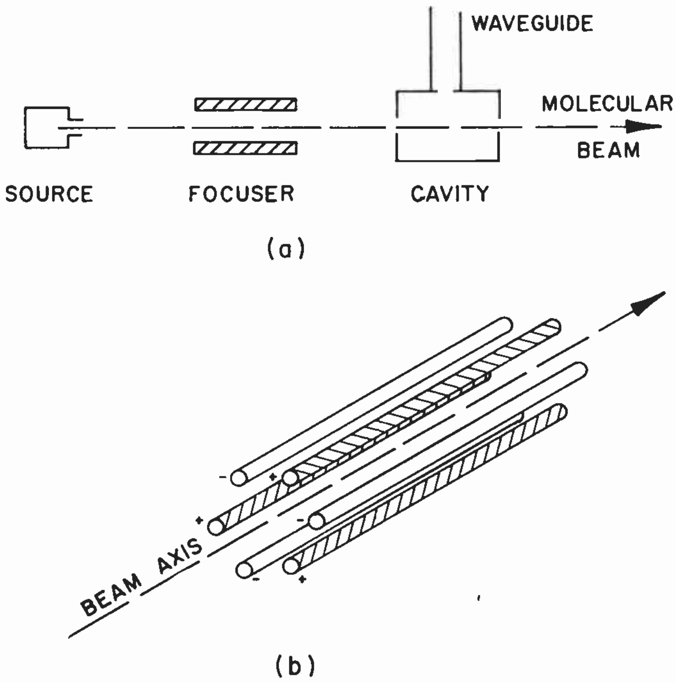


Fig. 8—(a) Schematic diagram of a molecular beam Maser. (b) A perspective view of the focuser.

The action of the focuser can be understood by reference to Fig. 9. This shows how the energies of the internal states of the molecules vary in an electric field; it is

[19] J. P. Gordon, "Hyperfine structure in the inversion spectrum of $N^{14}H_3$ by a new high-resolution microwave spectrometer," *Phys. Rev.*, vol. 99, pp. 1253–1263; August 15, 1955.
K. Shimoda, T. C. Wang, and C. H. Townes, "Further aspects of the theory of the Maser," *Phys. Rev.*, vol. 102, pp. 1308–3121; June 1, 1956.
J. C. Helmer, "Theory of a Molecular Oscillator," Microwave Lab. Rep. no. 311, Stanford Univ., Stanford, Calif.; June, 1956.

seen that the energy of the upper state increases in a field, and the energy of the lower state decreases. Because of this (quadratic) Stark effect on the energy levels, a molecule placed in a region where the electric field has a gradient will experience a force; this will be toward the low field region for a molecule in the upper energy state, but toward the high field region for a lower state molecule. An inhomogeneous field region thus acts on the original beam of molecules so as to split it into two physically separated beams, a focused one consisting of upper-state molecules and a diffuse, deflected one of lower-state molecules. The inhomogeneous field is produced in the focuser by a series of long, thin electrodes, parallel to the beam axis and surrounding it [see Fig. 8(b)]. Alternate electrodes are charged to high and low potentials: this produces zero field along the axis and high fields near the electrodes. The upper state molecules in the beam thus experience a focusing force toward the beam axis, while the lower state ones are defocused out of the beam. Clearly, only molecules having this type of Stark effect can be used in the Maser.

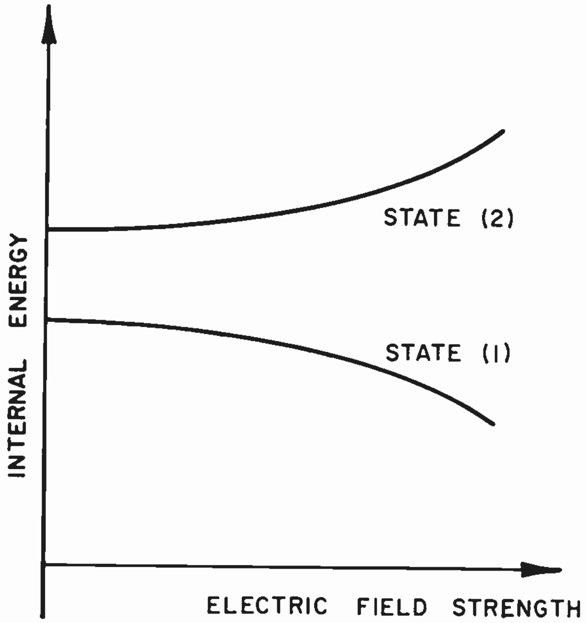


Fig. 9—Stark effect in an ammonia molecule.

If the power emitted by the molecules exceeds the losses in the walls, the device can act as an amplifier. If the power emitted exceeds the sum of wall losses and the power coupled out of the cavity into the connecting waveguides, oscillations will build up until saturation effects make the molecularly emitted power equal to all losses. The power emitted can be controlled by varying the beam density. Since it is possible to get beam densities high enough to sustain oscillations, the gain as an amplifier can be made essentially infinite, with, of course, a corresponding decrease of the amplification bandwidth to practically zero.

The bandwidth of the Maser is determined by a process not previously discussed. Since the beam is tenuous, gas collision broading is absent, and if the $TM_{010}$ mode in a cylindrical cavity is used, there is no axial variation in the microwave fields and Doppler effects are also missing. The coherent interaction time is then determined by the transit-time; *i.e.*, the time it takes for a molecule to travel the length of the cavity. The "molecular bandwidth" is of the order of[20]

$$(\text{Molecular bandwidth}) \approx \frac{1}{T_2} \approx \frac{v}{l},$$

where $l$ is the cavity length and $v$ the mean molecular velocity.

In a molecular-beam Maser, the internal energy emitted as microwave energy is acquired during the collisions of the molecules with the walls of the source chamber whereby they come into thermal equilibrium with this chamber. The kinetic energies of motion of the molecules arise in the same way. If the gas admitted to the source chamber is already at the temperature of the chamber, there is no further transfer of energy to the gas from the chamber.

When the Maser is acting as a linear amplifier, only a small fraction (a few per cent) of the total internal molecular energy is transformed to microwave energy, and hence the Maser efficiency is low. Acting as an oscillator, some nonlinearity of emission is inherent, and the efficiency can be somewhat higher. The kinetic energy of the molecules is not converted to microwave energy, but is lost when the "spent" gas is pumped away after traversing the amplifier cavity. If this is included in calculating the efficiency, the efficiency is low indeed, tas the kinetic energy of a molecule leaving a source a room temperature is typically hundreds of times its total internal energy available for conversion to microwaves.

One gas suitable for use in a Maser is ammonia: here the "inversion" transitions provide suitable pairs of levels for focusing. The strongest set of levels has a transition frequency at 23.870 kmc (1.26 cm). Cavities roughly 10 cm long give a transit-time bandwidth of about 6 kc; the amplification bandwidth is somewhat smaller, depending on the amplifier gain. Assuming a loaded cavity $Q_L = 5000$ and unloaded $Q_0 = 10{,}000$, the saturation condition for linear amplification (11) indicates that the incident signal power must be

$$P_{\text{inc}} < 10^{-10} \text{ watts}.$$

Another important property of the Maser is its noise figure. However, it is extremely difficult to measure noise figures as low as expected for such an amplifier, and no conclusive measurements have been made. The measurements made indicated a Maser noise figure of about one; *i.e.*, zero db.

[20] A suitable average over the distribution of velocities must of course be taken, but this only introduces a numerical factor of order one.

As an oscillator, the Maser has the very high spectral purity to be expected of a very narrow band amplifier. (The molecular bandwidth is about 3 parts in $10^7$ of the central frequency, giving an amplification bandwidth of only a few kilocycles.) Experimentally,[1] a spectral purity of about *4 parts in $10^{12}$* has been obtained for a period of the order of one second. The power generated by the device when oscillating was estimated to be a few times $10^{-10}$ watts.

The questions arise: what can be done to improve various operating characteristics, or, alternatively, to what extent can performance in one respect be enhanced at the cost of worsened behavior in another characteristic? As an amplifier, the main characteristics desired are high gain, broad bandwidth, low noise figure, and high power handling capabilities. As an oscillator, spectral purity and power are wanted. High gain in the Maser is easily assured. The amplification bandwidth can be increased if the transit time is reduced. This is most readily accomplished by shortening the cavity in the direction of beam travel. However, unless the beam geometry is altered, this reduces the number of molecules in the cavity at any one time and hence reduces the power emitted by the molecules. Unless the cavity losses are cut proportionately, the available gain is thereby reduced and the noise figure increased. One way to maintain a high number of molecules in a shortened cavity is to use multiple beams, but this introduces serious problems of pumping and obtaining a high cavity $Q$. Gain can also be traded for increased bandwidth by increasing the coupling of the cavity to the waveguide (reducing the loaded $Q_L$). The bandwidth can be increased by either method without loss of gain if the beam flux can be increased. Oscillator power can also be increased by increasing the flux. Considerable effort is being made to improve gas sources to increase beam densities.

There are several other practical problems in connection with the Maser that are receiving attention. Improvement of focusing methods is one. Also, since molecules must pass from the source through the focuser and cavity without colliding with other molecules, a high vacuum is needed, and various improved pumping schemes are under consideration. With its high spectral purity and its frequency basically determined by molecular properties, the Maser should provide a good atomic frequency standard. However, the exact frequency of oscillation can be "pulled" slightly by the resonant cavity or by stray fields. Various methods of eliminating or allowing for such pulling effects are being studied. Finally, the question of the tunability of a Maser amplifier arises. As seen above, the energy separation of the two states, and hence the molecular transition frequency, can be altered by an electric field. A magnetic field can have a similar result. However, problems of uniformity of tuning field (and degeneracy-splitting effects) make it practically impossible to change the resonance frequency more than a megacycle by these methods; the molecular beam Maser is essentially untunable.

### *B. Hot Grid State Separator*

Another type of molecular amplifier is the "hot grid cell" devised by Dicke.[21] The way a gas is maintained in an emissive condition in this device can be seen with reference to Fig. 10. Fig. 10(a) is a schematic cross section of one configuration of such a device. The microwave structure consists of a plane parallel plate waveguide, the plates being labeled "cold wall" and "hot wall" for a reason that will shortly become clear. The microwave fields are assumed to be in the form of a wave propagating between these conducting plates and into the plane of the paper. Adjacent to the hot wall is a grid of fine wires parallel to the hot wall and each other, and maintained at a high voltage with respect to the hot wall. This produces a region of intense electrostatic field around the wires and along the hot wall. In the following, it is assumed that the grid is completely transparent to gas molecules, to simplify the discussion. This clearly is only an approximation. However, by heating the grid structure itself, it becomes an "equivalent hot wall" and itself behaves as the actual hot wall. The region between the walls is filled with a gas, such as ammonia, that has a Stark effect as shown in Fig. 6, and is at a pressure such that the mean free path for gas-gas collisions is several times the distance between walls. Such collisions can thus be neglected. Fig. 10(b) shows the distribution of electric field strength in the cell; Fig. 10(c) is a plot of internal energy of a gas molecule as a function of position in the cell for molecules in the upper and lower of the two energy levels of interest. It is seen that in the bulk of the cell, between the cold wall and grid, there is no electric field (other than the microwave field) and the two energy levels are separated by the usual amount, $\Delta W = h\nu_0$, where $\nu_0$ is the frequency of the microwave field at resonance.

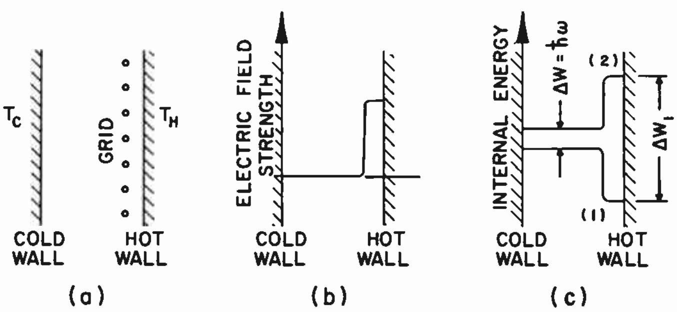


Fig. 10—Temperature, electric field strength, and internal energy as a function of position across a hot grid separation cell.

Gas molecules that strike the cold wall become "thermalized" at its temperature $T_C$. This means that they leave the wall with velocities characteristic of a Max-

[21] R. H. Dicke, private communication.

wellian distribution at $T_C$, and with a distribution between the two internal energy levels given by (4), with $T = T_C$. Thus the class of molecules moving toward the right after striking the cold wall contains more molecules in the lower energy state than the upper, and has a net absorptive effect. As the molecules move toward the hot wall, those in the upper energy state see a repulsive barrier in the region of high electrostatic field and tend to be reflected back toward the cold wall. Molecules in the lower state are attracted into the potential energy well along the hot wall [Fig. 10(c)] and hence strike the hot wall (or hot grid structure).

The height of the repulsive barrier seen by the upper state molecules depends on the strength of the electric field; however, with a Maxwellian distribution of velocities, some upper state molecules will possess sufficient kinetic energy to surmount the barrier and strike the hot wall, regardless of barrier height. The molecules that are reflected back to the cold wall form a second, smaller class of molecules; this class, composed as it is of upper state molecules only, is strongly emissive.

A third group of molecules consists of those leaving the hot wall and returning to the cold wall. These molecules leave the hot wall in equilibrium with it at its temperature $T_H$. However, the energy difference $\Delta W_1$ between the two states in the strong field at the wall is much greater than $\Delta W = h\nu_0$, and so, by (4), many more molecules leave the hot wall in the lower energy state than in the upper. Now it is the lower state molecules that have to surmount a Stark field barrier to return to the cold wall, while the field accelerates the upper state molecules in their return journey. The net result is that this third group of molecules, containing both upper and lower state molecules, is absorptive; *i.e.*, has a higher density of lower than of upper state molecules in it.

The net effect of these three groups of molecules depends on $T_C$, $T_H$ and the height of the electrostatic field barrier. If $T_C = T_H$, the gas in the field-free region between the "cold" wall and grid is in thermal equilibrium with the walls, and hence is absorptive, regardless of the high-field region. However, if the temperature of the cold wall is lowered, $T_C < T_H$, two effects occur. One is that the proportion of molecules in the lower state leaving the cold wall increases, by (4). This is a relatively unimportant effect. Lowering $T_C$ has the more significant result that the kinetic energies of the molecules leaving the cold wall are reduced, and more of these slower molecules are reflected at the field barrier, increasing the number of molecules in the emissive, reflected class. Raising the hot wall temperature $T_H$ has a similar result, for different reasons. By increasing $T_H$, more molecules leaving the hot wall are in the upper energy level [from (4)]. There is also the result that the molecules returning from the hot wall are speeded up by raising $T_H$, and since this is an absorptive class of molecules, by thus decreasing their transit time to the cold wall, their density, and hence absorptive effect, is reduced for a constant molecular flux. The net result can be that, for suitable $T_C$, $T_H$, and field barrier height, the gas is maintained in an emissive condition in the main part of the cell.

An exact analysis of such a grid cell is exceedingly difficult, as it depends on suitable averages to determine such parameters as the "effective field barrier height," a spatial average over the inhomogeneous field produced by the grid structure. However, estimates indicate that the "excess population," $(N_2 - N_1)$, can be of the order of one per cent of the total number of molecules for temperatures of the order $T_C \approx -100°C$, $T_H \approx 200°C$, and fields produced by appropriate grid structures at 15–20 kv voltages. This corresponds to a gain coefficient (13) of the order of $\alpha \approx 10^{-3}$ cm$^{-1}$. This is too small to make a nonresonant waveguide amplifier practical: an amplifier many meters long would be required to produce useful gain. It is nevertheless large enough to make a low-noise amplifier possible using a resonant cavity type of microwave structure; as seen above, with a cavity amplifier, the gain can be made arbitrarily large (with an unavoidable loss of bandwidth).

It might be expected that transit time effects across the cell would determine the amplification bandwidth. This however is not the case. For satisfactory operation of the cell, a rectangular cell cross section is desirable. To make a very low loss cavity with this geometry, one several wavelengths long is required. In this case, Doppler effects cannot be neglected, and indeed provide the dominant broadening mechanism. For ammonia as the active gas, the Doppler-determined molecular line breadth is of the order of 50 kc, about an order of magnitude larger than for a beam-type Maser. This greater bandwidth is accomplished without loss of gain, despite the small fractional upper state enrichment, by the much higher molecular densities achieved with the hot-grid cell compared with the molecular beam Maser.

The Maser, with its narrower bandwidth, is expected to be much more stable as an oscillator than the hot-grid cell. Thus, for frequency standard applications the beam type of device is to be preferred. On the other hand, the hot-grid cell can be operated as a sealed-off device without the continual supply of gas and its subsequent removal as in a beam amplifier.

The efficiency of a hot-grid cell is low. In addition to the loss taken to preserve linear operation, the thermal losses as the molecules travel from hot wall to cold are enormous relative to the microwave energy conversion. (Here again, as in the beam Maser, the internal energy comes from the thermal energy of the cell walls.)

An amplifier operative on the hot-grid state separation principle is under active investigation at RCA Laboratories at the present time.

### C. Pulse Inversion

One of the methods of achieving an emissive state discussed above is essentially a continuous process, *i.e.*, the hot-grid cell. The other, molecular-beam separation of states, while separating the exciting region from the utilization region, nonetheless uses a continuous flow of gas to achieve effectively cw operation. There are, however, other means of exciting a molecular system that are distinctly pulse-operation approaches. "Pulse inversion of states" is one of these. From (6) it follows that if a system is exposed to a pulse of microwave power on resonance ($\nu=\nu_0$) of such strength and duration that

$$\frac{pET}{h}=\frac{1}{2}, \tag{24}$$

molecules originally in the lower state will be raised to the upper state and vice versa; *i.e.*, the state populations will be *inverted*. Thus, such a pulse transforms an absorptive system into an emissive one. The pulse duration must be short compared to the relaxation times, or complete inversion is not achieved. This is not a serious limitation, however, as with readily available powers, pulses of under a microsecond duration satisfy (24). Two conditions are critical in this method of obtaining an emissive state: 1), if the pulse frequency is off resonance, complete inversion will not be obtained [see (6)], and 2), the product of pulse strength and duration must be accurately adjusted.

An amplifier can be made by putting a molecular system in a suitable microwave structure such as a resonant cavity and exciting it with an inverting pulse. For a period of time given by the relaxation processes, the cavity with its emissive loading will amplify microwave signals incident upon it in a frequency bandwidth corresponding to the gain bandwidth of the amplifier.

This excitation method offers the possibility of relatively high efficiency, on a molecular basis, if a system with $T_1 \gg T_2$ is used, and some time-dependence of the amplifier gain is allowed. Apart from resistive losses in the microwave structure, much of the energy supplied in the excitation pulse can be re-extracted in the amplification period. However, if the power requirements of the electronic devices needed to generate and control the inverting pulses are included in the calculation, even this excitation scheme results in low efficiency.

Such a device has obvious faults as an amplifier. It only amplifies for a fraction of the time, and is nonamplifying in the excitation and the "thermalization" period in which equilibrium is again attained before the next inverting pulse. The gain is therefore periodically varying. If the device operates at room temperature, the noise figure cannot be better than 3 db, from (23). This can be improved by going to low operating temperatures; at liquid helium temperatures, the minimum noise figure is under 0.1 db. A serious fault of this method of achieving state-population inversion is that it depends on a critical adjustment of frequency and pulse strength and duration. Means of overcoming these difficulties will be discussed in connection with a second pulsed method of obtaining an emissive state.

### D. Inversion by Adiabatic Fast Passage

Bloch[14] has shown that spin systems in a static magnetic field can be inverted by what is called *adiabatic fast passage*. His results can be generalized[22] to show that state populations in an arbitrary two-level system can be inverted by a similar technique. In adiabatic fast passages, the molecular system is subjected to a strong microwave field of amplitude $E$ and variable frequency. The frequency of this field starts far off resonance and is slowly swept through the resonant frequency until it is far off resonance on the other side. When this has been done, the state populations are inverted. Three conditions must be met for this to occur:

1) The passage must be adiabatic; that is, the frequency must be changed slowly compared to the internal motions of the molecule in response to the driving field $E$. This condition can be expressed as

$$\frac{d\nu}{dt} \ll \frac{pE}{h}.$$

2) The passage must be fast compared to the relaxation time; the time $\tau_s$ required to sweep the frequency from one side of resonance to the other must be short compared to the relaxation time $\tau_r$ of the system.

$$\tau_s \ll \tau_r.$$

3) The driving field $E$ must be larger than the maximum radiation field of the system. During the passage through resonance, the molecular system develops an oscillating dipole moment; the maximum radiation field of this moment must be smaller than the driving field or population inversion cannot be achieved.

A physical explanation can be given for each of these requirements. In an adiabatic fast passage, the molecular system goes through a series of quasi-stationary states. If the sweep rate is too large, the system cannot follow the changes adiabatically, and nonstationary states, such as are responsible for pulse inversion, are induced. The second condition merely assures that the inversion is completed before the competing relaxation processes can restore the system to thermal equilibrium. The third condition arises because the radiation field and driving field become out of phase as they approach each other in magnitude, and at equality of magnitudes, they just cancel each other in their effects.

Two other points of importance should be mentioned. One is that it does not matter in which direction the frequency sweep traverses the resonance: the initial frequency can be either above or below the resonant frequency. The second is that, if the resonant frequency of the molecular system can be altered by the application or change of an applied electric or magnetic field, for example, the inversion can be achieved by keeping the frequency of the applied microwave field fixed and *sweeping the resonant frequency* from one side of the applied frequency to the other.

[22] S. Bloom, private communication.

Adiabatic fast passage has two major advantages over pulse inversion as a method of achieving an emissive state: it is not necessary to control the frequency of the microwave exciter source accurately, and the exact duration and time dependence of the frequency sweep are similarly uncritical. The other difficulties of pulse inversion enumerated above are also met in adiabatic fast passage. However, as mentioned earlier, most, if not all, are susceptible to amelioration. Before discussing this, the characteristics desirable in a molecular system excited by adiabatic fast passage will be considered.

From (13), the gain coefficient $\alpha$ increases with the number of "excess" molecules $N$ and the dipole moment $p$, and decreases with increasing molecular bandwidth as

$$\alpha \sim \frac{p^2 N}{(\text{Bandwidth})}.$$

Since amplification bandwidth is at a premium, it is clear that systems with a large "excess" population are desirable: if state population inversion is to be used, a large excess at thermal equilibrium is wanted. From (5a), $N_1 - N_2 = N \tanh (h\nu/2kT)$, a large excess can be obtained in two ways, by increasing the molecular density or by going to lower temperatures.

Low temperatures are also desirable from a noise figure point of view, as is seen by considering (23). Neglecting guide losses,

$$F_{\min} = 1 + \left(\frac{N_2}{N_2 - N_1}\right)\frac{h\nu}{P_{N0}}.$$

Starting from a state of thermal equilibrium at a temperature of the molecular system $T_{\text{molec}}$, after state inversion

$$\frac{N_2}{N_1} = \exp\left(+\frac{h\nu}{kT_{\text{molec}}}\right)$$

and

$$\frac{N_2}{N_2 - N_1} \approx +\frac{kT_{\text{molec}}}{h\nu}.$$

The noise figure is referred to room temperature $T_0$,

$$P_{N0} \approx kT_0$$

and

$$F_{\min} = 1 + \frac{T_{\text{molec}}}{T_0}. \tag{25}$$

Thus noise figures of under 3 db can only be achieved by lowering the temperature of the molecular system: at dry ice temperatures, $T_{\text{molec}} = 194°\text{K}$, $F_{\min} = 2.3$ db, at liquid nitrogen temperatures, $T_{\text{molec}} = 77°\text{K}$, $F_{\min} = 1.1$ db, while at liquid helium temperatures $T_{\text{molec}} = 4°\text{K}$, $F_{\min} = 0.06$ db.

The desirability of both low temperatures and high densities points toward use of a solid molecular system. However, $\alpha \sim p^2$ argues for a gas, where electric dipole transitions with $p \sim 10^{-18}$ are available, and against a solid, where only paramagnetic resonance for which $p \sim 10^{-20}$ appears to provide sufficiently long relaxation times. Relaxation times in a gas of useful density are short, however, making the sweep technique of adiabatic fast passage very difficult to apply. Turning then to a solid (or liquid), the problem of finding a molecular system with a usably long relaxation time arises. Fortunately, systems are known with spin-lattice relaxation times of the order of many seconds.[23] These are systems of paramagnetic electrons localized on donor atoms in a silicon lattice. The donor atoms whose relaxation times have been measured include Group V elements (P, As, Sb) and lithium. The relaxation times vary with temperature and donor concentration, but for concentrations less than $10^{17}$–$10^{18}$ donors per cubic centimeter and at liquid helium temperatures, $T_1$ is of the order of a minute. Thus suitably doped silicon at liquid helium temperatures presents a high density, long relaxation time material suitable for making a low-noise amplifier[24] excited by adiabatic fast passage.

A possible amplifier using adiabatic fast passage excitation might consist of a doped-silicon sample in a microwave cavity in a helium cryostat located between the poles of an electromagnet. (See Fig. 11 opposite.) The time variation of the (dc) magnetic field $H$ for such an amplifier is shown in Fig. 12. At point $A$, the sample is in thermal equilibrium with its surroundings in a magnetic field $H_A$. From $A$ to $B$, the magnetic field is swept through $H_R$, which is the resonant field at the exciter microwave field frequency. This adiabatic fast passage inverts the spin state populations, leaving the sample in an emissive condition. From $B$ to $C$, the field is kept constant at $H_B$, and the device acts as an amplifier of microwaves at the frequency

$$\nu = \gamma H_B = 2.8 H_B \text{ mc}.$$

At $C$, the field is abruptly returned to $H_A$, reaching $H_A$ at $D$. Amplification at the frequency $\nu$ ceases, and the system is allowed to come to thermal equilibrium with its surroundings in the field $H_A$. Thermalized again at $A'$, the cycle is repeated. This cycle produces a pulsed amplifier, with an amplification period from $B$ to $C$ and a dead-time, during which it is re-excited, from $C$ to $B'$.

At first thought, it would seem that the dead-time $C$–$B'$ would greatly exceed the active time $B$–$C$, since a time $D$–$A'$ long compared to $T_1$ is required for the

[23] G. Feher and R. C. Fletcher, "Relaxation effects in donor spin resonance experiments in silicon," *Bull. Amer. Phys. Soc.*, vol. 1, p. 125; March 15, 1956 (Pittsburgh Meeting).

[24] J. Combrisson, A. Honig, and C. H. Townes, "Utilisation de la résonance de spins électroniques pour réaliser un oscillateur ou un amplificateur en hyperfréquences," *Compt. Rend.*, vol. 242, pp. 2451–2453; May 14, 1956.

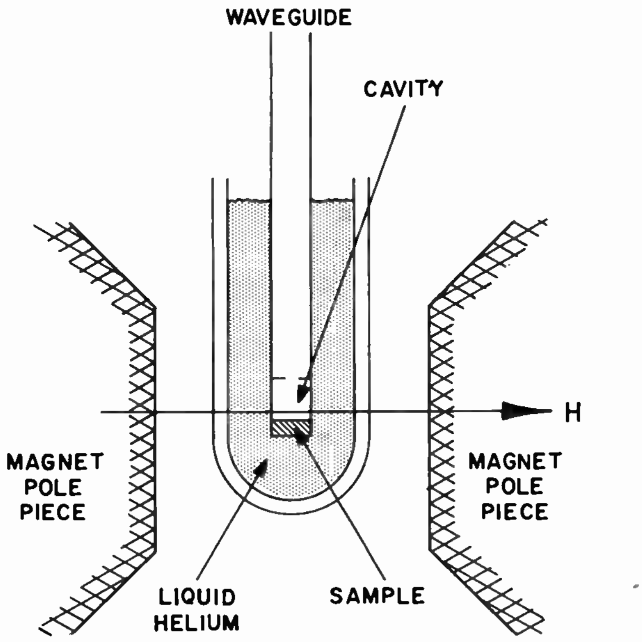


Fig. 11—Low-temperature paramagnetic amplifier.

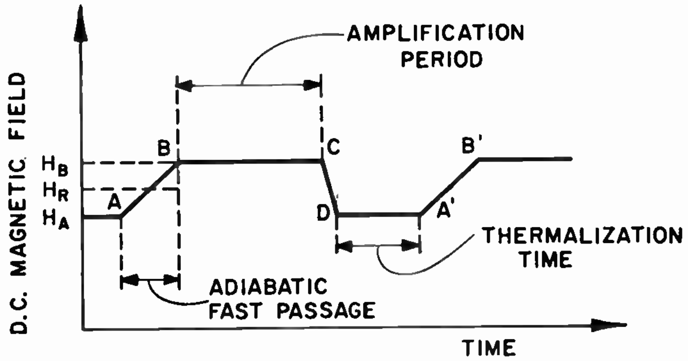


Fig. 12—Magnetic field vs time for a paramagnetic amplifier excited by adiabatic fast passage. $H_B$ is the field during the amplification period, $H_R$ the field at which the molecules are resonant at the exciter field frequency, and $H_A$ the field at which thermalization occurs.

sample to reach thermal equilibrium. Fortunately, however, with a doped-silicon sample something can be done to shorten $D$–$A'$ drastically. The long relaxation time ($\sim$1 minute) of this material at liquid helium temperatures is associated with the fact that the paramagnetic electrons are bound in the Coulomb-type potentials surrounding the donor atoms and that there are no free (conduction) electrons in the sample. If the sample (at liquid helium temperature) is illuminated with visible light, free carriers are released and $T_1$ is reduced to the order of a microsecond. A flash of light can thus shorten the required time for thermalization $D$–$A'$ to a few microseconds. With an adiabatic sweep time $A$–$B$ of a few milliseconds and an amplification time $B$–$C$ of several seconds, it is seen that a nearly continuously amplifying device can be made. (It should be mentioned that by exciting the sample externally to the amplifier cavity, and providing a continuous "flow" of excited material through the amplifier, a truly steady state, or continuous amplifier can be obtained.)

The power available from the molecular system after excitation is given by

$$P_{\text{emission}} = \frac{4\pi^2 N p^2 \nu T_2 H_{rf}^2}{h},$$

where $N$ is the excess number of spins in the upper state. The spectral line width is about

$$(\text{Bandwidth}) \approx \frac{1}{\pi T_2}.$$

Making the resistive losses in the cavity less than the emitted power by using high spin densities, using samples with sufficiently long $T_2$, and using a high-$Q$ cavity, a high-gain cavity amplifier can be obtained.

The molecular bandwidth associated with the doped-silicon samples discussed above is several megacycles, indicating a $T_2$ of less than a microsecond. This is radically lower than $T_1 \approx 60$ seconds. The difference is caused by magnetic interactions between the paramagnetic spins, and between these spins and other magnetic dipoles in the sample (nuclear moments). These interactions dephase the electron spin-radiation field interaction, and hence broaden the resonance, but do not provide a means of exchanging energy with the crystalline lattice, and hence do not shorten $T_1$. By removing some of the dephasing nuclear moments and replacing them with nonmagnetic isotopes, $T_2$ can be increased and hence the molecular emission power increased, at, of course, a sacrifice of some amplification bandwidth.

The energy input in adiabatic fast passage comes from two sources: the high-level microwave exciter field, and (possibly) the energy associated with the magnetic moment of the system if the applied field is varied. With available cavity $Q$'s, the losses from the exciter field are far greater than the microwave energy obtained in amplification, and so the efficiency of this system is low.

### E. Multilevel Excitation Methods

The methods of achieving an emissive state discussed thus far have dealt with molecular systems where only two energy levels were significant.[25] There are other ways of obtaining an emissive state if one considers multilevel schemes and admits the possibility of one or more exciter frequencies different from the amplification frequency. Several possibilities for three-state systems

[25] In these two-level systems, other levels are of practical importance in determining the fraction of the total molecules in the system in the two levels of interest.

have been discussed by Basov and Prokhorov.[26] One basic approach with a three-state system can be seen by reference to Fig. 13. In one embodiment [Fig. 13(a)], the upper pair of states 2) and 3) are separated by the microwave frequency desired to be amplified, $\nu_{23}=\nu_{sig}$. Energy states 1 and 3 are also coupled by a radiation field, of frequency $\nu_{13}=\nu_{ex}$. At thermal equilibrium, there are more molecules $N_{1eq}$ in state 1 than in either state 2, $N_{2eq}$, or state 3, $N_{3eq}$; in fact,

$$N_{1eq} > N_{2eq} > N_{3eq}.$$

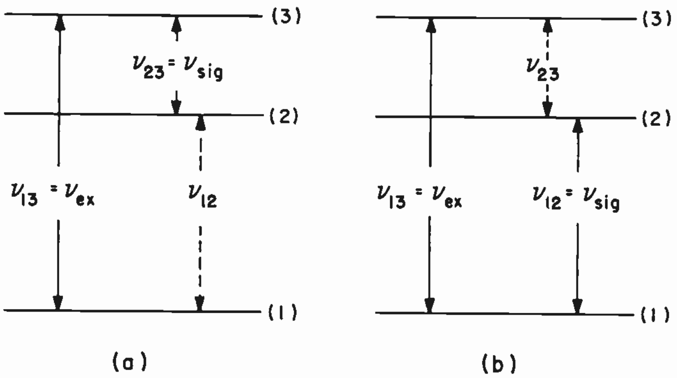


Fig. 13—Energy level diagram illustrating multilevel excitation methods.

If the resonance at $\nu_{ex}=\nu_{13}$ is saturated, however, this situation is altered. Here "saturated" means that the radiation field at frequency $\nu_{13}$ is so large that the effects of induced transitions completely override relaxation effects and the two states become equally populated. Saturating the transition at $\nu_{13}$ results in increasing the population $N_3$ of state 3. If this results in $N_3>N_2$, a steady state emissive condition is obtained. It must be noted that relaxation effects, despite the strong rf field at $\nu_{13}$, may be very important in the final effective state populations attained. For example, if the relaxation mechanisms bringing the state populations $N_3$ and $N_2$ into thermal equilibrium are very weak, while those thermalizing $N_1$ and $N_2$ are strong, as state 1 is depopulated by the radiation, the relaxation mechanisms will transfer molecules from state 2 to state 1, in an attempt to maintain the equilibrium $N_2/N_1$ ratio. This will make the effective $N_2$ less than that corresponding to a condition of thermal equilibrium for the whole sample. It is clear that with a complicated system of levels and relaxation effects, many variations on this theme are possible. An obvious one is shown in Fig. 13(b): here the $\nu_{13}$ transition is again saturated, but a fast relaxation bebetween states 3 and 2 together with a weak relaxation mechanism between states 1 and 2 now makes $N_2>N_1$ and amplification occurs at a frequency $\nu_{sig}=\nu_{12}$.

[26] N. G. Basov and A. M. Prokhorov, "Possible methods of obtaining active molecules for a molecular oscillator," *J. Exper. Theoret. Phys. USSR*, vol. 28, pp. 249–250; February, 1955.

This general method of achieving an emissive molecular system can have the exceedingly desirable feature of combining the high molecular densities available in a solid with the continuous excitation of the gas devices discussed, by using a solid with appropriate energy levels and relaxation times. Some practical problems are posed by this excitation method, however. Since excitation is a continuous process, the same cavity must be used for both excitation and amplification. This means a cavity simultaneously operating in two modes, at two frequencies. Furthermore, the modes must be chosen so that, insofar as possible, regions of strong field for the excitation processes and regions of strong interaction for the amplification process coincide. This requirement cannot be met completely, of course, due to cavity mode orthogonality. However, with small sample sizes, this is a minor problem. If the amplifier operates at liquid helium temperatures, the heating effects of the saturating exciter field must also be considered. As in excitation by adiabatic fast passage, the presence of a strong exciter field with its relatively high losses makes multilevel excitation methods have low efficiency.

Multilevel methods of excitation open up another possibility of achieving low-noise amplification without resorting to low temperatures. Neglecting guide losses, the noise figure can be written (25),

$$F = 1 + \frac{T_{molec}}{T_0},$$

where $T_{molec}$ is the "internal temperature" of the molecular system, defined by (2). With excitation methods that exchange, or invert, state populations, a $T_{molec}<T_0$ can only be achieved by physically cooling the molecular system ($T_0$ is *defined* to be room temperature). With the multilevel excitation scheme of Fig. 13, however, if the spin-lattice relaxation time $T_{12}$ coupling states 1 and 2 is very much shorter than the relaxation time $T_{23}$ coupling states 2 and 3, the internal temperature for the amplifier resonance between 2 and 3 is given by

$$T_{molec} = \left(\frac{\nu_{23}}{\nu_{12}}\right) \times (\text{Actual amplifier temperature}).$$

If $\nu_{23}$ is sufficiently below $\nu_{12}$, the actual amplifier temperature can be room temperature and low-noise figures can still be obtained.

A major problem associated with this multilevel excitation method is in finding a suitable molecular system. The requirements are rather stringent: 1) the system must have three levels as shown in Fig. 13, with separations so that $\nu_{ex}$ occurs at a frequency where oscillators of reasonable power exist, and so that $\nu_{sig}$ is a desirable amplification frequency; 2), the selection rules must permit the transitions at $\nu_{13}$ and $\nu_{23}$ to be induced by radiation; 3), the relaxation time $T_{13}$ must be long enough to permit saturation of the excitation resonance; 4), the relaxation time $T_{23}$ must be considerably greater than $T_{12}$; 5), molecular densities must be high enough so that the molecularly emitted power exceeds the losses at the signal frequency. Requirement 4) can be expressed mathematically by

$$T_{23} > T_{12}\left(\frac{N_{2eq} - N_{3eq}}{N_{2eq} + N_{3eq}}\right)\left(\frac{N_{1eq} + N_{2eq}}{N_{1eq} - N_{2eq}}\right)$$

where $N_{ieq}$ is the population of the $i$th level under conditions of thermal equilibrium (with no microwave fields present).

Fulfilling all five above requirements can only be done using very special systems. Such systems can be found in the energy level spectra of certain paramagnetic transition metals, using suitable compounds to achieve desired crystalline field splittings of otherwise degenerate energy levels. Work on an amplifier using multilevel excitation is being undertaken at Harvard University and elsewhere.[27]

### *F. Optical Pumping*

Still another way to achieve an emissive state utilizing several energy levels is known as "optical pumping."[28] In "optical pumping," the system is illuminated with suitably polarized light of a wavelength corresponding to an optical transition of the molecules. The absorption of this *polarized* light, and subsequent spontaneous emission of *unpolarized* light can result in a rearrangement of molecules among a group of states so that those in two states connected by a microwave transition are in an emissive condition. To see this in more detail, consider an atom (molecule) with the (partial) energy level system shown in Fig. 14. There are two sets of energy levels of interest: the upper set consists of five degenerate levels and is separated from the lower set of three degenerate levels by an "optical transition" ($\nu = 10^{14}$ to $10^{15}$ sec$^{-1}$). The levels are designated by the quantum numbers $F$ and $M_F$. $F$ is the total angular momentum of the atom, in units of $h/2\pi$, about the $Z$ axis, and $M_F$ is the component of $F$ along the $Z$ axis.

Consider what happens to an atom that absorbs a photon from a beam of circularly polarized light traveling in the $Z$ direction. Such a photon carries a $Z$ component of angular momentum of one unit and one unit of total angular momentum. Assume the sense of polarization is such that the photon has an effective $M_z = +1$. Then the selection rule for absorption is $\Delta M_F = +1$, and an atom originally in the $F=1$, $M_F = -1$ state goes into the $F=2$, $M_F=0$ state upon photon absorption. Atoms in the upper set of levels are highly excited, however, and re-emit a photon spontaneously in about $10^{-8}$ seconds. This photon is generally unpolarized, as the selection rule for spontaneous emission is $\Delta M_F = \pm 1$ or 0. This is indicated for the $F=2$, $M_F=0$ state on Fig. 14. It is seen that after a complete cycle of absorption and re-emission, an atom originally in the $F=1$, $M_F = -1$ state can be in any of the three $F=1$ states. An atom that was initially in the $F=1$, $M_F = +1$ state, however, must remain in the same state after a cycle, as the only transition possible in reemission is the $\Delta M_F = -1$ one.

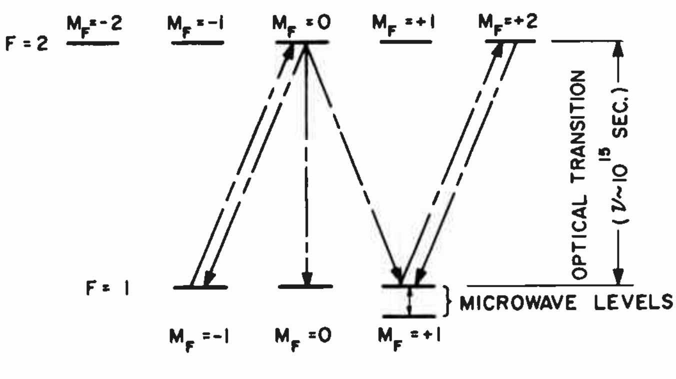


Fig. 14—Schematic energy level diagram showing optical transitions as used in "optical pumping." $M_F$ is the component of total angular momentum $F$ along the axis of the system.

It can therefore be seen that illumination of the system with polarized light can result in an effective depopulation of some levels ($F=1$, $M_F = -1$ in the case above) and an increase in population of others ($F=1$, $M_F = +1$ above). If a state that is increased in population is also coupled by a microwave transition to a lower state (as shown in Fig. 14), an emissive condition may be maintained with respect to these two levels. Such a gas of atoms could be used in a suitable microwave structure as an amplifying or oscillating medium.

There are several practical difficulties besetting the utilization of this method for obtaining a molecular system in an emissive condition. First, a molecule or atom having a suitable energy level scheme must be found. The alkalai metal vapors and atomic hydrogen fulfill this requirement; the optical transition is between the $^2S_{1/2}$ atomic ground state of the atoms and the first excited $P$ state, while the hyperfine structure of the ground state provides the microwave transition. Vapor pressures, which determine available gas densities, and, in the case of atomic hydrogen, recombination processes, must also be considered. If the optical transition is in either the near infrared or the ultraviolet, polarization difficulties may arise.

[27] N. Bloembergen, "Proposal for a new type solid state Maser," *Phys. Rev.*, vol. 104, pp. 324–327; October 15, 1956.
H. E. D. Scovil, C. Feher, and H. Seidel, "The operation of a solid state maser," (to be published).

[28] A Kastler, "Quelques suggestions concernent la production optique et la détection optique d'une inégalité de population des niveaux de quantification spatiale des atoms," *J. Phys. Rad.*, vol. 11, pp. 255–265; June, 1950.
J. Brossel, A. Kastler, and J. Winter, "Création optique d'une inégalité de population entre les sous-niveaux zeeman de l'état fondamental des atoms," *J. Phys. Rad.* vol. 13, p. 668; December, 1952.
W. B. Hawkins and R. H. Dicke, "The polarization of sodium atoms," *Phys. Rev.*, vol. 91, pp. 1008–1009; August 15, 1953.

In the above description of the "optical pumping" process, several simplifying assumptions were made. One was that the absorption process could be considered to go between an initial state and *one* final state. The upper optical levels are very broad, however, and the possibility exists of two such states, to both of which the selection rules permit transitions, "overlapping." In this case, complex "interference" effects occur; their result is that the excited atom "remembers" the angular momentum absorbed and tends to reemit it again, thus leaving no net "pumping" effect. It was further implicitly assumed that only one of the two microwave-coupled levels was optically coupled to the upper optical levels. Because optical lines are so broad, this cannot be achieved, and the populations of both of the microwave levels may be altered by "optical pumping." It can be shown, however, that in favorable cases both these possible sources of difficulty do not prevent an emissive state from being attained.

A sketch of one practical way in which an "optically pumped" amplifier might be achieved is shown in Fig. 15. The molecular system consists of rubidium vapor at

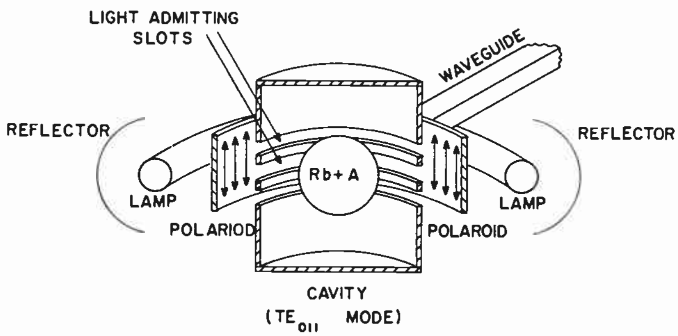


Fig. 15—"Optically pumped" amplifier.

a pressure of roughly $10^{-6}$ mm Hg. The microwave transition is between the hyperfine levels of the ground state, and occurs at 6834.68 mc. The optical transition is in the near infrared at about 7900 *A*. This can easily be linearly polarized using Polaroid sheets. The rubidium vapor is contained in a small glass bulb at the center of the microwave cavity. Argon at a pressure of about one mm Hg is also in the bulb. This increases the diffusion time of the rubidium vapor to the walls of the bulb, and effectively eliminates Doppler effects.[29] The cavity is cylindrical and resonant in the $TE_{011}$ mode. The light, polarized along the cavity axis, is incident on the vapor through circular slots cut in the walls of the cavity. This can be done without greatly affecting the $Q$ of the desired resonant mode. The effect of "optical pumping" on the population of the various hyperfine energy levels is shown in Fig. 16. At thermal equilibrium, the states are populated as shown on the upper series of energy levels. After each atom has absorbed and re-emitted an optical photon, the distribution is as shown on the lower series of levels. It is clear that for the three $\Delta M_F=0$ transitions, the system after "pumping" is highly emissive, and in fact corresponds to the very low negative internal temperature $T_{molec}=0.12°K$.

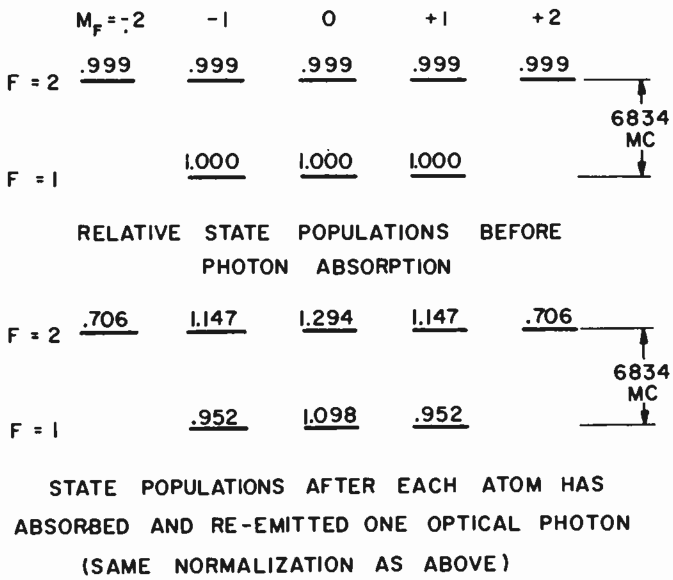


Fig. 16—Effect of "optical pumping" on the state populations of the magnetic substates of the hyperfine levels of the ground $S$ state of rubidium vapor. $F$ is the total angular momentum of an atom and $M_F$ its component along the axis of the system.

The major weakness of amplifiers based on this excitation method is a quantitative one. It seems necessary to go to a gaseous system to find a molecular energy level scheme that is suitable. Furthermore, the density of the active gas must be relatively low, as indicated above, or "light trapping" occurs; that is, the optical photon re-emitted by one excited atom is reabsorbed by another before it can "escape" from the gas. Since the re-emitted photon is unpolarized, its reabsorption tends to "unpump" the atom absorbing it. Thus the total "excess population" $N$ is very low. Also, at least in the molecular systems listed above, the microwave transition is a *magnetic* dipole one, with the accompanying low dipole moment $p$.

Because of these factors, to get amplification it is necessary to make the bandwidth extremely small. For this reason, optically pumped amplifiers may be limited to applications where this is an advantage: high spectral purity oscillators and frequency standards. A frequency standard, or "atomic clock" using optically pumped rubidium vapor is under study at Princeton University.[30] The expected molecular bandwidth of this rubidium amplifier, with its utilization of a "buffer gas" to eliminate Doppler effects, is of the order of 100 cps. Since the

[29] R. H. Dicke, "The effect of collisions upon the Doppler width of spectral lines," *Phys. Rev.*, vol. 89, pp. 472–473; January 15, 1953.

[30] I am indebted to Dr. T. R. Carver of Princeton University for permission to quote the results of calculations on this device that he has made. The numerical example given in the text is based on his work.

amplification bandwidth, or noise-induced frequency modulation as an oscillator should be much less than this, a very high frequency stability should be achieved.

The calculation of an "oscillator parameter $L$" (see Section III, C) for this device is difficult, as several unknown factors enter. However, an estimate indicates that a loaded cavity $Q$ of about 60,000 is required to sustain oscillations (*i.e.*, for $L=1$). With the glass bulb in the cavity and the slots for the admission of the polarized light, this may be an unattainable $Q$. Even if it ultimately so proves, however, such a device will provide an extremely sharp resonance that can be used for frequency control purposes.

The efficiency of an optically-pumped amplifier is very low. Nearly all of the (relatively) huge amount of energy supplied in the optical frequency range is lost, either in optical reradiation or in resistive losses to heat in the cavity.

## V. Discussion

The above perhaps gives some idea of the wide variety of devices and systems covered by the common title of "molecular amplifier." While it is believed that the most promising amplifier schemes now proposed or being worked on have been included in this discussion, it is also quite certain that this listing is not a final, definitive one. The wide range of approaches to emissive states indicated have all been invented or developed within the past few years, and it seems highly unlikely that the field is already exhausted of novel solutions to its problems.

This diversity has in some respects prelcuded a general discussion of molecular amplifiers and oscillators, and necessitated separate descriptions for each of the excitation methods and molecular systems. The basic properties of all molecular amplifiers are quite similar however. The weakness of the individual molecular interactions in the microwave region, of the order of $10^{-5}$ to $10^{-4}$ electron volts per molecule, argues strongly for the use of a resonant microwave structure for the region of interaction between molecules and radiation. With such a structure, which includes built-in positive feedback, gain is of itself no problem: by increasing feedback, *i.e.*, cavity $Q$, any desired gain can be obtained. The problem of power losses in the microwave structure, on the other hand, is a real one. Extremely loss-free structures are needed so that the weak molecular power exceeds losses by a sufficient margin to permit low-noise operation.

There are two main sources of noise in a molecular amplifier: noise due to spontaneous incoherent emission from the excited molecules, and noise from emission induced by thermal radiation from the microwave structure.[81] The latter can often be reduced by cooling the microwave structure. Despite these noise sources, noise figures of under 1 db can be achieved, and in many communications and radar systems where the signal source is such that the background noise entering the amplifier is well below thermal noise at room temperature, this indicates a tremendous increase in sensitivity over conventional microwave amplifiers with noise figures in the range 5–15 db.

Low-noise figure is directly related to the spectral purity when the amplifier is operated as an oscillator. It was seen that the heretofore unattainable short-term frequency stability of a few parts in $10^{12}$ can be achieved with the molecular beam Maser, thus opening up new possibilities for the design of super-stable frequency standards based on atomic processes. These provide a standard of frequency at any place on the earth independent of astronomical observations, radio-transmitted standard frequency signals, etc.

The power handling capabilities of molecular amplifiers have been seen to be low. In some applications, this is a definite disadvantage; for example, in military radar such an amplifier could easily be saturated by a jamming signal. In other applications, however, the low power handling ability of molecular amplifiers provides no serious limitation to their use.

A major problem of molecular amplification, in addition to that of reduction of losses in the microwave structure, is that of bandwidth, or, more exactly gain-bandwidth. As seen above, gain can always be traded for bandwidth, by using an inhomogeneous Stark or Zeeman field to broaden the resonance, by increasing molecular density and hence shortening relaxation times, or by increasing the microwave coupling to the interaction cavity. However, when the molecular emission power is spread over a broad bandwidth, the gain and noise figure worsen. Achievement of a high total available molecular power therefore is of great importance. For gaseous molecular systems, molecular densities are low and bandwidths of a few kilocycles to a few tens of kilocycles are available. With solid-state devices, bandwidths of megacycles should be obtainable.

In addition to gain, noise figure, and bandwidth, frequency coverage is of importance. If the amplification frequency is determined by internal molecular properties, such as in ammonia and possibly in some solid state systems, the frequency is essentially fixed. In some applications, such as a frequency standard, this is desirable, or at least not objectionable. In case a tunable amplification frequency is required, a system such as a paramagnetic spin system where the frequency is determined by the strength of an applied static field can be used. An important possibility arises in this case. Since the excitation frequency is not necessarily the amplification frequency, it is possible to generate microwaves at frequencies higher than those of excitation. This is a possible approach to the problem of the generation (or amplification) of millimeter waves. Methods such as

[81] Additional noise is introduced when the number of interacting molecules fluctuates in time. This is generally relatively unimportant.

the molecular beam Maser that do not utilize rf electromagnetic excitation also offer a possibility for the generation of discrete frequencies in the millimeter and submillimeter region, by using transitions between relatively widely separated energy states. Vibrational transitions and certain rotational transitions, as, *e.g.*, in ammonia, lie in the far infrared and submillimeter part of the frequency spectrum and might prove suitable.

Thus molecular amplifiers provide a means of achieving low noise amplification of microwaves. At present, bandwidths are quite low, but developments in the near future should greatly improve this situation. Below microwave frequencies, the gain fall-off with frequency will probably preclude their use; in the millimeter and submillimeter range, they offer new possibilities for amplification and generation. Many developments both in excitation methods and in molecular systems suitable for use in these devices can be expected in the near future.

## Appendix

Consider a molecule with two energy levels, with an oscillating microwave field $E \cos(2\pi\nu t)$ inducing transitions between them. If the state function of the molecule is written as a linear combination of pure energy state functions, as in (1),

$$\psi = \sum_j a_j(t)\psi_j, \tag{1}$$

then the quantum mechanical equations for the time dependence of the $a_j$'s are given by:[32]

$$\frac{da_j}{dt} = \frac{2\pi}{ih}\sum_n W_{jn}a_n \exp(2\pi i\nu_{jn}t). \tag{26}$$

Here $i=\sqrt{-1}$, $\nu_{jn}=(W_j-W_n)/h$, $W_j$ is the energy of the $j$th state, etc., and $W_{jn}$, the $jn$th element of the perturbation matrix, is given by

$$\begin{aligned} W_{jn} &\equiv \int \psi_j^* W\psi_n dv \\ &= \int \psi_j^*(pE\cos(2\pi\nu t))\psi_n dv \\ &= E\cos(2\pi\nu t)\int \psi_j^* p\psi_n dv. \end{aligned} \tag{27}$$

The integral represents $p_{jn}$, the strength of the dipole moment coupling the states $j$ and $n$.

In the two-state problem at hand, (26) and (27) become

$$\frac{da_1}{dt} = \frac{2\pi E\cos(2\pi\nu t)}{ih}\left[p_{11}a_1 + p_{12}a_2\exp(2\pi i\nu_{12}t)\right]$$

$$\frac{da_2}{dt} = \frac{2\pi E\cos(2\pi\nu t)}{ih}\left[p_{21}a_1\exp(2\pi i\nu_{21}t) + p_{22}a_2\right].$$

[32] L. I. Schiff, "Quantum Mechanics," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 189–190; 1949.

Here $p_{ii}$ indicates the dipole moment possessed by a molecule when it is in the state of energy $W_i$. Such static dipoles do not lead to resonant transition effects and will be neglected; also $p_{12}=p_{21}=p$, in the cases of interest at present, so

$$\frac{da_1}{dt} \approx \frac{2\pi pE}{ih}\cos(2\pi\nu t)\exp(2\pi i\nu_{12}t)a_2$$

$$\frac{da_2}{dt} \approx \frac{2\pi pE}{ih}\cos(2\pi\nu t)\exp(2\pi i\nu_{21}t)a_1.$$

Also, $\nu_{21}=-\nu_{12}=\nu_0$, and defining $\nu\equiv\nu-\nu_0$,

$$\frac{da_1}{dt} = \frac{\pi pE}{ih}\left\{\exp[2\pi i\Delta\nu t] + \exp[-2\pi i(\nu+\nu_0)t]\right\}a_2$$

$$\frac{da_2}{dt} = \frac{\pi pE}{ih}\left\{\exp[2\pi i(\nu+\nu_0)t] + \exp[-2\pi i\Delta\nu t]\right\}a_1.$$

The terms with $(\nu+\nu_0)$ represent very high frequency perturbations on the basic behavior of the system. As they do not lead to any net transition effects over a reasonable averaging time, they can also be neglected, leading to

$$\begin{aligned} \frac{da_1}{dt} &= \frac{\pi pE}{ih}\exp(2\pi i\Delta\nu t)a_2 \\ \frac{da_2}{dt} &= \frac{\pi pE}{ih}\exp(-2\pi i\Delta\nu t)a_1. \end{aligned} \tag{28}$$

It can be readily shown that the expressions

$$a_1 = \exp(\pi i\Delta\nu t)\left[\cos\lambda t - \frac{\pi i\Delta\nu}{\lambda}\sin\lambda t\right]$$

$$a_2 = \frac{\pi ipE}{h\lambda}\exp(-\pi i\Delta\nu t)\sin\lambda t$$

satisfy (28) if

$$\lambda = \pi\left[(\Delta\nu)^2 + \left(\frac{pE}{h}\right)^2\right]^{1/2}.$$

Furthermore, at $t=0$ they reduce to

$$a_1(0) = 1$$

$$a_2(0) = 0.$$

For such a molecule in energy state 1 at time $t=0$, the probability that it has undergone a transition and is in state 2 at time $t$ is:

$$\text{Transition probability} = |a_2(t)|^2$$

$$= \frac{\left(\frac{pE}{h}\right)^2\sin^2\left\{\pi\left[(\Delta\nu)^2+\left(\frac{pE}{h}\right)^2\right]^{1/2}t\right\}}{(\Delta\nu)^2+\left(\frac{pE}{h}\right)^2},$$

which is (6) in the text.

# The Spacistor, A New Class of High-Frequency Semiconductor Devices*

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*Summary*—**New devices are considered in which electrons or holes are injected directly into space-charge regions of reverse-biased junctions avoiding the diffusion of carriers through field-free regions. The case considered is one in which the junction is biased at a voltage such that the injected carriers are multiplied by the avalanche process. A device of this type shall be called a spacistor. It is shown that negative resistance devices and amplifying devices may be constructed. The difficulty is the accumulation of the generated carriers in front of the emitting contact. Experimentally, it has been found that this accumulation of carriers results in the spontaneous relaxation-type oscillations of a transistor-like structure operated at collector voltages greater than the punch-through voltage. It is suggested that the accumulation effect may be diminished by the use of small emitters and magnetic fields.**

## Introduction

THE FREQUENCY range of transistors has been extended considerably in the past few years.[1–6] An investigation was made to determine types of oscillating or amplifying semiconductor devices which are, in principle, capable of extending the useful frequency range of transistors to even higher frequencies. In addition to extrinsic base resistance and the collector capacity, the $\alpha$-cutoff frequency essentially limits the frequency response of a transistor. The $\alpha$-cutoff frequency can be considered to arise from the finite "transit time" of a minority carrier through the base region and the collector space-charge region. The base region of the transistor is either field free or has a relatively weak built-in field with a potential drop of 0.1 to 0.2 volts. Thus the "transit time" in present transistors has been determined almost completely by the characteristics of the base region. Therefore new devices were constructed which no longer use the diffusion of minority carriers through virtually field-free regions.

The first attempt in this direction has been made by Shockley[7] with the so-called unipolar field-effect transistor. It appears, however, that this device will not exceed the frequency response of the transistor.[8] Another approach to the same problem made by Shockley[9] used the transit time through the space-charge regions of reverse-biased junctions to obtain negative resistance in a small frequency interval. This problem has recently been reconsidered by Read[10] who proposes to obtain negative resistance by the use of the transit time through space-charge regions in conjunction with the build-up time of avalanche breakdown. The latter approaches[9,10] may result in useful oscillators at a frequency which is of the order of the inverse transit time of a carrier through the space-charge region of a junction and may, thus, be of the order of $10^4$ mc.

In the present paper, new devices are suggested which also use processes in space-charge regions of reverse-biased junctions. However, in all the devices proposed here, a third injecting contact is made to the space-charge region. Only devices will be considered for which the junction is biased with a voltage high enough so that avalanche multiplication takes place to an appreciable extent. A device using these principles shall be called a spacistor. Two possible geometrical arrangements will be discussed in detail. In the first, an injecting "point" contact is made into the space-charge region of a reverse biased $p$-$n$ junction. In the second, a transistor-like structure is discussed. The base region has a width such that the collector space charge extends to the emitter junction or "punches-through" at a voltage which is near the avalanche breakdown voltage.

It has not yet been possible to construct the former device successfully; however, extensive experimental studies have been made of the latter structure. From these studies it became apparent that there is a tendency for carriers generated by the avalanche multiplication mechanism to accumulate in front of the injecting emitter and, thus, greatly alter the space charge there. This accumulation results in relaxation-type oscillations when the injecting contact is biased with a high resistance source. It may be possible that use of a small emitter in conjunction with a magnetic field in the base will eliminate the carrier accumulation and thus prevent the undesired oscillations.

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[2] 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.

[3] H. Kromer, "Zur theorie des diffusions- und des drifttransistors," *Archiv der Elek. Übertragung*, vol. 8, pp. 223–228; May, 1954.

[4] H. Statz, W. F. Leverton, and J. Spanos, "Transistors with an Exponential Impurity Distribution in the Base Region," Paper presented at the AIEE-IRE Semiconductor Devices Res. Conf., Philadelphia, Pa.; June, 1955.

[5] M. Tanenbaum and D. E. Thomas, "Diffused emitter and base silicon transistors," *Bell Sys. Tech. J.*, vol. 35, pp. 1–22; January, 1956.

[6] C. A. Lee, "A high-frequency diffused base germanium transistor," *Bell Sys. Tech. J.*, vol. 35, pp. 23–34; January, 1956.

[7] W. Shockley, "A unipolar 'field effect' transistor," Proc. IRE, vol. 40, pp. 1365–1376; November, 1952.

[8] G. C. Dacey and I. M. Ross, "The field-effect transistor," *Bell Sys. Tech. J.*, vol. 34, pp. 1149–1189; November, 1955.

[9] W. Shockley, "Negative resistance arising from transit time in semiconductor diodes," *Bell Syst. Tech. J.*, vol. 33, pp. 799–828; July, 1954.

[10] W. T. Read, Jr., "A Proposed High-Frequency, Negative Resistance Diode Using Breakdown and Transit Time," Paper given at AIEE-IRE Semiconductor Devices Res. Conf., Purdue Univ., Lafayette, Ind.; June, 1956.

## A Negative Resistance Device

A $p$-$n$ junction with an applied reverse bias is shown in Fig. 1(a). The $p$-type region is considered to be doped the more heavily of the two so that the space charge extends mainly into the $n$-type region. If the doping in the $n$- and $p$-type region is uniform, then a plot of the potential is distributed parabolically as a function of distance on either side of the junction. In Fig. 1(b), the potential distribution in the space-charge region is shown for two applied voltages $V_1$ and $V_2$. Next, consider a small $p$-type contact situated in the space-charge region as shown in Fig. 1(a). This contact will be referred to as a "bond."

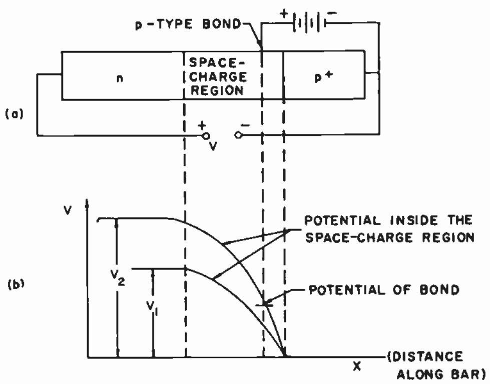


Fig. 1—(a) Negative resistance device, and (b) potential distribution in space charge of junction.

If this bond is electrically open-circuited, then it will essentially assume the potential that exists in the space-charge region immediately under it. However, if the small bond region is biased positively with respect to the potential in the underlying space-charge region, holes will be emitted into this region. Since the geometry of the device is not one-dimensional, a calculation cannot be carried out readily; however, it is expected that the emission of holes is "space-charge limited" and qualitatively similar to the one-dimensional case.[11] If the small $p$-type bond is biased negatively with respect to the potential of the underlying space charge, no emission of holes will take place. Some holes which are generated thermally in the $n$-type region within a diffusion length of the junction and which flow through the space-charge region will be collected by the small bond.

Suppose that the $p$-type bond is connected in series with a battery to the $p$-region as indicated in Fig. 1(a). The voltage of the battery is chosen such that the bond has a potential as indicated in Fig. 1(b). For the lower applied voltage $V_1$, the bond emits holes into the space-charge region; for the higher applied voltage $V_2$, no holes are emitted. In the following discussion only the emitted current will be considered. A sketch of the dependence of emitted bond current on applied voltage $V$ is shown in Fig. 2(a). Applied voltages which are less than the bond bias voltage are not of interest here. This voltage region is marked by a shaded area in Fig. 2. In this case, holes will not only flow to the $p$-side but will also be emitted into the $n$-region. In the $n$-region there is virtually no field and the holes will diffuse and recombine. This diffusion process is a relatively slow one. However, consider the situation when the applied voltage exceeds the bond bias. Obviously, with no avalanche multiplication, the emitted hole current will not cause any additional current to flow through the terminals across which $V$ is applied. A current can be induced in this circuit only if the emitted holes create electron-hole pairs at the junction. The electron-hole pairs are separated by the field of the junction, and a current flows. The magnitude of this current is

$$I = (m - 1)I_{\text{Bond}}, \tag{1}$$

where $m$ is the avalanche multiplication factor and $I_{\text{Bond}}$ is the current emitted by the bond. The multiplication factor is not necessarily the same as in the isolated junction since the bond distorts the field. The electric field in the vicinity of the junction will depend, to an appreciable degree, on the voltage applied to the bond which, of course, is independent of the applied voltage $V$. Therefore, the variations of $m$ with applied voltage are expected to be smaller when the bond is biased than when it is not. Thus, the diode current characteristic will be as indicated in Fig. 2(b); there may be a maximum in the $I$, $V$ curve as indicated by the dashed line. It is, however, difficult to predict the exact shape of the curve without a detailed calculation.

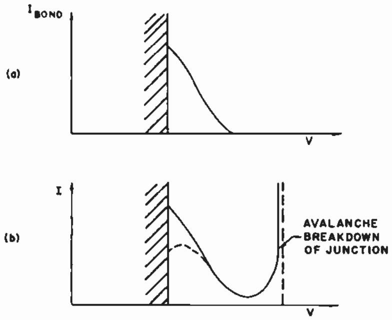


Fig. 2—(a) Bond current as a function of applied voltage $V$. (b) Diode current as a function of applied voltage $V$.

It can be stated, quite generally, that there must be a negative resistance region because the current emitted by the bond becomes zero at a certain voltage; and thus, the induced current also becomes zero. The small current still flowing at the minimum of the $I$, $V$ curve is the multiplied saturation current of the junction. If the applied voltage is increased to even higher values, the junction eventually breaks down. Devices with negative resistance regions may be used as high-frequency oscillators. Also, bistable devices for high speed switching can be built.

[11] W. Shockley and R. C. Prim, "Space-charge limited emission in semiconductors," *Phys. Rev.*, vol. 90, pp. 753–758; June, 1953.
G. C. Dacey, "Space-charge limited hole current in germanium," *Phys. Rev.*, vol. 90, pp. 759–763; June, 1953.

Consider the size of a typical space-charge region. Suppose the avalanche-breakdown voltage occurs at 200 volts. This implies, for example, in germanium[12] that the $n$-type side contains about $6\times10^{14}$ donor atoms per cm$^3$. The width of the space-charge region is

$$W = \sqrt{\frac{2\epsilon}{\rho} V}, \tag{2}$$

where $V$ is the applied voltage, $\epsilon$ is the dielectric constant of the semiconductor, and $\rho$ is the charge density due to the ionized donors. For $V$ equal to 200 volts, this width becomes, for the above mentioned concentration of donors, $W=2.4\times10^{-3}$ cm. For $V$ equal to 100 volts, the width will be about $1.7\times10^{-3}$ cm. The problem of making bond contacts much smaller than the above widths has not been solved. In experiments performed at this laboratory with unformed tungsten points placed into the space-charge region no emission of holes was observed. The emission of holes from a point contact depends upon the formation of an inversion layer under the point contact. To ensure hole emission it therefore seems desirable to create a small $p$-type region by doping.

There are, however, many ways by which the above described task can be simplified. For example, nonconstant doping may be used (see Fig. 3). On the left of the

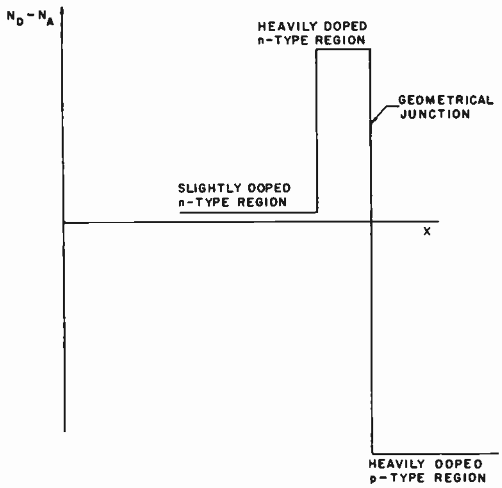


Fig. 3—Scheme of donor minus acceptor concentration for a proposed negative resistance device.

junction there is a heavily doped $n$-type region followed by a slightly doped $n$-type region. One may choose the width and impurity concentration of the heavily doped region such that the space charge includes this region for an applied voltage at which appreciable multiplication takes place. A relatively small increase in applied voltage will then cause a rapid increase in the space-charge width because of the adjacent nearly intrinsic region, without affecting the multiplication factor appreciably. By a scheme like this, the space-charge region can be made relatively wide, and at the same time, multiplication will occur at relatively low applied voltages.

## Devices With Transistor-Like Structures

As an example of another structure, consider a regular alloyed-junction transistor with a relatively small base width. The base width is assumed to be of such a magnitude that punch-through takes place at a collector voltage which is an appreciable fraction of the avalanche-breakdown voltage. In this transistor, the collector corresponds to the $p^+$-side of the diode of Fig. 1(a), the emitter corresponds to the $p$-type bond, and the base region corresponds to the $n$-side of the diode. Because of the plane parallel geometry of emitter and collector junction, significant differences exist between the two devices.

First, consider the floating potential of the emitter junction as a function of the voltage applied between the base region and the collector.[13] Before punch-through, the emitter is slightly reverse biased, and the voltage difference is given by[14,15]

$$V_f = -\frac{kT}{q}\ln(1-\alpha),$$

where $k$ is Boltzmann's constant; $T$, the absolute temperature; $q$, the electronic charge; and $\alpha$, the emitter to collector current amplification factor. This floating potential rarely exceeds 0.1 to 0.2 volt. However, after punch-through has taken place, the voltage difference between emitter and collector remains constant as the collector junction bias is varied. This can be seen schematically in Fig. 4. In Fig. 4(b) three cases are shown. In the first case, the applied voltage $V_{BC}$ is less than the punch-through voltage $V_P$; the space charge does not extend to the emitter. In the second case, $V_{BC}=V_P$ and the space charge extends up to the emitter junction. In the third case, $V_{BC}>V_P$; the solution of Poisson's equation is the same for this case as for $V_{BC}=V_P$ except for an additive constant in the potential. The reason is that the electric field must vanish in front of the emitter junction just as it does for $V_{BC}=V_P$. Thus, $V_{BC}-V_{BE}=V_P$, where $V_{BE}$ is the potential difference between base and emitter [see Fig. 4(b)].

If a voltage is applied between emitter and collector which is somewhat larger than the punch-through voltage, a current will flow. This current is independent of the voltage $V_{BC}$ as long as $V_{BC}>V_P$. Results on a typical experimental unit when $V_P$ is equal to 12 volts are

[12] S. L. Miller, "Avalanche breakdown in germanium," *Phys. Rev.*, vol. 99, pp. 1234–1241; August, 1955.

[13] H. Schenkel and H. Statz, "Voltage punch-through and avalanche breakdown and their effect on the maximum operating voltages for junction transistors," *Proc. Nat. Elect. Conf.*, vol. 10, pp. 614–625; 1954.

[14] W. Shockley, "The theory of P-N junctions in semiconductors and P-N junction transistors," *Bell Syst. Tech. J.*, vol. 28, pp. 435–489; July, 1949.

[15] W. Shockley, M. Sparks, and G. K. Teal, "P-N junction transistors," *Phys. Rev.*, vol. 83, pp. 151–162; July, 1951.

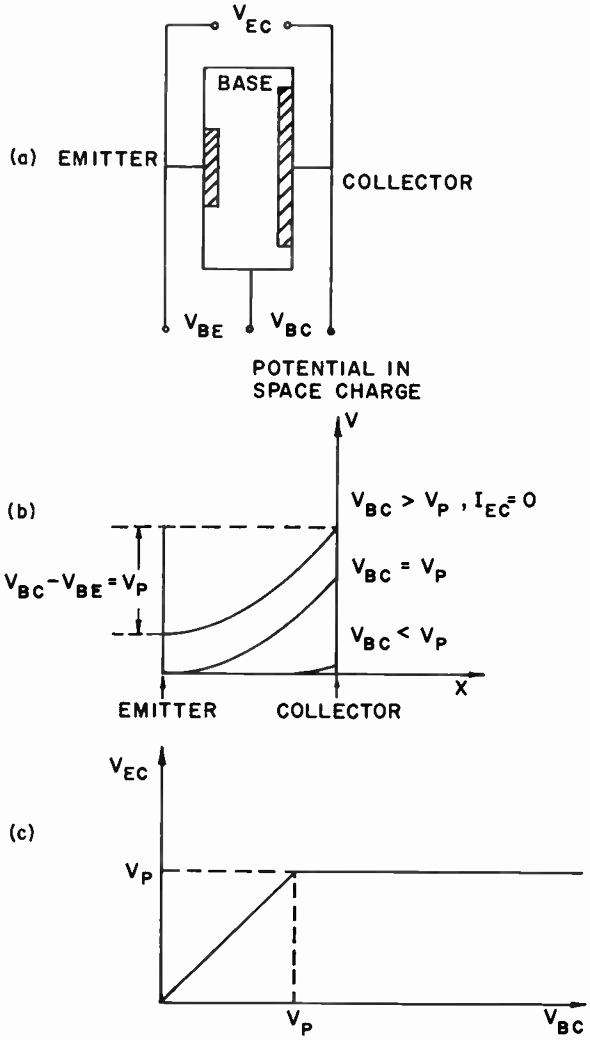


Fig. 4—(a) Cross section of transistor-like structure. (b) Potential distribution in base for three voltages $V_{BC}$. (c) Open circuit voltage between emitter and collector as a function of voltage between base and collector.

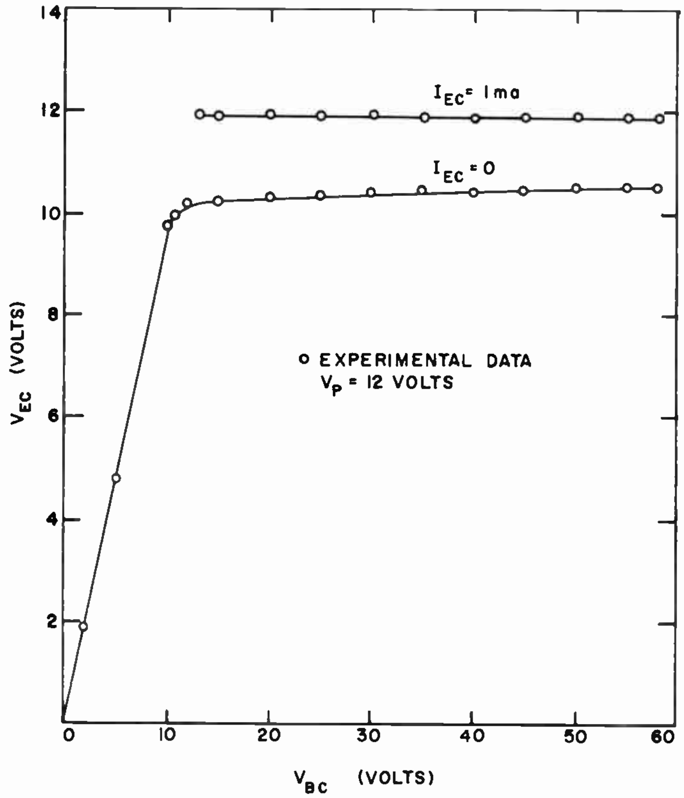


Fig. 5—Emitter to collector voltage as a function of base to collector voltage for $I_{EC}=0$ and $I_{EC}=1$ ma.

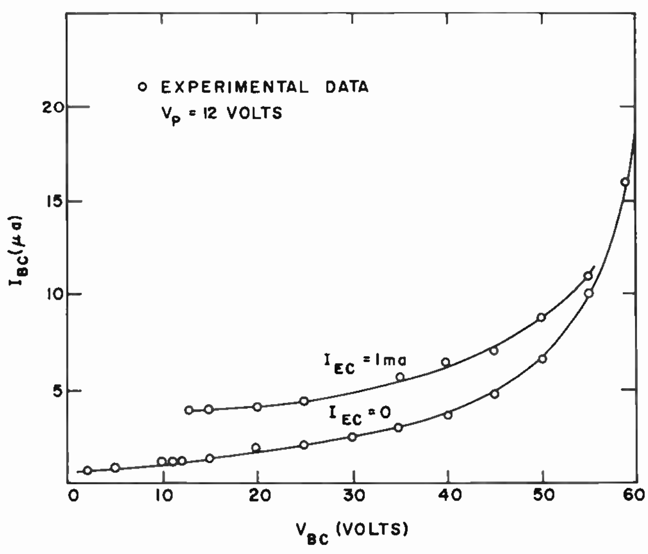


Fig. 6—Base current as a function of base to collector voltage for $I_{EC}=0$ and $I_{EC}=1$ ma (punch-through voltage approximately 12 volts).

shown in Fig. 5. When $I_{EC}$, the current flowing from emitter to collector is zero, the voltage $V_{EC}$ varies slightly with $V_{BC}$; however, when $I_{EC}$ equals 1 ma, $V_{EC}$ is practically independent of $V_{BC}$ provided the latter exceeds the punch-through voltage. The input resistance $dV_{EC}/dI_{EC}$ is a function of the current $I_{EC}$. This resistance decreases with increasing current. If multiplication takes place near the collector junction, electron-hole pairs will be generated. The electrons should not flow into the emitter but should rather flow out into the base; and thus, a current $I_{BC}$ may be induced in the high resistance collector circuit. If the ratio of the output resistance to the input resistance is high enough, and if the avalanche multiplication factor is large enough, then it should be possible to obtain power amplification. In Fig. 6, the base current $I_{BC}$ is shown as a function of collector voltage for $I_{EC}$ equal to zero and 1 ma (for the transistor as of Fig. 5). It may be seen that some leakage current flows across the collector junction. As expected, the current in the base-collector circuit is somewhat higher for $I_{EC}=1$ ma. However, in this case, the multiplication is very small. From these curves, a value of approximately 0.0026 is obtained for $(m-1)$. From the constant difference between the collector currents it is seen that $(m-1)$ is essentially independent of collector voltage for $V_{BC} > V_P$. (At high values of $V_{BC}$, heat generated in the collector causes deviations.) The factor $m$ is nearly constant because the field distribution between emitter and collector is essentially independent of the voltage applied between base and collector when $V_{BC} > V_P$.

The disadvantages of the transistor-like geometry become quite apparent when units having considerably higher multiplication factors than the above are investigated, *i.e.*, transistors in which the punch-through voltage $V_P$ is a larger fraction of the avalanche-breakdown voltage $V_A$. In all the transistors investigated, for which the ratio $V_P/V_A$ was approximately 0.5 or greater, spontaneous oscillations have been observed. A typical circuit in which these oscillations occur is shown in Fig. 7. The spontaneous oscillations are due to the fact that there is essentially no transverse field along the surface of the emitter so that in a *p-n-p* type structure the generated electrons tend to accumulate in front of the emitter.

In Fig. 8, the emitter-collector voltage $V_{EC}$ and the base current $I_{BC}$ are shown as a function of time; departures from these sketches attributable, probably, to ringing in the circuit, have been omitted. The curves suggest a relaxation type of oscillation which can be explained as follows: There is always some stray capacity in the external circuit between emitter and collector, and the device itself has some internal capacity. If initially $V_{EC}=0$, *i.e.*, there is no voltage difference between emitter and collector, the battery in Fig. 7 proceeds to charge this capacity. Experiments show that the charge time can be explained if the device is assumed to be nonconducting during this time interval. Letting $C$ represent the total capacity, the charge time $T$ is equal to $CV_P/I$, if a constant current source $I$ is used, as shown in Fig. 8. When the capacity has charged to a value $V_{EC}\approx V_P$, the device becomes conducting, and completely discharges the capacity, that is, $V_{EC}$ decays to essentially a zero value. The interval of discharge $T_d$ was so short that it could not be measured with accuracy with the available equipment, when $C$ was the residual stray capacity ($\approx 20$ $\mu\mu$f). ($T_d$ is the time in which $V_{EC}$ decays from 0.9 to 0.1 of its initial value.) This time is estimated to be of the order of a few millimicroseconds for the transistors investigated. The transistors used were of the CK762-type having cutoff frequencies between 14 and 25 mc at a collector bias of 6 volts.

It is interesting to observe that the base current is delayed relative to the discharge and it flows for a much longer period of time compared to the discharge time $T_d$. It can be shown that the maximum repetition rate is related to the time it takes for the accumulated electrons to leave the base region. Repetition rates of 50 mc have been observed.

The energy bands along a line connecting the emitter and collector appear in Fig. 9 (next page). There is a potential maximum just in front of the emitter which limits the flow of holes from the emitter region.[11] If the emitter supply voltage between emitter and collector is made slightly greater than $V_P$, holes will flow from emitter to collector. At the collector, each hole will create on the average $(m-1)$ new electron-hole pairs. The generated electrons will flow toward the emitter and become trapped in the potential energy minimum. Since the electric field at the emitter junction has no component parallel to this junction, the trapped electrons will move to the sides of the emitter by diffusion. The dimensions of the CK762 emitter are large enough (di-

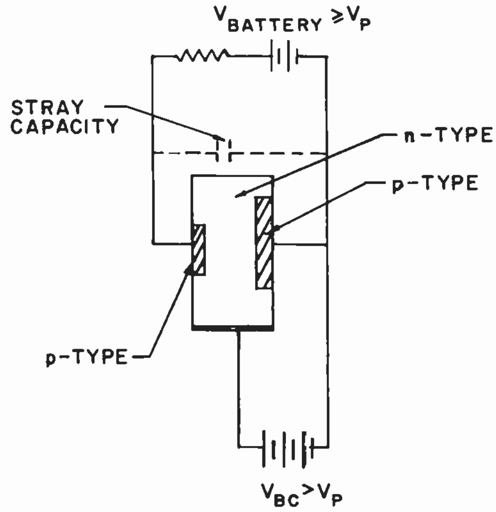


Fig. 7—A circuit for the observation of relaxation type oscillations.

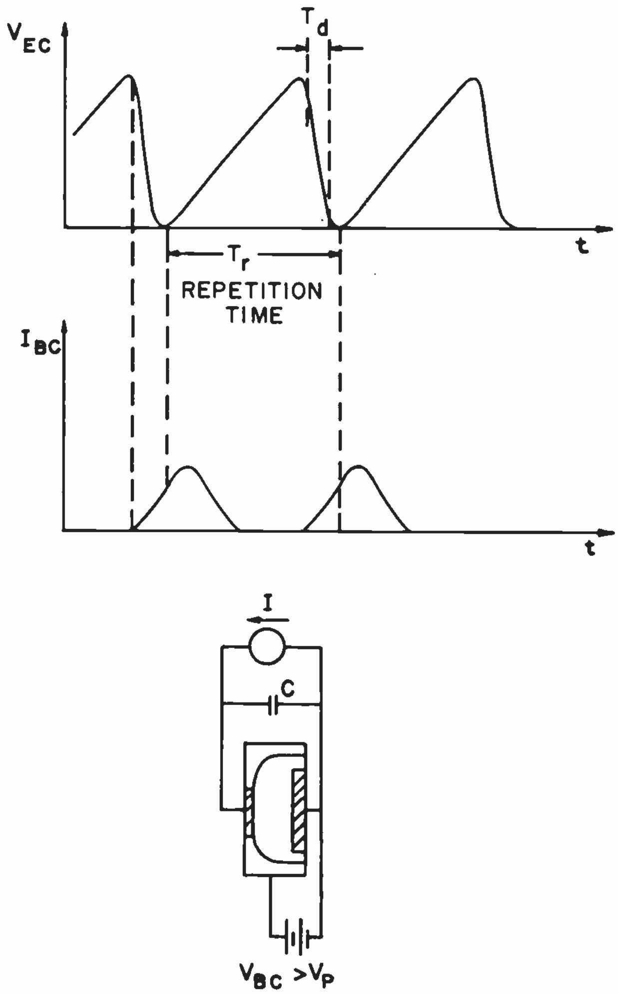


Fig. 8—Emitter to collector voltage and base current as a function of time for the circuit shown.

ameter equal to $2.5\times10^{-2}$ cm), so that the removal of these electrons proceeds rather slowly. The possibility that electrons flow into the emitter can be discarded for the reason that the concentration of $c_1$ of electrons at the potential-energy minimum and the concentration $c_2$ at the base side of the emitter are related as follows:

$$c_1 \approx c_2 e^{\Delta E/kT},$$

where $\Delta E$ is the potential energy difference between the emitter and the potential minimum for an electron in the conduction band; $k$ is Boltzmann's constant; and $T$ is the absolute temperature.

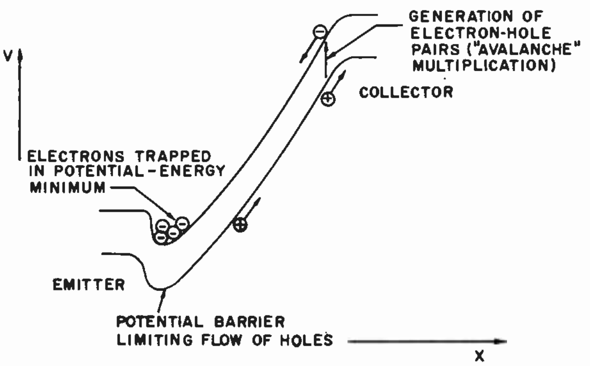


Fig. 9—Plot of energy bands along a line connecting emitter and collector after punch-through.

Since the emitter is field free, any electrons which enter must travel by diffusion. An electronic emitter current as low as 1 microampere requires a concentration of approximately $10^{12}$ electrons per cm$^3$ at the base side of the emitter, assuming a lifetime of $10^{-5}$ second for the electrons. For a potential barrier of 0.5 ev, a concentration of more than $10^{20}$ electrons/cm$^3$ is required at the potential-energy minimum. For higher currents, the concentration requirement becomes correspondingly greater. A very high concentration, of course, will alter the potential distribution and long before such a concentration is reached, the potential minimum model (Fig. 9) will no longer apply.

As electrons in the potential barrier become trapped, the barrier for hole flow will be lowered. This increased flow of holes creates more electrons, and the electrons in turn give rise to a still higher hole current, etc. This process permits the hole current to build up to a very high value. A criterion of such an instability can be postulated as follows: Let one hole create $(m-1)$ electrons. Suppose now that an electron, while it is in the potential minimum, induces the flow of $f$ additional holes (for simplicity $f$ is assumed to be constant). These holes in turn will create new electrons and these electrons, new holes, and so forth. The additional number of holes that are induced to flow because of the first injected hole is given by the infinite series:

$$\begin{aligned} N &= (m-1)f + (m-1)^2f^2 + \cdots \\ &= (m-1)f/[1-(m-1)f]. \end{aligned} \tag{3}$$

This series diverges if $(m-1)f \geq 1$ which means physically that the current builds up rapidly with time. Eq. (3) explains why the device is stable for low values of $m$, since a stable condition corresponds to $(m-1)f<1$.

Once the capacity begins to discharge through the device, Fig. 9 is no longer applicable, since the potential drop between emitter and collector has dropped below $V_P$, which implies that the space charge does not extend throughout the region between emitter and collector. In Fig. 10, the space-charge regions are shown schematically. There is a region in front of the emitter which is virtually field-free. The trapped electrons, which were generated at the collector, not only neutralize the positive donors but also holes which are present in this region. Some electrons probably flow out through the "pinched-off" channel between the space-charge regions adjacent to the emitter and collector.

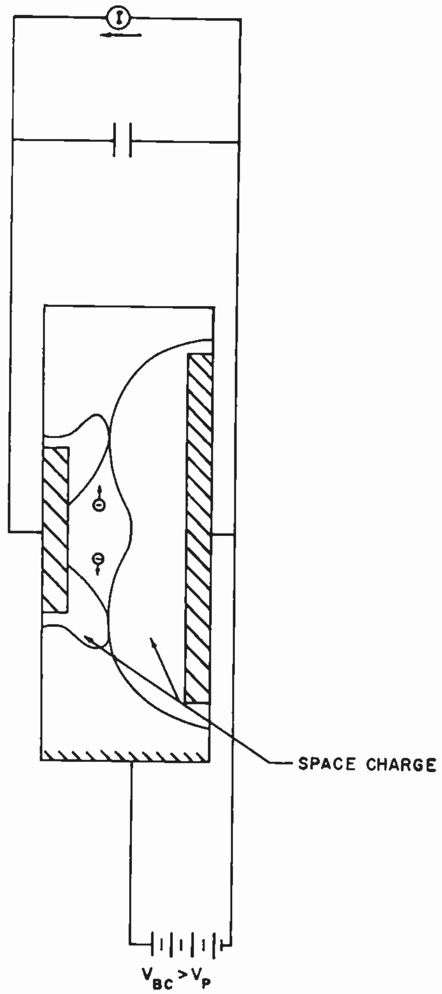


Fig. 10—Scheme of space charge around emitter and collector during discharge of capacity.

From experimental data, rate of flow of electrons through this channel is less than rate at which electrons are created by the avalanche process. Thus, most of the electrons generated near the collector are trapped in front of the emitter junction. To test the above model, a calculation of the discharge time will be made.

### Estimate of the Discharge Time

In a regular $p$-$n$-$p$ transistor the current $I$ flowing from emitter to collector is given approximately by

$$I_{EC} = AD_p q p_0 / W. \quad (4)$$

Here, $A$ is the emitter area; $D_p$, the diffusion constant for holes; $q$, the electronic charge; and $p_0$, the density of holes in the base region at the emitter side; and $W$, the width of the base region or, more exactly, the distance between the base-side boundaries of the emitter and collector space-charge regions. The total charge of the excess holes in the base region is given by

$$Q_h = \frac{Aqp_0W}{2}. \quad (5)$$

From (4) and (5),

$$I_{EC} = \frac{2D_pQ_h}{W^2}. \quad (6)$$

During the discharge interval the hole current is large, therefore, so-called high-injection conditions exist. The total electronic charge $Q_e$ in the base region is approximately equal in magnitude to the hole charge $Q_h$, since the charge arising from the ionized donors can be neglected. Thus

$$Q_e \approx Q_h. \quad (7)$$

From (6) and (7),

$$I_{EC} = 2D_pQ_e/W^2. \quad (8)$$

The width $W$ is related to the voltage difference between emitter and collector. If the distance between the geometrical emitter and collector junctions is $l$ then the punch-through voltage is

$$V_P = \frac{\rho l^2}{2\epsilon} \quad (9)$$

where $\rho$ is the charge density of the ionized donors and $\epsilon$, the dielectric constant. When the voltage between emitter and collector has decreased from $V_P$ to $V_P - V'$, the following relation

$$V_P - V' = \rho\frac{(l - W)^2}{2\epsilon} \quad (10)$$

exists, from which one obtains

$$\frac{W}{l} = 1 - \sqrt{1 - V'/V_P}. \quad (10a)$$

$I_{EC}$ is related to the rate of discharge of the capacity $C$ as follows:

$$I_{EC} = C\frac{dV'}{dt}. \quad (11)$$

From (8), (10a), and (11) it follows that

$$C\frac{dV'}{dt} = 2D_pQ_e/l^2(1 - \sqrt{1 - V'/V_P})^2. \quad (12)$$

If it is assumed that $Q_e$ is the total electronic charge generated near the collector and that none of this charge has passed to the base, then

$$Q_e = \int_0^{V'} [m(V') - 1]dQ_h = C\int_0^{V'} [m(V') - 1]dV'. \quad (13)$$

In the above, the small fraction of injected holes which are needed to maintain the neutral region have been neglected. Using

$$m(V') = \frac{1}{1 - \left(\frac{V_P - V'}{V_A}\right)^n} \quad (14)$$

given by Miller,[12] where for 0.5 to 1 ohm cm $n$-type germanium $n$ is approximately equal to three, the integral becomes

$$\frac{Q_e}{CV_P} = \int_0^{V'} \left(\frac{V_P - V'}{V_A}\right)^3 \left[1 - \left(\frac{V_P - V'}{V_A}\right)^3\right]^{-1} \frac{dV'}{V_P}. \quad (15)$$

This integral is plotted in Fig. 11 as a function of $V'/V_P$, with $V_P/V_A$ as a parameter.

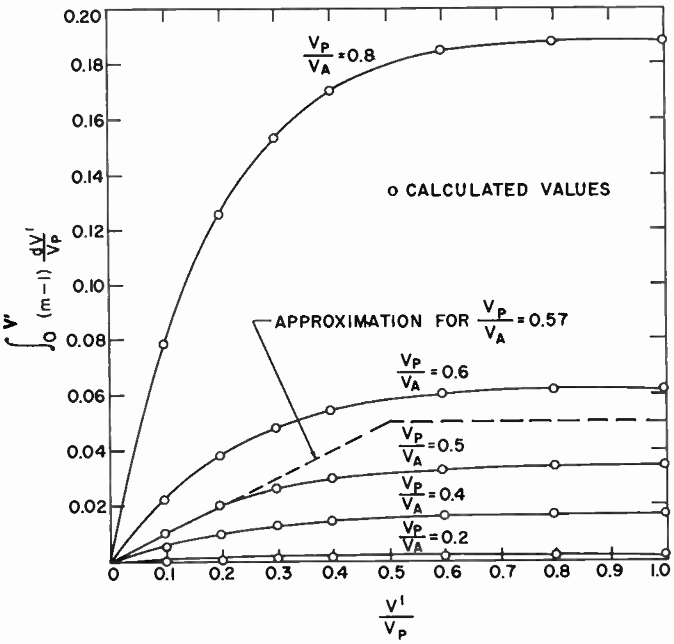


Fig. 11—$\int_0^{V'}(m-1)\, dV'/V_P$ as a function of $V'/V_P$ for various values of $V_P/V_A$.

In the investigations, a transistor with $V_P = 24$ volts and $V_A = 42$ volts was studied extensively. For this case, $V_P/V_A = 0.57$. Eq. (12) has been integrated analytically for the dashed curve in Fig. 11, and $V_{EC}$ was obtained as a function of time. The result is shown in Fig. 12. The discharge time is seen to be about $3 \times 10^{-9}$ second. It is interesting to note that in the calculations, the quantity $C$ drops out [as may be seen by inserting (13) in (12)]. Experimentally, it is found that the discharge time depends to some extent on the capacity. Experimental results for the discharge time as a function of capacity are shown in Fig. 13 for the transistor being

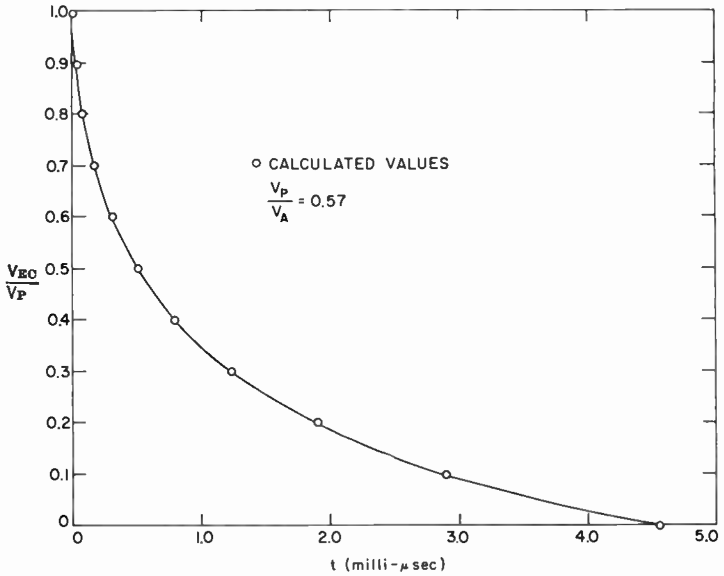


Fig. 12—$V_{EC}/V_P$ as a function of time during discharge of capacity for a transistor with $V_P/V_A = 0.57$.

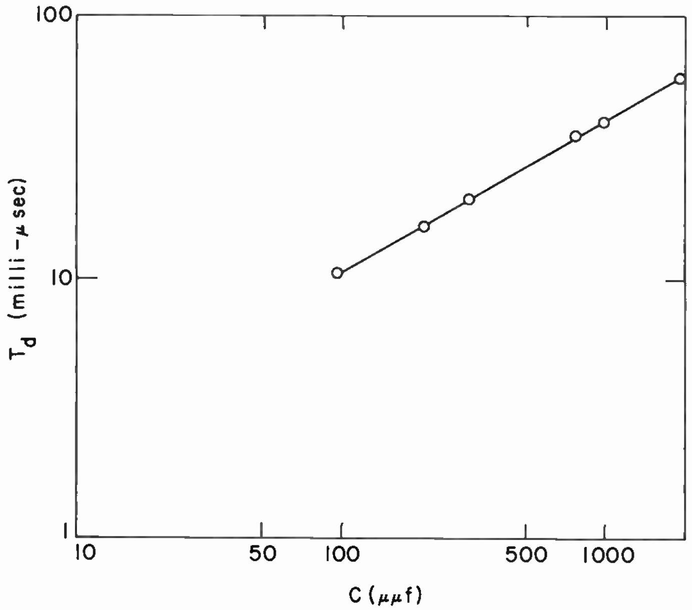


Fig. 13—Observed discharge time as a function of the capacity between emitter and collector.

considered. Because of limitations in the available equipment, values for the discharge time $T_d$ below $10^{-8}$ second could not be measured. It may well be that for very small capacities, the experimental and theoretical values may approach each other very closely.

It should be mentioned that an abrupt change in the behavior of the devices does not occur when the collector voltage is decreased below the punch-through value. Oscillations have been observed at voltages as low as 0.7 $V_P$. In this case the oscillations begin because the extrinsic base resistance is large and $\alpha$ is greater than unity. After the capacity is partially discharged the two space-charge layers from emitter and collector touch and the discharge process described above is repeated. The voltage $V_{EC}$ still goes to zero, which can never occur in a circuit having a regular transistor with $\alpha$ greater than unity and a large external base resistor.

## Conclusion

The above example of a transistor-like structure shows that in these devices where the injection of carriers into space-charge regions and the avalanche multiplication are used simultaneously, the problem of the accumulation of generated carriers in front of the emitting contact is serious. A similar problem may also arise in the first diode structure discussed if the dimensions of the emitting bond are too large. In the transistor-like structure, tendency of accumulation may be reduced by decreasing the size of the emitter.

In addition, this tendency may be reduced even further by using the deflection produced by a magnetic field. Fig. 14 shows the path of electrons and holes when a magnetic field is applied. The angle $\theta$ which the electron current, for example, makes with the direction of the electric field is given approximately by the formula

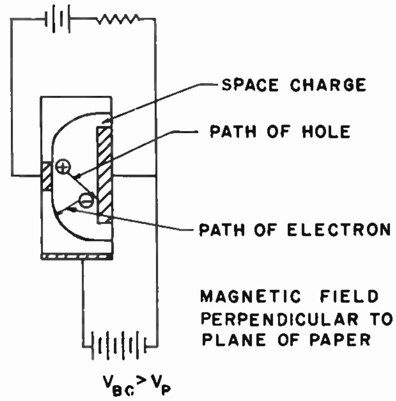


Fig. 14—Path of electrons and holes in a magnetic field.

$$\tan \theta = \mu_n H/c. \quad (16)$$

$\mu$ is the mobility of electrons in esu, which is assumed to be constant; $H$, the magnitude of the magnetic field in emu; and $c$, the velocity of light. For $\mu_n = 3600$ cm²/volt second and $H = 2 \times 10^4$ oersted, $\theta \approx 36°$. In applying (16), however, the dependence of $\mu_n$ on the electric field should be taken into account.

At present, it appears that the combination of a small emitter and a magnetic field may result in a useful amplifier. It is important that the relation between $V_{BC}$ and $V_{EC}$ shown in Fig. 4 is maintained. For small emitters, deviations from the relationship shown in Fig. 4 are expected. In practical units a compromise must be made regarding the size of the emitter.


## Acknowledgment

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# Ferrite Apertured Plate for Random Access Memory*

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*Summary*—Plates with an array of apertures molded from rectangular loop ferrite can be used for random access memory. With the relatively close center-to-center spacing equal to twice the hole diameter there is negligible interaction between adjacent holes. The nonconductive ferrite plate can be coated with a conductive pattern to form a "printed" winding linking all holes in series. The plates can be stacked and the whole stack threaded with word address selecting wires. The printed windings serve for reading and writing the digits of the word.

An experimental prototype plate was developed. It has 256 holes of 0.025 inch in a square $0.830 \times 0.830$ inch. A current of 330 ma reverses the magnetization around a hole in 1.5 μsec and produces 30 mv. The hysteresis loop has good rectangularity and the properties of the holes are uniform within ±5 per cent.

The memory plates can be driven by a switch, itself made of plates. Several novel switching principles are proposed: switching over all holes of a switch-row but the selected one, using memory plates in pairs, and operating by a flux-limited fashion. This results in read-out signals free of disturbs, fast access, and large operating tolerances. Conventional current coincident operation is possible also.

The making, testing, and assembling of ferrite memory plates with printed windings requires much less time and labor than corresponding techniques with conventional cores. The plates open possibility of random access memories with capacities of millions of bits.

## Introduction

To STORE digital information, it is natural to provide a discrete cell for every bit. This was not the case in most of the early solutions to the problem of the high-speed memory of reasonably large storage capacity—acoustic delay lines and electrostatic storage tubes—which resorted to storing the bits on nonsegregated areas of a homogenous medium. The advent of magnetic materials with fast switching and square hysteresis loop provided the first practical means of building relatively large numbers of individual storing cells, each being merely a tiny toroidal core.[1-3] The core memory proved much more reliable and better in most respects than the earlier types and is now the classical solution to the problem.

The two directions of magnetic remanence provide natural storage for a bit of information and the square hysteresis loop allows the storing magnetic element itself to participate in the switching required for its selection in a memory system. Some years ago when we first realized that these properties were ideal for a large capacity high-speed random access memory, we sought various artifices to fabricate at once arrays of large numbers of cells. At first we used individual cores mostly to test the ideas of complete memory systems. We found that the making, testing, and assembling of individual cores was not too laborious a task as long as the number of cores was in the hundreds or even thousands. The fact is that, with presently well-evolved techniques, the cost of assemblies with thousands and even hundreds of thousands of cores is comparable to the cost of the auxiliary electronic equipment required for operating the memory. But for larger storage capacities, millions or scores of millions of bits, the fabrication, the testing, and the assembly of separately made individual cores becomes prohibitive.

For this reason, the original investigations for ways to fabricate whole arrays of storage cells was carried on, along with experiments with single cores. Some years ago, we conceived the idea of molding a plate with an array of holes out of ferrite material having a square hysteresis loop. Such a plate can provide the many storage sites inherent in the use of a continuous medium and yet preserve the distinctiveness of the cells. This is because the areas on the plate are artificially defined by the apertures to form separate storing cells. Pierced metal sheets were considered by R. C. Minnick and R. L. Ashenhurst.[4] These authors believed it preferable to etch out individual toroids from a sheet cemented on a bakelite support.[5]

The development of our apertured plate has now reached a stage which opens possibilities of memories of very large capacities—millions of bits. Because it requires much less driving power, it promises also to reduce and simplify the associated electronic circuits. Furthermore, it makes possible very compact memories of relatively small capacity.

This paper describes the principle of operation of the aperture plate, its fabrication by molding, means of making printed windings, and the characteristics of an experimental prototype plate with $16 \times 16 = 256$ apertures of which thousands were made. A system is proposed to drive the memory plates by a switch, itself made from plates. The operation of the system is based on several novel switching methods which make possible large storage capacities, read-out signals free from disturbs, fast access, and large tolerances in the amplitudes of the electronic driving circuits. The use of the plates in a conventional current coincident operation is considered.

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[5] Since the writing of this paper the author has learned of the work of R. H. Meinken, "Characteristics of a Memory Array in a Sheet of Ferrite," reported at the Conference for Magnetism and Magnetic Materials, Boston, Mass.; October 18, 1956.

### The Apertured Plate

Consider a regular array of round holes in a plate of magnetic material having a perfect square hysteresis loop. Let the direction of remanent magnetization around each hole store one bit of information. It should be expected that each hole acts independently as though it were the hole of an individual core, and that there is no effect due to repeated access to adjacent holes. This is for the following reason. For a given current linking an aperture the magnetizing force $H$ diminishes gradually with distance. Near the hole, nearly perfect radial symmetry can be assumed, so that the $H$ lines are circles and the value of $H$ is inversely proportional to the radius. Therefore, there will be some circle of radius $R_0$ within which the magnetizing force is more than some critical value $H_0$ required to reverse the flux in square loop materials and beyond which it is less than this critical value. Hence, there should be complete reversal within the circle and none without. For a judicious choice of the energizing current, bringing this circle to less than half the width of the legs separating adjacent holes, there should be no interaction between adjacent holes.

In reality, there is a family of hysteresis loops and the loops are not completely square so that there is a gradual rather than a discontinuous radial change of flux. To determine whether these deviations from perfect rectangularity would produce detrimental "cross-talk," a number of experimental plates were made, all of as identical a material as was possible to make. Each plate had an array of 9 holes of a given diameter $d$ and spaced at a center-to-center distance $D$, the ratio $d/D$ being different for the various plates (Fig. 1). The current $I_0$ in the center hole was adjusted in every case to produce maximum discrimination under the conventional two-to-one current regime. The percentage change in the signal obtained from the excitation of $I_0$ as a result of repeated excitations of an adjacent hole by a current $I_1$, made equal to $I_0$, was taken as a measure of the interaction. It is apparent from Fig. 1 that the interaction becomes significant only when the holes are so close that the distance between hole peripheries is less than the radius of the holes. The current producing optimum discrimination is seen to increase linearly with hole size except for a slight deviation from linearity for large holes. The resulting output voltage curve also starts by rising linearly with hole size because the optimum current renders the cross section area of the material in which flux reverses proportional to the hole diameter. The voltage reaches a maximum and then decreases as the available cross section area diminishes because the legs between adjacent holes become thinner. The curve of maximum discrimination rises, first slowly, then more rapidly. This is because the region where flux reverses, between the aperture and its nearest neighbors, becomes thinner; therefore the hysteresis loop, which is an average of elementary zonal hysteresis loops, is squarer as the averaging extends over a smaller radial spread.

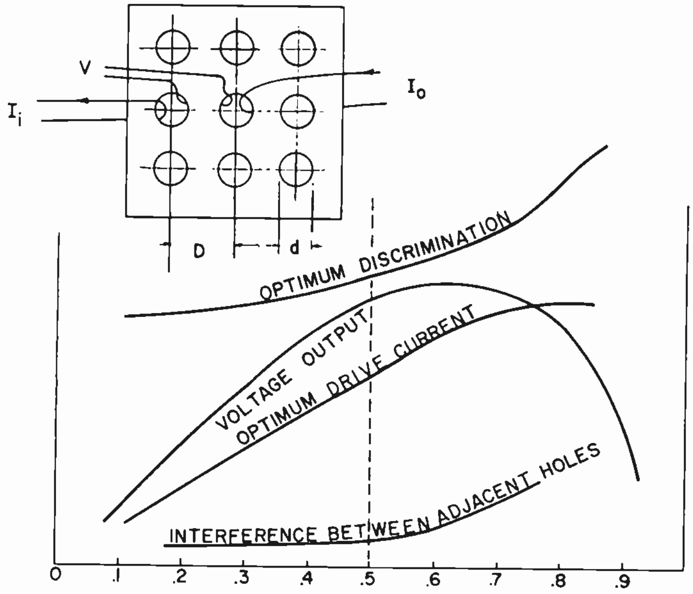



Fig. 1—Effect of hole size and spacing. Ratio of hole diameter to hole spacing.

Because of unavoidable differences in the nature of the material of the various plates and the difficulties of measurements, no absolute numerical significance can be attached to the plots of Fig. 1. However, these plots indicate the general behavior of the interaction, amount of useful signal, and discrimination, as a function of the geometry of the array. The ratio of hole diameter $d$ to hole spacing $D$ of one half is approximately in the center of an optimum range for which there is practically no cross-talk, the signals are high, and the discrimination is relatively large. This simple ratio was adopted for an experimental prototype plate.

### The Experimental Prototype Plate

A prototype plate design was adopted for most of our experiments. The plate has an array of $16 \times 16 = 256$ holes. The holes are 0.025 inch in diameter and are spaced 0.050 inch center-to-center. The plate is a square $0.830 \times 0.830$ inch. (See Fig. 2.) The plates were molded with the array of holes. This required fairly elaborate punches and dies. Considerable experimentation was necessary to develop the right powder, composition, binder, and consistency to obtain satisfactory molding and satisfactory magnetic properties. The firing of the plates is somewhat more critical than the firing of individual cores. This is mostly because precaution must be taken to keep the temperature uniform throughout the area of the plate. Prefiring in tunnel-type furnaces in air, followed by final firing in controlled atmospheres, yielded uniform plates with fairly square loop characteristics. These characteristics are described in more detail below.

### The Printed Winding

The ferrite plate is nonconducting, in fact a fairly good insulator. It is possible to coat it with a metallic

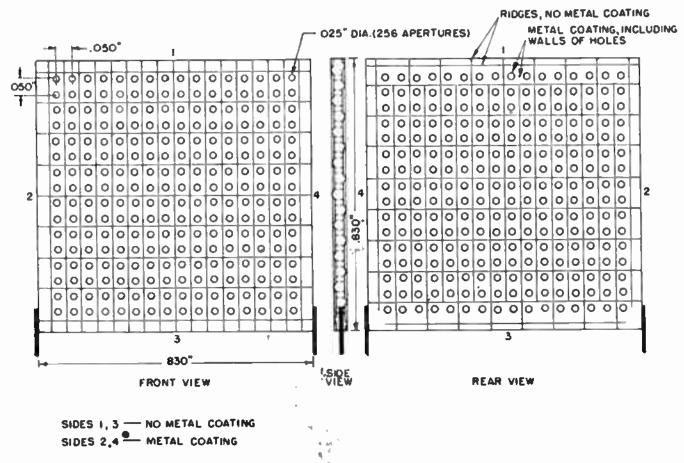



Fig. 2—The experimental prototype apertured ferrite plate.

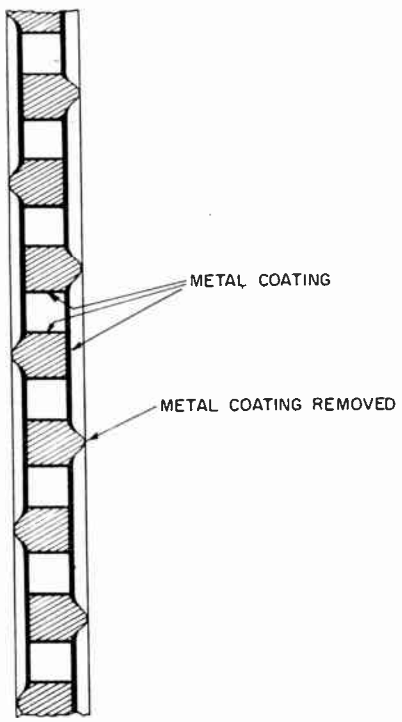



Fig. 3—Enlarged cross section of ferrite plate.

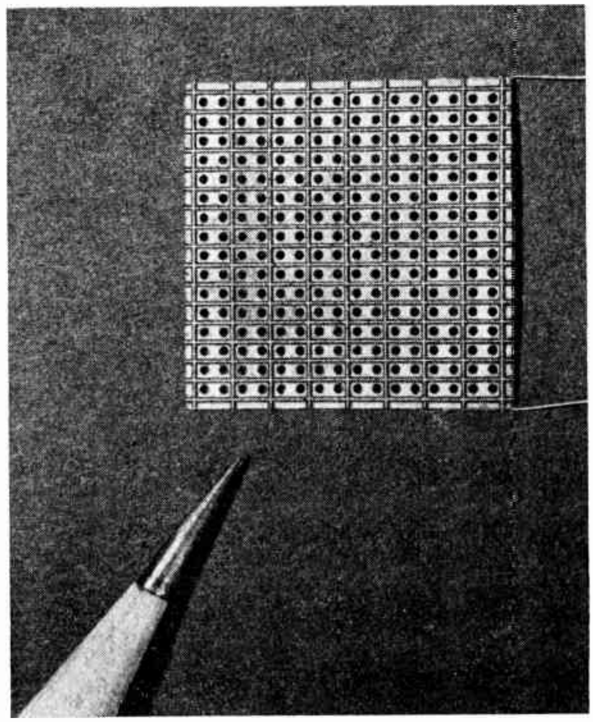


Fig. 4—Photograph of experimental prototype memory plate with an array of $16 \times 16 = 256$ apertures.

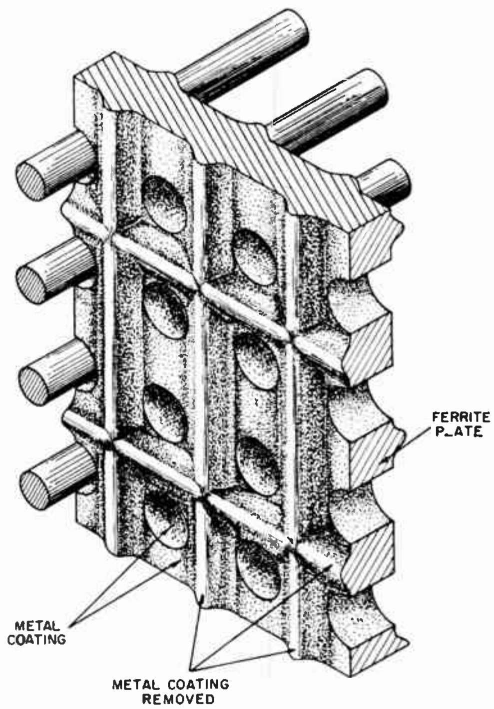



Fig. 5—Perspective view of section of prototype plate.

conductive layer so as to link the holes and thereby obtain a "winding" without the necessity of threading a wire through the holes. This method is particularly suitable for making a winding which links all the holes in series. To obtain such a winding the ferrite plate is first coated with a metal conductive layer on its entire surface including the inside walls of the holes. Then, the coating is removed along cross-lines so that little islands of metal remain on the surface of the plates. (See Figs. 2–5.) Each island is a rectangle comprising two holes. The rectangles on one side of the plates are staggered with respect to those on the other side. The continuity of current flow is, therefore, through one island, then through the coating of the walls of the hole, then through the island on the other side, back through the coating on the walls of the next hole to the next island, etc. (See Fig. 3.) The pattern of islands is so designed that a winding, linking all holes in series, is obtained. The removal of the metallic coating along appropriate straight lines can be done in various ways such as grinding grooves, using photo-engraving techniques, etc. A particularly convenient method is to mold the plates with a pattern of ridges corresponding to the lines along which the coating is to be removed. These ridges are only a few thousands of an inch high. After spraying the entire surface of the plate it is merely sufficient to remove the metal coating on the top of the ridges to obtain the desired pattern. Fig. 4 is a photograph of the plate with its printed winding. The two terminals are 0.007-inch bare copper wires, silver pasted to two sides of the plate. (Slides 2 and 4 on Fig. 2.) The silver on the two other sides (1 and 3) of the plate is removed to prevent a short between the terminals.

The plates undergo considerable shrinkage during firing, about 20 per cent in linear dimensions. The exact final size depends on the nature of the ferrite, the pressure of molding, and the heat treatment. It is found to be remarkably constant, perhaps within $\pm 0.001$ inch for the 0.830-inch side of the plate, provided accurate control of parameters is maintained. But during the developmental stage, when different materials and processing are used and the size may vary considerably, the ridges are a particularly convenient way of insuring that there is no misalignment between the pattern of lines and the pattern of holes. The ridges have also another important advantage in that they tend to strengthen the plate, particularly when it is very thin. Plates were made with a thickness of only 0.007 inch at the significant location around the holes, although a thickness of about 0.020 inch was selected for the prototype plates.

## Plate Testing

After being provided with its printed winding, the entire plate can be tested easily in a single operation by means of a special tester. The tester has an array or "forest" of parallel conductors over which the plate is inserted. The electric circuit of the conductors is closed by shorting the conductors after insertion of the plate. The pins are energized serially by means of a magnetic switch, itself made of plates (of a type described below). The outputs from the successive holes are obtained from the printed winding and are conveniently displayed as a superposition of traces on an oscilloscope. A glance at the bundle of traces suffices to determine whether or not the plate is acceptable. An experimental semiautomatic tester was built (Fig. 6) which had only 128 pins so that two insertions are necessary to test the whole array of 256 holes.

## Plate Characteristics

Many thousands of prototype plates were fabricated in an experimental batch. Fair uniformity was observed on sample tests of about a hundred plates each. Typically, the amount of flux reversed around the aperture for a given current excitation varied $\pm 5$ per cent from the average, within a given plate as well as between plates.

The operation under the conventional two-to-one current drive is shown on Fig. 7. An aperture was energized by a train of current pulses

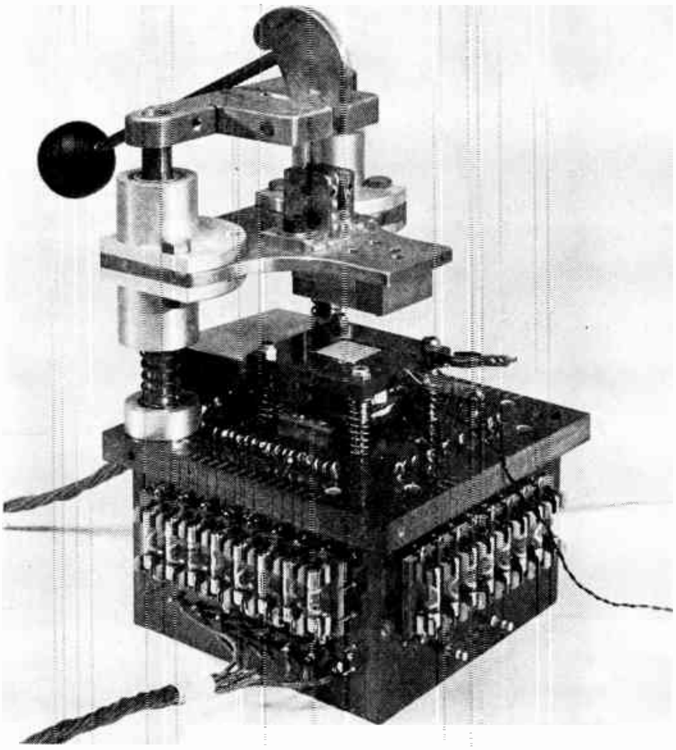


Fig. 6—Semiautomatic experimental plate tester (photo).

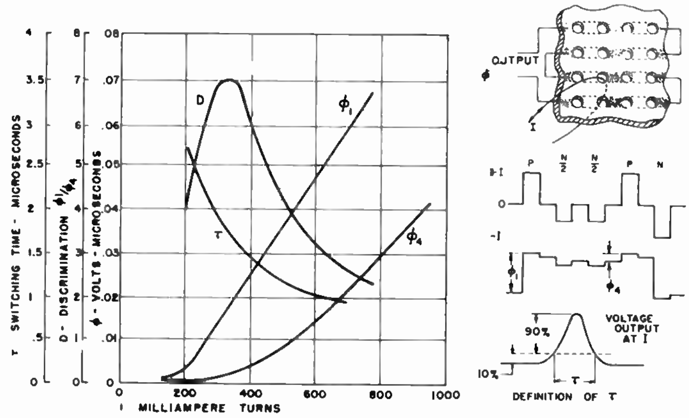



Fig. 7—Flux changes, discrimination, and switching time around aperture of prototype plate.

$$\left(P \frac{N}{2} \frac{N}{2} PN \text{ as shown on Fig. 7}\right)$$

used in routine checking of discrimination for conventional cores. The wanted flux reversal $\phi_1$ for a given current excitation and the unwanted disturb flux $\phi_4$ obtained after two disturbs of half the given current excitation are plotted as a function of the current excitation. These flux reversals do not include the reversible part subtracted at the end of the current pulses and thus include both reversible and irreversible flux changes. The ratio of wanted-to-unwanted fluxes $\phi_4/\phi_1$ exhibits a maximum for an excitation of about 330 ma. The switching time $\tau$ is plotted also as a function of $I$ and is seen to be about 1.5 $\mu$sec for the nominal excitation of 330 ma.

Read-out signals obtained from the printed winding are shown by the photograph of oscilloscope traces on Fig. 8. The full current was at the nominal amplitude of 330 ma. The wanted signal 1 occurs at the first positive $P$ pulse, the disturb signal 4 of interest occurs at the second positive $P$ pulse after two half amplitude negative demagnetization pulses. The discrimination of this experimental ferrite plate is comparable to that of early cores and somewhat inferior to that of present day relatively evolved cores. On the other hand, the aperture of the plate requires about 3 times less current and only about $2\times10^{-8}$ joules or about 5 times less energy than the conventional memory core. This economy results from the fact that it is practical to use only a very small volume of material in a plate of manageable dimensions, while an individual core of the same volume would be unmanageably small.

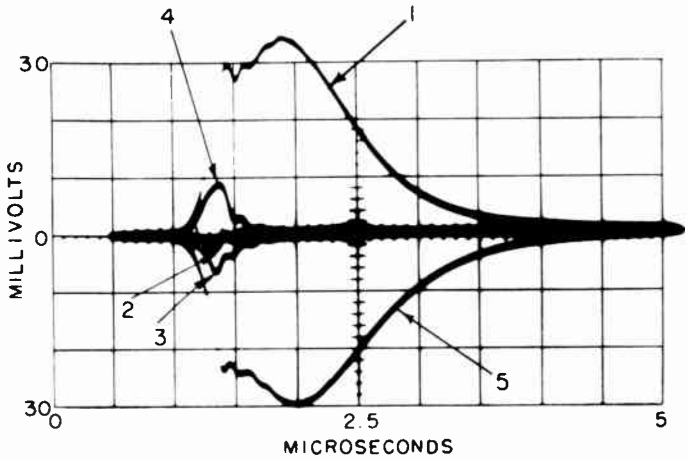



Fig. 8—Read-out signals from the aperture plate.

The observed amount of "cross-talk" between adjacent holes of the prototype plate is very small, as illustrated by the plots of Figs. 9 and 10. The spread of flux from an energized aperture outward is apparent on Fig. 9 which is a plot, as a function of current amplitude, of the total flux $\phi_1$ around the aperture, the flux $\phi_2$ in the region up to the nearest adjacent holes, and finally of flux $\phi_3$ in the next surrounding zone. This spread is as should be expected by assuming that the flux reverses up to a critical circle of radius proportional to current. If the current is kept below the value for which the circle bisects the leg between adjacent holes, there should be no interaction, as was mentioned. This is verified approximately by the measurements of the variations of flux in a leg between two adjacent holes as a function of current excitations in one and then the other hole, as shown on Fig. 10. The entire plate is first brought to a standard condition of flux distribution by a current $I_0$ through the printed winding (directions of flux around the holes in a checkerboard pattern). Then a current $I_1$ is applied through one of the holes and produces a flux change $\phi_A$ in the intermediary leg. If, before the flux is restored by a new pulse $I_0$, a pulse of current $I_2$ (equal to $I_1$) is applied to the other hole, the resultant flux change $\phi_B$ in the intermediary leg has not, in general, the same value as $\phi_A$, since some of the flux in that leg was previously reversed by the pulse $I_1$. However, the values of $\phi_B$ and $\phi_A$ are almost identical for values

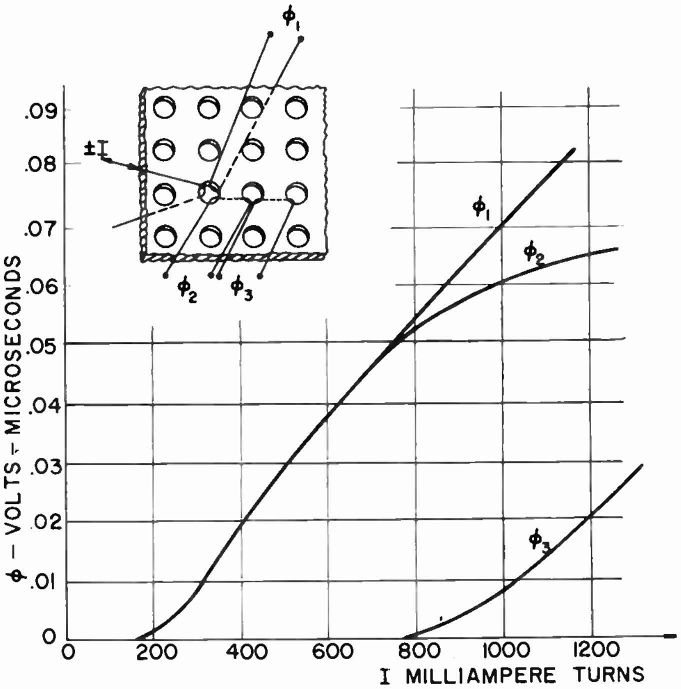



Fig. 9—Flux spread in apertured plate.

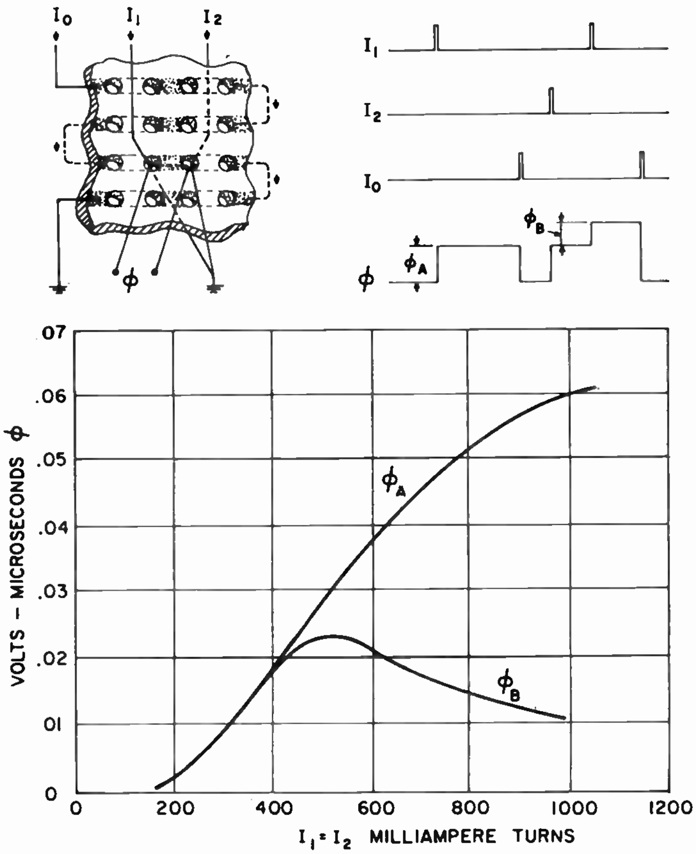



Fig. 10—Effect of access to adjacent hole.

of current that produce about half the total possible flux change in the intermediary leg. For the optimum operating current of about 330 ma, the value of $\phi_B$ is only slightly less than $\phi_A$, showing that there is only negligible cross-talk between the apertures which, for all practical purposes, can be considered to operate independently. A considerably larger current, about twice the optimum value, may still be used, even though it will reduce, in general, the strength of the signal obtained from an adjacent aperture.

## Current Coincident Plate Memory

For most applications of high-speed memories, random access to a word composed of a number $M$ of bits is desired. This number $M$ is typically between a few and a hundred. The plates are particularly suitable for such a parallel memory as they are conveniently stacked with all their apertures in perfect register. The stack is wired as a unit with address selecting wires so that the same location in each of the $M$ plates (or plate pairs) is reached simultaneously. The printed or "digit" winding on the individual plates serves for read-out and write-in of the digits of the word.

The plate memories can be operated in a current coincident mode, just as do the conventional core memories. The holes of the stack of plates are threaded back-and-forth by rows $X$ and columns $Y$, as shown on Fig. 11. This threading is not too exacting a task, even though the holes are only 0.025 inch or half the ID of conventional memory cores, because the wires go straight through the holes which are in perfect axial register.

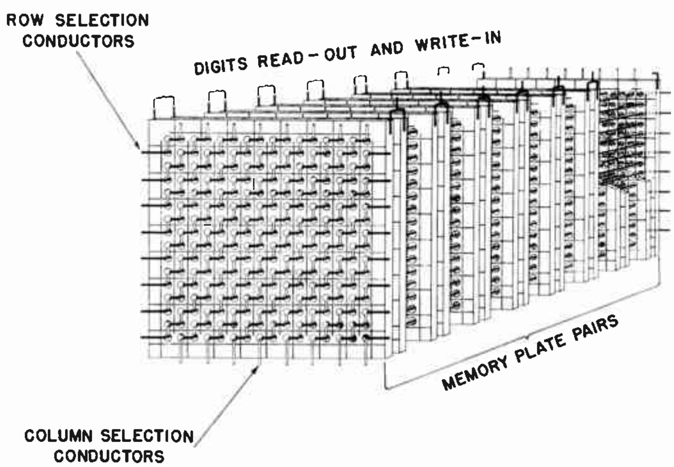



Fig. 11—Current-coincident ferrite plate memory.

This wire threading is a unique operation and need not be repeated separately for each plane, as is customary with memory core planes. The numerous separate digit windings, so tedious to thread in core planes, are obtained by the printed technique on the plates. It is apparent therefore that the complete "wiring" of the stack of plates is much simpler than with individual cores. However, only three conductors are provided in each aperture: the printed digit $Z$ winding in the form of metal coating on the wall of the hole and the $X$ and $Y$ selecting wires. It is not convenient (although possible with a modification of the printed technique) to provide two digit windings, one for read-out and one for write-in, as is customary with core planes, so that the single $Z$ winding must serve for both functions. This poses some problems.

It will be recalled that in most current-coincidence operating modes there are two cycles: a reading cycle in which the selecting wires $X$ and $Y$ are energized with $+I/2$ and the signal is read-out from the $Z$ winding, and a writing (or rewriting) cycle in which the selecting wires $X$ and $Y$ are energized with $-I/2$ and the $Z$ windings of the various planes are either not energized or energized with inhibiting currents of $+I/2$ depending on the nature of the digits to be written in. The logic of the writing requires that the three windings $X$, $Y$, and $Z$ link every aperture in the plate in the same sense. This is achieved by threading the $X$ and $Y$ windings back-and-forth through the stack to conform with the physical checkerboard pattern of the $Z$-printed windings, as is shown on Fig. 11. Therefore, during the read-out, the voltages induced by the half excitations of the apertures on the selected row and column, add up, rather than tend to cancel each other as in conventional read-out windings, and produce an appreciable masking disturbance. This disturbance could be neutralized by a fixed voltage derived from an auxiliary core excited at the instant of read-out.

A better way to cancel the disturb signals is to use a pair of plates for every digit. A bit is stored by magnetizing around an aperture in one plate in one direction, and around the corresponding aperture in the other plate, in the opposite direction. The $Z$ windings of the plates are connected in series opposition, as shown on Fig. 11. Therefore on reading, the disturb signals of the two plates tend to cancel each other, and the polarity of the net signal is indicative of the stored bit. For writing, the inhibiting current is sent through the $Z$ winding of one or the other plate. The use of a pair, instead of a single plate per word digit, is reasonable and not as extravagant as would be the use of two-core arrays. The cost of the molded plate with its printed winding is inherently low and the doubling of the number of plates has no effect on the labor involved in threading the stack.

To obtain storage capacities with more words than there are holes in a plate, an array of plates can be used for each digit, the $Z$ windings being connected in series. The threading of the $X$ and $Y$ wires in the larger stack of plate arrays can be made in a single operation or by connecting together individually threaded plate stacks. The limit to the possible number of words depends mostly on the method used to overcome the effect of the disturbs. Any of the methods developed for core memories are applicable. For example: strobing at a time where the disturb voltage has decayed and the voltage due to the slower irreversible flux is near maximum, dividing the read cycle in two to allow the disturb due the excitation of $X$ to decay before $Y$ is excited, or integrating over a positive and negative excitation cycle. The hysteresis loop of the plates is almost as square as that of the conventional individual cores, but the uniformity of properties between apertures is not quite as good as that resulting from the most severe core segregation, thus a somewhat greater reliance must be laid on the above methods.

## Switch Driven Plate Memory

In the early stages of development of the ferrite apertured plate, there was some doubt whether sufficient loop rectangularity could be obtained for operating by current coincidence. A method tolerating poor plate characteristics was devised. It consists of driving the memory plate stack by a switch which energizes the selected word location without half exciting any other locations. In this way the function of address selection is performed by the switch and that of storing by the memory stack. This method is advantageous even when the presently available prototype plates with good rectangularity are used, because the read-out signal is inherently devoid of any disturbances and is simpler to detect, the operation can be faster, and the drives can have larger tolerances.

The address selecting switch is itself made of a stack of plates identical with those of the memory stack, but without the printed windings, as shown on Fig. 12. The

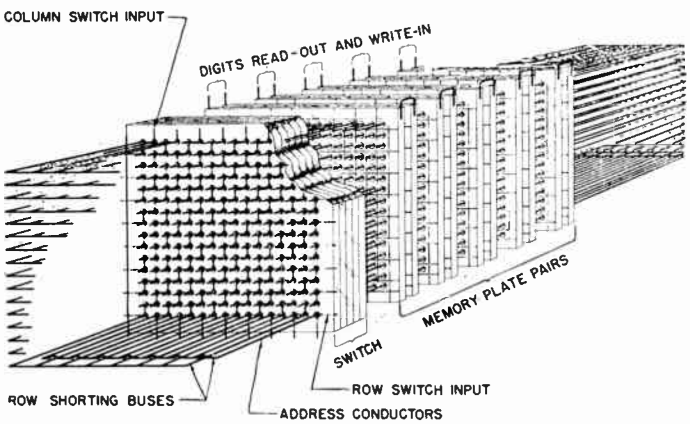
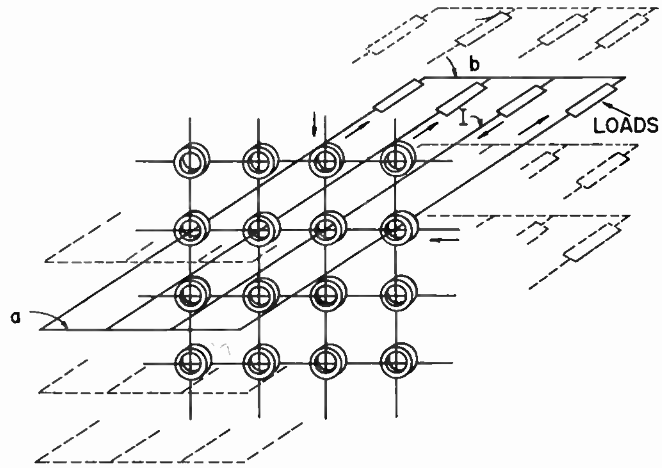



Fig. 12—Switch driven ferrite plate memory.

switch stack is threaded by $X$ and $Y$ windings, and is set in perfect geometrical register with the memory stack. Both stacks are threaded through straight address selecting conductors, one for each aperture. The threading of this "forest" of wires is relatively simple as it amounts to dropping a straight wire into each of the apertures. The ends of these wires are shorted by rows in a manner to be described.

The "end-on" switch can produce outputs on one of the address selecting conductors as a result of the energization of the corresponding row $X$ and column $Y$ in a number of ways. For example, it could be biased with a pulsed or dc inhibiting drive overcoming one of the lines drives, or it could be driven by driving the selected row and inhibiting all columns but the selected one, as was done in previous types of memory driving switches.

A novel method called "set-a-line" was found which has many advantages over previous types. Consider an array of switching elements, cores or apertures in a plate, shown for simplicity as an array of $n \times n$ cores on Fig. 13 ($n=4$). The loads to be driven by the switch




(a)

| | | | |
|---|---|---|---|
| N | N | N | N |
| N | N | N | N |
| N | N | N | N |
| N | N | N | N |

(b)

| | | | |
|---|---|---|---|
| N | N | N | N |
| P | P | P | P |
| N | N | N | N |
| N | N | N | N |

(c)

| | | | |
|---|---|---|---|
| N | N | N | N |
| P | P | N | P |
| N | N | N | N |
| N | N | N | N |

(d)

Fig. 13—Operation of set-a-line magnetic switch. (a) and (b) initial; (c) set-a-line; (d) intermediary.

and assumed of equal impedance $Z$, are connected in parallel by rows, as shown on Fig. 13(a). Let all cores be initially in a normal state $N$ (Fig. 13(b)). In a first step, a current pulse is applied to the selected row tending to drive all cores of the row toward $P$ and at the same time (or with a slight advance) another pulse is applied to the selected column tending to drive all the cores of the column towards $N$. As a result, all cores of the selected row are switched to the state $P$, except the one on the selected column which is kept in its initial state by the column inhibiting current. This produces the intermediary pattern of core states, Fig. 13(d). The voltage $V_1$ induced on the $(n-1)$ cores which are switched over, causes currents to flow in the corresponding loads and returns through the selected load. The equivalent circuit is that of a source $V_1$ connected in series with the selected loads and the remaining $(n-1)$ loads in parallel. Therefore the current $I$ in the selected load is $+(V_1/Z)(n-1/n)$ while the current in each of the remaining $(n-1)$ loads is $-(V_1/Z)(1/n)$. In a second step, a current is applied to the selected row tending to restore all cores to $N$. The inhibiting current on the selected column may or may not be left on during this step. This causes the $(n-1)$ cores which were switched in the first step to switch back and induce voltages $V_2$ and thereby cause the flow of a current $I$ in the selected load equal to $-(V_2/Z)(n-1/n)$ and a current $+(V_2/Z)(1/n)$ to flow in the remaining $(n-1)$ loads. For both steps the currents in the unselected loads are $(n-1)$ times smaller than the selected load current and therefore negligible in most practical cases when $n$ is fairly large. The rate of change of flux of the switching cores can be made very high as the logic of the switch operation does not depend on the amplitude of the drive row current which can be made as large as desired. Of course, the inhibiting column current must be made consequently larger. Therefore the voltages pulses $V_1$ and $V_2$ can be made of high amplitude and short duration. For symmetrical excitation $V_1$ and $V_2$ will be symmetrical, *i.e.*, of equal and opposite amplitude. Asymmetrical operation is possible also, yielding a high and short positive pulse followed by a low and long negative pulse or vice versa.

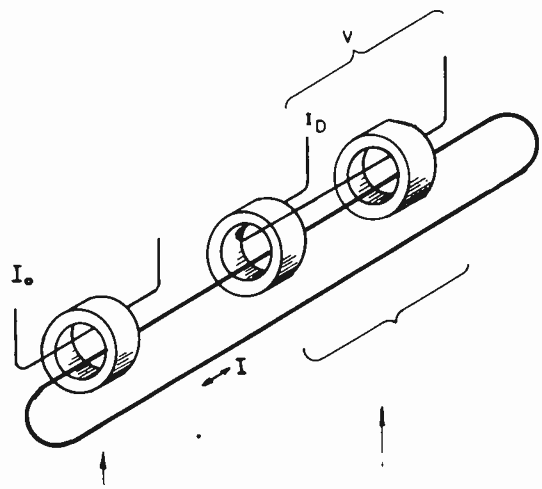
The polarities of energization of the load address conductors must alternate along rows and columns to conform with the checkerboarding due to the printed $Z$ windings of the memory plates. This determines the back-and-forth threading of the row and column windings on the switch and the use of interlaced shorting row buses on the ends of the address conductors, shown on Fig. 12.

It is of interest to note parenthetically that the first operating step could be split in two; a row drive and a subsequent column drive. The row drive switches over or "sets" all cores of the selected row and brings about the pattern of Fig. 13(c). These switch-overs produce a voltage between the shorting buses $a$ and $b$, but no load currents, as there is no available return path for the current. The subsequent column drive brings about the intermediary state shown on Fig. 13(d) and causes the same currents to flow in the loads as was the case when the row and drive pulses were simultaneous. This mode of operation involves no coincidences of driving currents. It may be useful in cases where the logic of addressing furnishes row information prior to column information.

The switch produces very small disturbing currents in the unselected loads corresponding to the selected row and column. On the row, these currents are due to the fact that the unselected loads are a part of the return circuit for the current of the selected load. They are small for all practical cases when there are many columns (*i.e.*, $n>8$), as was mentioned. On the column, there are small disturbing voltages due to reversible flux changes around the apertures driven further into saturation by the column inhibiting current. Therefore the read-out signal from the switch driven plate memory is almost solely due to the drive of the selected address conductor and is practically devoid of any masking disturbance.

A pair of memory plates is used for every bit of the word. A bit is stored by magnetizing around corresponding apertures of the two plates in opposite directions, in one way or the other. This pairing was proposed for cancelling masking disturbance on read-out in the coincident current operation, but is not necessary for that purpose in the switch drive mode since there are practically no such signals. The purpose of the pairing is primarily to render constant the load presented to the switch by the memory. Irrespective of the nature of the stored word, there will be precisely $M$ apertures around which flux reverses irreversibly and $M$ apertures around which flux reverses reversibly, since in every plate pair the two apertures are left in opposite states.

The use of plates in pairs permits operation of the system in a "flux-limited" fashion. To understand the system consider first the case of three identical square loop cores linked by a shortening loop of negligible ohmic resistance, as shown on Fig. 14. Initially, the




| | SWITCH CORE | MEMORY CORE | |
|---|---|---|---|
| INITIAL | N | N | P |
| READ | P | N | N |
| REWRITE OR WRITE | N | N | P |

Fig. 14—Flux limited switch drive.

switch core and one of the memory cores are assumed at $N$ and the other memory core at $P$. In a first, read-out step, the switch core is reversed from $N$ to $P$ by a drive current $I_0$. Because the ohmic voltage drop in the shorting loop is negligible, the change of flux in the switch core must produce an equal total irreversible flux change in the memory cores and therefore cause the memory core initially at $P$ to switch to $N$, bringing both cores to $N$. The polarity of the read-out voltage induced on a digit winding linking the two cores in series opposition identifies which of the two cores switches over. In a second, write-in step, the switch core is restored from $P$ to $N$ by a drive current $I_0$ of the opposite polarity. At the same time, or preferably with a slight advance, a current $I_D$ is sent through the digit winding in one or the other direction depending on the digit to be written or rewritten-in. This current $I_D$ favors the turning over of one of the memory cores and hinders that of the other. Consequently the flux from the switch core will be transferred to the memory cores in unequal parts, a larger part to the favored and a lesser part to the hindered core. For a sufficiently large amplitude $I_{D0}$ of the digit write current $I_D$, all the flux will appear in one of the cores. For smaller amplitudes, there will still be a difference between the fluxes transferred to the two cores and consequently a logical write-in which can subsequently produce a read-out signal of correct polarity even though it is of reduced amplitude.

The example of the three identical cores illustrates the operation at one address of the flux-limited end-on switch driven plate memory system. Let the number of plates making up the switch be such that the maximum amount of flux that can be reversed in the switch be just equal to the amount of flux to be reversed in $M$ memory plates, one for each pair. When plates of the same thickness as the memory plates are used this number is equal to $M$. Fewer thicker plates can be used also. The elements of the idealized circuit of Fig. 14 are approximated at every location as follows: the switch core by the $(n-1)$ row apertures around which the flux reverses, the shorting loop by the address conductor in series with the $(n-1)$ other conductors in parallel, and the two memory cores by the $2M$ apertures of the memory plates. The circuit includes also the selected aperture of the switch, which does not switch-over during selection, and which adds negligible impedance to the loop.

The flux limited mode of operation has a number of advantages. Because the amount of flux transferred to the memory is strictly limited by that available in the switch, very large drive currents $I_0$ (and consequently large but short loop current pulses $I$) can be used to obtain fast switch-over and large read-out voltages for relatively small stored-in write energy. Considerably faster speeds are possible than those obtained with a nominal current adjusted for maximum two-to-one discrimination. There is considerable tolerance in the amplitude of the write-in digit current. This is illustrated by typical plots of the net read-out signal (difference in flux between the two plates of the pair) as a function of that current. (See Fig. 15.) The middle full curve is obtained when the two plates of the pair are identical and the dotted curves are indicative of curves that could be obtained with plates differing in thickness or some other parameter. There is a minimum to the allowable digit write-in current necessary to compensate for possible differences between plates. There is a maximum above which the write-in digit current would produce irreversible demagnetizations in the nonselected apertures of the plate. This maximum is considerably greater than the value for which substantial read-out signals are obtained. The allowable extreme values depend somewhat

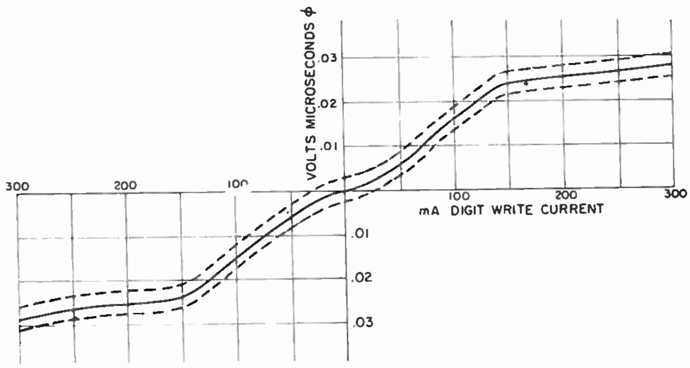



Fig. 15—Output signal as a function of digit write current.

on the switchover time. In practice, there is a tolerance of about two to one in the amplitude of the digit write-in current.

## Associated Circuits

The plates are provided with a single digit winding obtained by a fairly simple printed technique. The more complex techniques for "printing" two windings are not necessary because the use of a pair of plates per digit makes it possible to design fairly simple circuits operating on a single winding for both sensing the read-out and writing the desired digit.

Such a typical circuit is shown by the block diagram of Fig. 16. The digit winding is coupled through a trans-

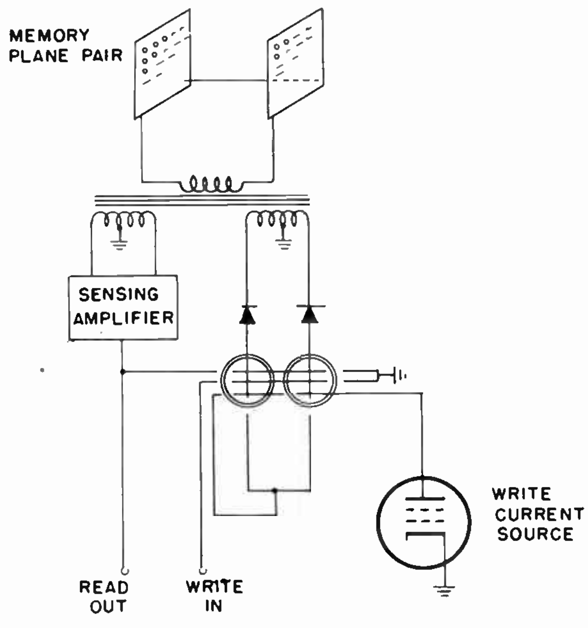



Fig. 16—Digit sensing and writing circuit

former to a sensing amplifier which produces a current of one or the other polarity depending on the polarity of memory output. This current sets an auxiliary pair of square loop cores, one core in state $N$ and the other in state $P$ or vice versa. Rewriting is obtained by steering a current from a driving source through one or another of two windings wound in opposite directions on the transformer and thereby inducing a current in the digit winding in one or the other direction. During rewrite the sensing amplifier is gated off. For writing new information, the auxiliary cores are set by the write-in signal instead of the sensing amplifier. It is convenient to use transistors for the sensing amplifier and a tube drive current source common to all digit circuits.

The address circuits for the plate memory can be one of the many types used with core memories. In general the requirements are less severe. In the current coincident mode considerably less driving power is required than that required for the core memory, because the energy stored per bit is about 5 times smaller around the aperture of a plate than in a conventional core. Also the smaller amplitude of the row and column currents gives a better match to tubes or transistors. The switch driven plate memory requires more power because the switch itself dissipates power. However, the very large tolerances of the driving currents make possible simpler circuits.

## Conclusion

Apertured ferrite plates are proposed for building high speed random access memories. They provide a simple means to fabricate large numbers of discrete storing cells in convenient packages which can be easily assembled into large memory systems. Furthermore, the cells can be made small so as to require little energy to store each bit.

A prototype experimental plate and means to use it in memory systems were developed which realize to a large extent the expectations of this proposal. The findings of the development may be summarized as follows:

1) The prototype is a molded square plate 0.830 $\times$0.830 inch having $16\times16=256$ apertures 0.025 inch in diameter and spaced on 0.050 inch centers. Measurements show that:
   a) There is negligible interaction between apertures.
   b) In lots of hundreds of plates taken from thousands the properties between apertures of a given or different plate are approximately within $\pm5$ per cent of an average value.
   c) The hysteresis loop around each aperture has good rectangularity.
   d) At 330 ma drive the flux reversal around each aperture occurs in 2 $\mu$sec and produces a peak voltage of 30 millivolts. The energy stored per bit is less than $2\times10^{-8}$ joules.
2) The fabrication, the testing, and the assembly of the memory plates is relatively simple and economical.
3) The memory plates can be used with conventional two-to-one current coincidence drive. This is particularly suitable for compact, lower power consuming, transistor driven, memories of relatively small capacity.
4) A new method is proposed of driving memories particularly suitable for plates. The memory plates are driven by an "end-on" switch itself made of plates. The switch operates in a "set-a-line" and "flux-limited" fashion and the memory uses plates in "push-pull" operated pairs. This results in a mode of operation permitting fast access, large storage capacities, and allowing large tolerances in the timing and amplitudes of driving currents. The systems of switching developed for the plate memory have broader utility. The use of a pair of memory cores driven by a switch core of limited flux may find applications in fast core memories and logic circuits.

We ventured to predict three years ago[3] that microsecond random access memories with capacities reckoned in millions of bits, megabits, would be available at a relatively low cost in a distant future. We believe that the ferrite aperture plate is now ready to usher in the era of such memories. It will follow the present era of core memories with capacities of hundreds of thousands of bits. But the demand for larger and faster memories is incessant and we may look forward to the development of new techniques making possible storage of billions of bits.

## Acknowledgment

The success of this work is due to many members of RCA Laboratories. C. Wentworth developed the plate processing. W. Wales[6] and G. R. Briggs worked in the early phases of the investigation. H. D. Crane[7] and W. F. Kosonocky worked in the later stages and contributed many ideas. A. W. Lo was active throughout the project and developed the sensing circuit. Most of the mechanical work was done by J. Wallentine.[8] B. M. Quinn and C. H. Morris contributed to the shop work.

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# Stability and Power Gain of Tuned Transistor Amplifiers*

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*Summary*—The transistor is a nonunilateral device which, if appropriately terminated, can become unstable at frequencies where its "internal feedback" is sufficiently large. At such frequencies, the maximum power gain is infinite and the transistor may oscillate. This paper discusses the maximum power gain realizable as a function of a required degree of stability.

A "stability factor" is defined in terms of the transistor parameters and terminations (the admittance matrix is used as an example, but the approach is analogous using other representations). The maximum stable power gain of an isolated amplifier stage and the terminating admittances required for the realization of this maximum power gain are then computed as functions of the stability factor. The computations are extended to include bandwidth requirements and limitations. (It is found that, although bandwidth requirements *may* impose limitations on the power gain, there is no simple relationship tying together bandwidth and power gain.) The treatment of multistage amplifiers is outlined with the conclusion that the gain realizable in an $n$-stage amplifier is smaller than $n$ times the gain of a one-stage amplifier having the same stability factor as the stages of the $n$-stage amplifier. The respective advantages of different representations for different circuit configurations are discussed.

In an appendix, the theoretical considerations are applied to tuned transistor amplifiers in common-emitter and common-base configurations. The stability factor is related to the tolerances in transistor parameters and terminating impedances. Examples are given for the maximum realizable stable gain as function of parameter tolerances.

## INTRODUCTION

TRANSISTORS are *nonunilateral* devices; a signal applied to the output port of a transistor amplifier results in a response at the input port. The existence of *internal feedback* is expressed by the fact that, if the transistor is described by the "$z$," "$y$," "$h$," or "$g$" matrices, the matrix element having the subscript 12 (*i.e.*, $z_{12}$, $y_{12}$, $h_{12}$, or $g_{12}$) is different from zero.

If properly terminated at its ports, a device having sufficient *internal* feedback may become unstable (*i.e.*, may oscillate) even in the absence of *external* feedback. For example, in the "tuned plate-tuned grid" vacuum tube oscillator, feedback is provided entirely by the grid-to-plate capacitance of the tube.

It has been shown[1,2] that transistors exhibit *potential instability* within certain frequency ranges. The frequency range of potential instability is different for the three transistor configurations (common-emitter, -base and -collector).[3]

At frequencies where the transistor configuration is unconditionally stable (*i.e.*, where the transistor cannot become unstable, no matter what passive terminations are used), the maximum available power gain can be calculated.[1] However, at frequencies, where potential instability exists, the transistor may oscillate and the maximum power gain is infinite. Consequently, at such frequencies, the unqualified "maximum available power gain" is not a very useful concept. On the other hand, engineers engaged in the development of transistor circuitry do know very well that a practical measure for the maximum power gain realizable in a stable amplifier is indeed useful and very much needed. Furthermore, experience shows that such a maximum stable gain does actually exist.

Various attempts have been made to define a practical measure indicative of the maximum realizable power gain. For example, it has been shown that the transistor has a finite maximum gain if conjugately matched at its output, the termination at the input being either a pure resistance[4] or a resistance in combination with a reactance tuning out the reactive component of $h_{11}$.[5] It also has been thought that the "neutralized gain" may be a useful measure of stable amplifying capability and, besides other expressions, Mason's $U$ function[6] has been suggested as a practical indication of attainable stable power gain.

These and other approximations for the maximum realizable stable power gain have the common advantage of presenting the practicing engineer with a measure useful as a reference. Their common disadvantage lies in the fact that the physical conditions under which these gain expressions can actually be realized are chosen essentially arbitrarily and are quite different from the situation most frequently occurring (and most desirable) in practice: the case of an unneutralized amplifier using impedance transforming ("matching") tuned circuits at both input and output ports.

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In order to obtain an expression for the maximum realizable stable power gain which is clean and acceptable from both practical and conceptual points of view, it seems desirable to tie in the maximum realizable power gain with a measure indicating the stability of the amplifier. If this is done, the maximum power gain is finite for stable amplifiers and can appear as a *function of the degree of stability* of the amplifier.

The purpose of this paper is to discuss the maximum power gain of transistor amplifiers using a measure of stability which appears to be both simple and practical. The following considerations can be applied directly to other active nonunilateral two-port elements.[7]

## The Stability Factor "$k$"

The following calculations make use of the "$y$" matrix elements (the admittance parameters) of the transistor and external circuit elements are represented as admittances. This procedure is arbitrary; analogous calculations can be carried out using the "$z$" matrix elements and external impedances or the "$h$" (or "$g$") matrix elements and suitable combinations of admittances and impedances. It will be shown that these various representations have their respective advantages depending on the types of resonant circuits used in the tuned amplifier.

If $y_{ij}$ are the admittance parameters of the transistor, one can write, separating real and imaginary components:

$$y_{11} = g_{11} + jb_{11} \tag{1}$$

$$y_{22} = g_{22} + jb_{22} \tag{2}$$

$$y_{12}y_{21} = M + jN. \tag{3}$$

The absolute value of the "internal loop gain" ($y_{12}y_{21}$) is

$$L = |y_{12}y_{21}| = \sqrt{M^2 + N^2}. \tag{4}$$

It has been shown[1-3] that if

$$L + M \geqq 2g_{11}g_{22} \tag{5}$$

the transistor *may* oscillate without external feedback, if properly terminated.

If the transistor is terminated by a generator admittance $Y_G$ and a load admittance $Y_L$ (Fig. 1), both $Y_G$ and $Y_L$ may be complex:

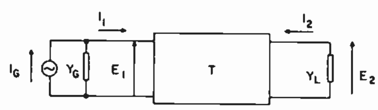


Fig. 1—Transistor with terminating admittances.

$$Y_G = G_G + jB_G \tag{6}$$

$$Y_L = G_L + jB_L. \tag{7}$$

The transistor and the conductive components of the terminating admittances can be considered as constituting a new two terminal pair element $C$ (Fig. 2). Applying

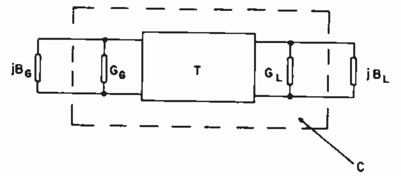


Fig. 2—Transistor and conductive terminating components forming composite network $C$.

inequality (5) to $C$, one finds the condition of potential instability:

$$L + M \geqq 2(g_{11} + G_G)(g_{22} + G_L). \tag{8}$$

Note that (8) does not imply actual instability: in order that the arrangement be actually unstable, $C$ must be terminated by suitable susceptances $B_G$ and $B_L$. However, in most tuned amplifiers, the terminating (or coupling) susceptances are variable and are actually varied when "tuning up" the amplifier. In arrangements having reasonably narrow band, the terminating susceptances will assume a wide range of values. Consequently, it can be assumed that, in most cases, during the tuneup procedure, $B_G$ and $B_L$ will assume values resulting in oscillation, provided that (8) is satisfied. Therefore, for many practical purposes, it can be stated that, if (8) is satisfied, the amplifier is not only *potentially unstable*, but is very likely to become *actually unstable.*

Unconditional stability can be achieved if

$$(g_{11} + G_G)(g_{22} + G_L) > (L + M)/2. \tag{9}$$

In other words, for stability, one requires

$$(g_{11} + G_G)(g_{22} + G_L) = k(L + M)/2 \tag{10}$$

where $k > 1$. The larger $k$, the more remote is the likelihood of instability. Consequently, $k$ can be considered as a measure of stability and will be called *stability factor*.

For any individual transistor, whose parameters are known, stability can be achieved by chosing $G_G$ and $G_L$ such as to make $k$ slightly larger than 1. However, considering that transistor parameters (*i.e.*, the $y_{ij}$) vary with transistor age, dc bias, temperature, etc., that transistors belonging to the same type have slightly different parameters and, finally, that certain variations in $G_G$ and $G_L$ must also be tolerated (especially if the terminations are other transistor amplifier stages), it is usually desirable to design the amplifier with $k$ well in excess of unity in order to insure interchangeability of

[7] An approach similar to the one presented in this paper has been used independently by G. Bahrs in his doctoral thesis at Stanford University, Stanford, Calif.

components and stability under all reasonable conditions of operation. (Values of $k$ up to 10 may be desirable, depending on the various tolerances involved.)

An infinite number of combinations of $G_G$ and $G_L$ correspond to the same value of $k$. Although the degree of stability of all these arrangements is the same, they may lead to different values of power gain. In the following section, the power gain will be maximized for a given value of $k$.

## Analysis of a One-Stage Amplifier

### *The Power Gain*

The transducer gain $G_T$ is defined as the ratio of the power delivered to the load admittance $Y_L$ to the available power of the generator having an admittance $Y_G$. The transducer gain is:

$$G_T = \frac{4|y_{21}|^2 G_G G_L}{|(y_{11}+Y_G)(y_{22}+Y_L) - y_{12}y_{21}|^2}. \tag{11}$$

Defining:

$$g_{11} + G_G = G_1, \tag{12}$$

$$g_{22} + G_L = G_2, \tag{13}$$

$$b_{11} + B_G = B_1, \tag{14}$$

$$b_{22} + B_L = B_2, \tag{15}$$

one finds for a given stability factor $k$:

$$G_1 = k(L+M)/2G_2. \tag{16}$$

Then, the transducer gain is:

$$GT = \frac{N}{D} = \frac{4|y_{21}|^2[k(L+M)/2G_2 - g_{11}](G_2 - g_{22})}{[B_1B_2 - k(L+M)/2 + M]^2 + [B_1G_2 + B_2k(L+M)/2G_2 - N]^2}. \tag{17}$$

It is desired to find the value of $G_2$ which leads to maximum $G_T$ for a given $k$. In order to find $G_T$ as a function of $G_2$ alone, it is convenient to eliminate the two other variables $B_1$ and $B_2$. Since $B_1$ and $B_2$ occur only in the denominator $D$ of (17), this can be done by determining the values of $B_1$ and $B_2$ which minimize the denominator. It should be remembered that, $B_1$ and $B_2$ representing susceptances, must be positive or negative *real* numbers.

The process of minimizing $D$ in (17) can be simplified by a change of variables. One defines

$$y_1 = B_1G_2 + B_2k(L+M)/2G_2 \tag{18}$$

$$y_2 = B_1B_2. \tag{19}$$

$D$ can then be written:

$$D = (y_2 - A)^2 + (y_1 - N)^2 \tag{20}$$

where

$$A = k(L+M)/2 - M. \tag{21}$$

Note that for $k>1$, $A$ is always positive. $B_1$ and $B_2$ can be expressed in terms of the new variables $y_1$ and $y_2$:

$$B_1 = (y_1 + \sqrt{\Delta})/2G_2 \tag{22}$$

$$B_2 = (y_1 - \sqrt{\Delta})G_2/k(L+M) \tag{23}$$

where

$$\Delta = y_1^2 - 2k(L+M)y_2. \tag{24}$$

Since $G_2$ and $k(L+M)$ are always real numbers, $B_1$ and $B_2$ will be real, whenever

$$\Delta \geqq 0. \tag{25}$$

In the $y_1 - y_2$ plane, the curve $\Delta = 0$ separates the region of real values of $B_1$ and $B_2$ from the region in which these quantities are complex. The equation

$$\Delta = y_1^2 - 2k(L+M)y_2 = 0 \tag{26}$$

is that of a parabola in the $y_1 - y_2$ plane (Fig. 3), the

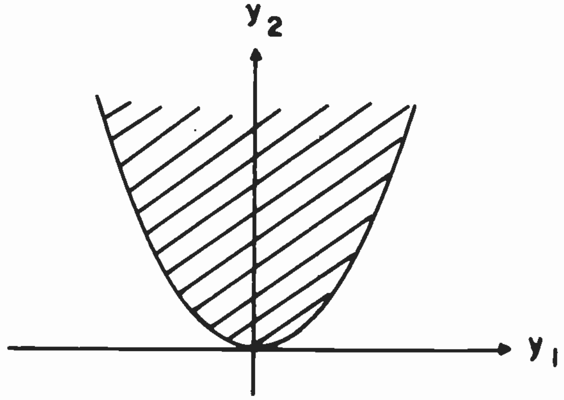


Fig. 3—Parabola separating region of complex $B_1$ and $B_2$ (shaded) from region in which $B_1$ and $B_2$ are real.

shaded area "inside" the parabola representing the region of complex values of $B_1$ and $B_2$ and the area "external" to the parabola being the region of real values of $B_1$ and $B_2$.

In a three-dimensional coordinate system with axes $D$, $y_1$ and $y_2$, (20) represents a paraboloid of revolution, whose axis is perpendicular to the $y_1 - y_2$ plane and intersects this plane at the point $Q(N, A)$. In the same space, (26) is that of a parabolic cylinder. [Fig. 4 shows the parabola (26) and the projection of "constant $D$ circles" on the $y_1 - y_2$ plane.] The "interior" of this cylinder is the space within which $D$ cannot lie if $B_1$ and $B_2$ are real and one is interested to determine the minimum value of $D$ "outside" the parabolic "wall." [The vertex $Q$ of the paraboloid having coordinates $(N, A)$, where $A$ is always positive for $k>1$, will be "inside" the parabolic cylinder, since it corresponds to infinite gain, which of course, is excluded, in view of the assumptions regarding stability.]

An analysis of the geometrical situation shows that the minimum of $D$ (*i.e.*, the maximum of the gain) will

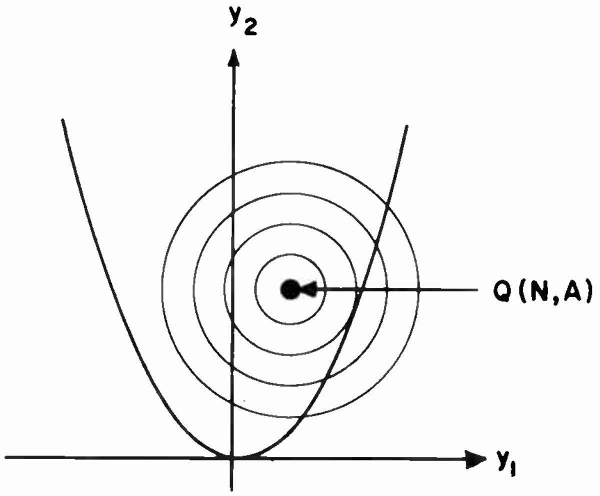


Fig. 4—Parabola (26) and projections of constant D circles on the $y_1-y_2$ plane.

occur along the intersection of the paraboloid with the parabolic cylinder. In Fig. 4, circles with increasing diameter correspond to increasing values of the denominator $D$ (and, consequently, to decreasing values of the gain). The minimum realizable value of $D$ with real $B_1$ and $B_2$ occurs where a circle is tangent to the parabola, *i.e.*, at a point common to both paraboloid and parabolic cylinder. Geometrical considerations also show that two cases are possible:

1) If the location of the point $Q(N, A)$ is such as shown in Fig. 5(a) (with only one point $P_0$ at which a circle corresponding to constant $D$ is tangent to the parabola), the intersection of the paraboloid and the parabolic cylinder has *one* minimum at point $P_0$ corresponding to $D=D_0$.

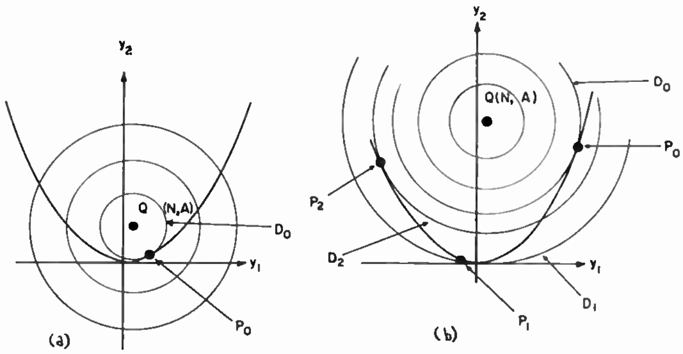


Fig. 5—Situations leading to one minimum (a) and three extrema (b) respectively.

2) If there are three points at which "constant $D$ circles" are tangent to the parabola, there is a minimum at $P_0$ corresponding to a low value $D_0$ of $D$. There is a second minimum at $P_2$ corresponding to the higher level $D_2$ of $D$ and there is a maximum at $P_1$ corresponding to an even higher level $D_1$ of $D$. In this case the maximum gain is given by the smaller minimum occurring at $P_0$.

These qualitative considerations can be put into quantitative form. Since $D_0$ occurs along the intersection of the paraboloid with the parabolic cylinder, one must determine the minimum of (20) with condition (26).

Defining the variable $z$ as

$$z = y_1/\sqrt{k(L+M)} \tag{27}$$

$D$ can be written:

$$D = z^4/4 + [k(L+M) + 2M]z^2/2 - 2N\sqrt{k(L+M)}z + A^2 + N^2. \tag{28}$$

The minimum of $D$ can be found by setting

$$0 = dD/dz = z^3 + [k(L+M) + 2M]z - 2N\sqrt{k(L+M)}. \tag{29}$$

As expected from the foregoing qualitative analysis, (29) leads to two possibilities:

1) (29) has one real root $z_0$, the other two being conjugate complex. In this case $z_0$ corresponds to $D_0$.

2) (29) has three real roots. Two roots ($z_0$ and $z_2$) represent minima of $D$, $z_0$ corresponding to the smaller minimum, while the third root ($z_1$) is a maximum.

In either case, $z_0$ leads to the maximum realizable gain. Substituting $z_0$ into (28), the minimum value $D_0$ of $D$ is found. Note that, since both $z_0$ and $D$ are functions of $k$ and $(y_{12}y_{21})$ only, $D_0$ depends only on $k$ and $(y_{12}y_{21})$ and is independent of other transistor parameters and of conductive terminations.

Assuming now that $D_0$ is known, the gain (maximized with regard to $B_1$ and $B_2$) can be written:

$$G_T = (4|y_{21}|^2/D_0)[k(L+M)/2G_2 - g_{11}](G_2 - g_{22}). \tag{30}$$

Generator and load conductances are assumed to be positive and one must have $G_1 > g_{11}$ and $G_2 > g_{22}$. In other words:

$$k(L+M)/2g_{11} > G_2 > g_{22}. \tag{31}$$

Setting $dG_T/dG_2 = 0$, one finds that for maximum realizable power gain:

$$G_2 = \sqrt{k(L+M)/2}\sqrt{g_{22}/g_{11}} \tag{32}$$

and, using (16):

$$G_1 = \sqrt{k(L+M)/2}\sqrt{g_{11}/g_{22}}. \tag{33}$$

Substituting these expressions into (30), the maximum realizable gain, corresponding to the stability factor $k$, is found:

$$G_{\max k} = (4|y_{21}|^2/D_0)[\sqrt{k(L+M)/2} - \sqrt{g_{11}g_{22}}]^2. \tag{34}$$

Eq. (34) expresses the maximum power gain realizable with a given stability factor $k$. As $k$ increases, $G_{\max,k}$ decreases. Consequently, one can say that increased stability can be achieved by sacrificing gain. It is, however, important to know, that moderate gain does not necessarily imply improved stability: $k$ is determined by the product $(G_1G_2)$ and this product can be chosen such that while $k$ is small (close to instability), the gain is also small due to a disadvantageous ratio $(G_1/G_2)$. In this case gain is sacrificed without improvement in stability.

From (32) and (33) the required values of the terminating conductances are found:

$$G_G = G_1 - g_{11}$$
$$= g_{11}\left[\sqrt{k(L+M)/2}/\sqrt{g_{11}g_{22}} - 1\right] \quad (35)$$
$$G_L = G_2 - g_{22}$$
$$= g_{22}\left[\sqrt{k(L+M)/2}/\sqrt{g_{11}g_{22}} - 1\right].\text{[8]} \quad (36)$$

The required values of the susceptances can be determined, using (22), (23), and (27):

$$B_1 = B_G + b_{11} = G_1z_0/\sqrt{k(L+M)} \quad (37)$$
$$= z_0\sqrt{k(L+M)}/2G_2$$
$$B_2 = B_L + b_{22} = G_2z_0/\sqrt{k(L+M)} \quad (38)$$

where $G_1$ and $G_2$ are given by (32) and (33). Eqs. (37) and (38) show that $B_1$ and $B_2$ have identical sign (that of $z_0$). This, of course, is not necessarily true for $B_G$ and $B_L$.

## Bandwidth Limitations

The preceding computations were carried out without regard to possible selectivity requirements. In reality, however, when designing a tuned amplifier, the bandwidth requirements are very important and, usually, the following problem must be solved: "design a stable amplifier with given center frequency and bandwidth and obtain as high gain as possible."

Simple considerations show that realizing the maximum gain imposes certain limitations on the bandwidth and, on the other hand, requiring a certain bandwidth imposes certain restrictions on the gain. These limitations belong to two categories:

1) Limitations due to the transistor (device limitations) and
2) Limitations due to external circuit elements.

Some aspects of the gain vs bandwidth relationship can be explained by considering the transistor terminated by synchronous single tuned circuits. Since the $y$ parameters were used in the preceding calculations, it is convenient to use parallel tuned circuits, as shown in Fig. 6. It is assumed that the transistor parameters

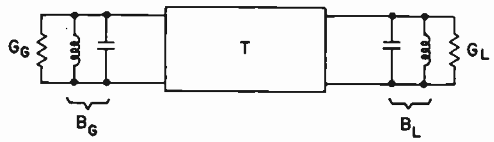


Fig. 6—Transistor with parallel tuned terminations

are frequency independent throughout the frequency band of interest.

The device limitations can be understood *qualitatively* in the following manner. The values of $G_G$ and $G_L$ leading to maximum gain are given by (35) and (36). Maximum gain will be realized only if the values of $B_1$ and $B_2$ are those given by (37) and (38). Considering now, say, the output selectivity one may assume, for example, that the required value of $B_L(=B_2-b_{22})$ is positive. This implies that the load susceptance must be capacitive and since the load consists of $L_L$ and $C_L$ in parallel, the required value of $B_L$ can be realized only if $\omega_0C_L>(B_2-b_{22})$, $\omega_0$ being the center frequency of the amplifier. Consequently, for maximum gain the tuned circuit must contain a capacitance $C_L$ exceeding the value of $(B_2-b_{22})/\omega_0$. However, this and the possible capacitive component of the transistor driving point admittance impose an upper limit on $L_L$. This results in an upper limit of bandwidth realizable with the maximum gain given by (34). If larger bandwidth is required, a smaller capacitance $C_L$ will be used, implying that the denominator of the gain expression in (17) will not be $D_0$ but larger and, therefore, the gain will be reduced. Similarly, if $B_L$ is required to be negative, meaning an inductive load susceptance, there is a maximum admissible value of $L_L$. This again imposes an upper limit on the bandwidth realizable with the calculated maximum gain. Analogous considerations do apply, of course, to the input selectivity.

The bandwidth limitations resulting from device properties are not important in the case of most amplifiers having reasonably narrow bandwidth. Their analytical treatment is complicated and will not be dealt with here.

The limitations imposed by the required bandwidth upon the realizable gain due to external circuit elements are caused by the finite "$Q$" of practical inductances and can be calculated.

In the circuit of Fig. 7, $g_{x1}$ and $g_{x2}$ represent the

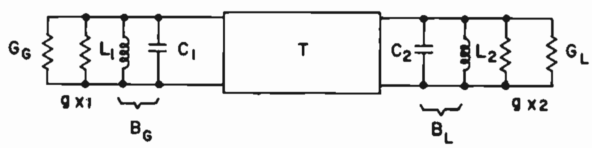


Fig. 7—Transistor with tuned circuits having finite "$Q$."

equivalent parallel loss conductances of inductances $L_1$ and $L_2$. $Q_0$ being the "unloaded $Q$" of the inductances and assuming that $Q_{01}=Q_{02}=Q_0$, one has:

[8] Eqs. (35) and (36) show that $G_G$ and $G_L$ will be positive if the bracketed expressions on the right hand side of these equations are positive, *i.e.*, if the stability factor $k$ of the terminated amplifier exceeds the stability factor $k_0$ of the unloaded amplifier. If the unloaded amplifier is stable in itself, for positive terminations, one must have $k>k$.

$$Q_0 = 1/\omega_0 L_1 g_{x1} = 1/\omega_0 L_2 g_{x2}. \quad (39)$$

The transistor is still treated in terms of its $y$ parameters but the loss conductances $g_{x1}$ and $g_{x2}$ are considered to be parts of $g_{11}$ and $g_{22}$ respectively. One has:

$$g_{11} = g_{11}' + g_{x1} \quad (40)$$

$$g_{22} = g_{22}' + g_{x2} \quad (41)$$

where $g_{11}'$ and $g_{22}'$ are the short circuit input and output conductances of the transistor itself. The transducer gain is:

$$G_T = (4|y_{21}|^2/D_0)[k(L+M)/2G_2 - g_{11}' - g_{x1}](G_2 - g_{22}' - g_{x2}). \quad (42)$$

Assuming that the tuned circuits at the input and the output are synchronously tuned but may have different fractional bandwidths $B^{(1)}$ and $B^{(2)}$ ($B = \Delta\omega/\omega_0$, where $\Delta\omega$ is the 3-db bandwidth):

$$B^{(1)} = (G_G + G_{in})/Q_0 g_{x1} \quad (43)$$

$$B^{(2)} = (G_L + G_{out})/Q_0 g_{x2}. \quad (44)$$

$G_{in}$ and $G_{out}$ are the conductive components of the transistor driving point admittances when the transistor is terminated.

$$G_{out} = g_{22} - (MG_1 + NB_1)/(G_1^2 + B_1^2). \quad (45)$$

Using (37) and defining:

$$W = \frac{M + z_0 N/\sqrt{k(L+M)}}{1 + z_0^2/[k(L+M)]} \quad (46)$$

one finds:

$$G_{out} = g_{22} - [2W/k(L+M)]G_2. \quad (47)$$

Similarly:

$$G_{in} = g_{11} - (MG_2 + NB_2)/(G_2^2 + B_2^2). \quad (48)$$

Using (38), one finds:

$$G_{in} = g_{11} - W/G_2. \quad (49)$$

Eqs. (48) and (49) can be substituted into (43) and (44), yielding:

$$B^{(1)} = [k(L+M)/2G_2 Q_0 g_{x1}][1 - 2W/k(L+M)] \quad (50)$$

$$B^{(2)} = [G_2/Q_0 g_{x2}][1 - 2W/k(L+M)]. \quad (51)$$

Substituting $g_{x1}$ and $g_{x2}$ from (50) and (51) in the gain expression (42), $G_T$ is obtained as a function of $G_2$ alone. Equating $(dG_T/dG_2)$ to zero, the value of $G_2$ leading to maximum gain is found:

$$G_2 = \sqrt{k(L+M)/2}\sqrt{(1 - P/Q_0B^{(1)})/(1 - P/Q_0B^{(2)})} \cdot \sqrt{g_{22}'/g_{11}'}. \quad (52)$$

The corresponding value of $G_1$ is:

$$G_1 = \sqrt{k(L+M)/2}\sqrt{(1 - P/Q_0B^{(2)})/(1 - P/Q_0B^{(1)})} \cdot \sqrt{g_{11}'/g_{22}'}. \quad (53)$$

where:

$$P = 1 - 2W/k(L+M). \quad (54)$$

Substituting $G_2$ in expression (42), the maximum realizable gain is:

$$G_{max,k} = (4|y_{21}|^2/D_0)[\sqrt{(k/2)(L+M)} \cdot \sqrt{(1 - P/Q_0B^{(1)})(1 - P/Q_0B^{(2)})} - \sqrt{g_{11}'g_{22}'}]^2. \quad (55)$$

If $B^{(1)} = B^{(2)} = B$, (55) becomes:

$$G_{max,k} = (4|y_{21}|^2/D_0)[(1 - P/Q_0B)\sqrt{k(L+M)/2} - \sqrt{g_{11}'g_{22}'}]^2. \quad (56)$$

Eqs. (55) and (56) show that the finite $Q$ of external circuit elements reduces the maximum realizable gain at *small bandwidths*, whereas the device limitations have been shown to reduce the maximum realizable gain at large bandwidths.

## Multistage Amplifiers

The method used in the previous sections to determine the maximum realizable transducer gain maintaining a given degree of stability can also be applied to multistage amplifiers. In such amplifiers, interstage coupling is usually performed by "impedance matching"[9] tuned transformers. The problem is to determine the turns ratio $m$ of the transformers and the conductances terminating the amplifier at its input ($G_G$ of the first stage) and its output ($G_L$ of the last stage) terminal pairs. When carrying out such calculations for multistage amplifiers, the following points should be observed:

1) The expression to be maximized is the transducer gain of the amplifier. The transducer gain of the multistage amplifier is, however, *not* the product of the transducer gains of the individual stages, but is the transducer gain of the first stage multiplied by the product of the actual power gains of the remaining stages.
2) The interstage transformer having turns ratio $m$ transforms impedances in both directions. Therefore, since individual stages are terminated by the transformed driving point admittances of adjoining stages, load and source admittances of individual stages cannot be chosen independently. The condition for maximum over-all gain of the amplifier, all stages being operated with a certain de-

[9] The word "matching" is inappropriately used in an amplifier with prescribed stability factor $k$, since, in order to insure stability, the stages are deliberately mismatched.

sired stability factor $k$, does not imply that the individual stages are operated for maximum individual gain. Consequently, the maximum over-all gain of an $n$-stage amplifier, all of whose stages have the stability factor $k$, is less than $n$ times the maximum gain realizable with a single stage having the same degree of stability.

It should be mentioned that, already in the case of two-stage amplifiers (the analysis of which is outlined in Appendix II) the algebraic and numerical evaluation of the equations becomes exceedingly involved. The complete evaluation of such or more complicated cases (for example, that of iterative stages) should be handled with the aid of appropriate computing machinery.

## The Use of Different Matrix Representations

In the preceding sections, the discussion was based on the use of the admittance parameters. Since condition (5) can be stated in analogous form in terms of the "$z$," "$h$," and "$g$" parameters, all calculations could be carried out in terms of these other matrix representations. In fact, the calculations could have been carried out in terms of generalized "$k$" parameters, provided that the generator immittance is given the dimension of $k_{11}$ and load immittance the dimension of $k_{22}$.

However, the different matrix representations have their respective advantages, depending on the nature of the terminating (or interstage coupling) tuned circuits. The following sets of parameters should be used with different types of tuned circuits:

1) $y$ parameters with parallel tuned circuits at input and output (or parallel-parallel interstage networks)—see Fig. 8(a).
2) $z$ parameters with series tuned circuits at input and output (or series-series tuned interstage networks)—see Fig. 8(b).
3) $h$ parameters with series tuned circuit at the input and parallel tuned circuit at the output (or parallel-series tuned interstage networks)—see Fig. 8(c).
4) $g$ parameters with parallel tuned circuit at the input and series tuned circuit at the output (or series-parallel tuned interstage networks)—see Fig. 8(d).

With transistors, cases 1) and 3) have practical importance. Cases 2) and 4) are almost never used because the transistor driving point immittances would lead to inconvenient values of inductances and capacitances.

Some numerical calculations are carried out in Appendix I, using the $h$ parameters.

## Conclusion

It has been shown that using an appropriately defined "stability factor" $k$, the "maximum gain" of a transistor amplifier is a meaningful concept even at frequencies where the transistor exhibits potential instability. The maximum gain corresponding to a given stability factor $k$ has been computed for a one-stage amplifier having tuned generator and load admittances. The gain limitations due to bandwidth requirements have also been discussed for this case.

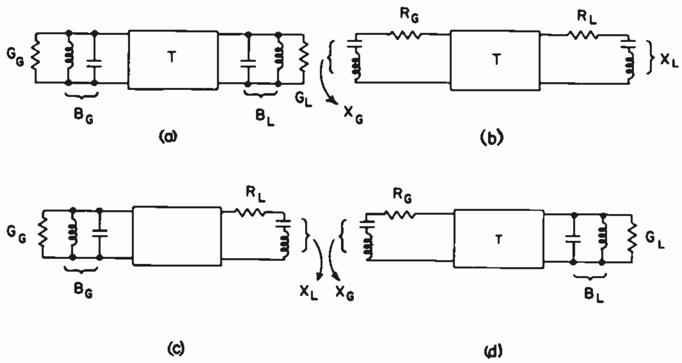


Fig. 8—Various tuned amplifier configurations.

## Appendix I

## Some Numerical Examples

We consider a transistor having the following nominal device parameters:

Low-frequency common-base short circuit current amplification: $a_0 = 0.99$.
Cutoff frequency of the common-base short circuit current amplification: $\omega_a/2\pi = 10$ mc.
Collector capacitance: $C_c = 10$ uuf.
Base spreading resistance: $r_b' = 100$ ohms.
Emitter diffusion resistance (at 1-ma emitter current): $r_e = 25$ ohms.

We are interested in the maximum power gain realizable with this transistor in *common emitter* configuration with various given values of the stability factor $k$. The operating frequency is assumed to be $\omega/2\pi = 0.5$ mc and, consequently: $\rho = \omega/\omega_a = 0.05$.

The circuit used is shown in Fig. 9 and calls for the use of the $h$ parameters. Using the approximate equivalent circuit of Fig. 10 the parameters of interest can be calculated.

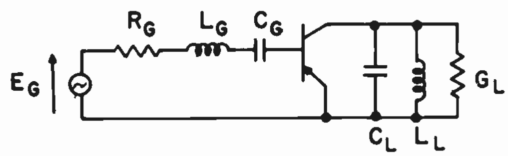


Fig. 9—Common emitter circuit with series tuned input and parallel tuned output.

The real component of $h_{11}$ is:

$$r_{11} \cong r_b' + r_e(1 - a_0)/\rho^2 \cong 200 \text{ ohms.} \qquad (57)$$

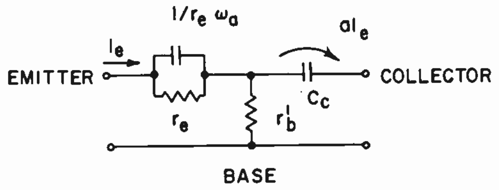


Fig. 10—Approximate equivalent circuit of a junction transistor.

The real component of $h_{22}$ is:

$$g_{22} \cong \omega_a C_c \cong 6 \times 10^{-4} \text{ mhos.} \tag{58}$$

The "internal loop amplification" ($h_{12}h_{21}$) is:

$$\begin{aligned} h_{12}h_{21} &\cong [\omega_a C_c r_e + j\omega_a C_c r_e(1 - a_0)/\rho](1/j\rho) \\ &\cong 0.06 - j\,0.3. \end{aligned} \tag{59}$$

The device is potentially unstable, since

$$[\,|\,h_{12}h_{21}\,| + \text{Re}\,(h_{12}h_{21})]/2 \cong 0.19 > r_{11}g_{22} \cong 0.12. \tag{60}$$

The choice of $k$ depends on the admissible tolerances in device parameters and in the load. For the *nominal* parameters and corresponding terminations we want to have

$$(R_g + r_{11})(G_L + g_{22}) = k[\,|\,h_{12}h_{21}\,| + \text{Re}\,(h_{12}h_{21})]/2. \tag{61}$$

Since Im $(h_{12}h_{21}) \gg$ Re $(h_{12}h_{21})$, the right-hand side of (61) will vary like [see (59)]

$$\text{Im}\,(h_{12}h_{21}) \cong \omega_a^2 C_c r_e/\omega. \tag{62}$$

We may, for example, assume that $\omega_a$, $C_c$, $r_e$ will each have a tolerance (in the upper direction) of 3 per cent. Furthermore, we may assume that the two factors in the left-hand side of (61) will vary by no more than 3 per cent each in the lower direction (these tolerances are too small to be realistic and they are also inexact, since both $r_{11}$ and $g_{22}$ are functions of the device parameters). We would then require $k \cong (1.03)^6 \cong 1.2$. Computing the corresponding maximum gain, we find

$$G_{\max,1.2} \cong 36 \text{ db.}$$

By increasing the tolerances to, say, $+10$ per cent in each factor on the right-hand side of (62) and to $-10$ per cent on the left-hand side of (61), we require $k \cong 1.7$, leading to

$$G_{\max,1.7} \cong 33 \text{ db.}$$

By increasing the tolerances even further, to $+30$ per cent in $\omega_a$ and $r_e$, $+50$ per cent in $C_c$, $-10$ per cent in $(G_G + g_{11})$ and $(G_L + g_{22})$, we require $k \cong 4$. This leads to

$$G_{\max,4} \cong 30 \text{ db.}$$

A further increase in tolerances may require $k = 8$ and leads to

$$G_{\max,8} \cong 28 \text{ db.}$$

The example shows that the tolerances in device parameters ($\omega_a$, $C_c$), emitter current ($r_e$ is principally a function of the emitter current) and terminations ($G_G$ and $G_L$) have considerable influence on the maximum realizable stable gain.

Similar calculations can be carried out for the *common base* stage. Here we have:

$$r_{11} \cong r_e \cong 25 \text{ ohms} \tag{63}$$

$$g_{22} \cong \omega^2 C_c^2 r_b' \cong 10^{-7} \text{ mhos} \tag{64}$$

$$h_{12}h_{21} \cong -j\omega C_c r_b' a_0 \cong 3.10^{-3}. \tag{65}$$

The device is potentially unstable, since:

$$\begin{aligned} [\,|\,h_{12}h_{21}\,| + \text{Re}\,(h_{12}h_{21})]/2 &\cong 15 \times 10^{-4} \\ r_{11}g_{22} &\cong 25 \times 10^{-7}. \end{aligned}$$

Calculating the maximum gain for the same values of $k$ as in the common emitter case, we find:

$$\begin{aligned} k &= 1.2 & G_{\max,1.2} &\cong 44 \text{ db} \\ k &= 1.7 & G_{\max,1.7} &\cong 36 \text{ db} \\ k &= 4.0 & G_{\max,4} &\cong 28 \text{ db} \\ k &= 8.0 & G_{\max,8} &\cong 25 \text{ db.} \end{aligned}$$

Note that in the common base case, these values of $k$ correspond to *different tolerances* in the device parameters than in the common emitter case. With the particular set of device parameters chosen, the common base gain is higher for low (impractical) values of $k$, whereas at higher values of $k$, the common emitter gain is superior.

Note furthermore, that for a different circuit (for example, if parallel tuned circuits are used at both input and output, where the $y$ parameters should be employed instead of the $h$ parameters used in the above example), different values of $k$ will correspond to the same permissible tolerances in the device parameters and different values of maximum gain may be realized.

## Appendix II

### Outline of the Analysis of a Two-Stage Amplifier

The schematic circuit of an amplifier consisting of two stages $T_1$ and $T_2$ is shown in Fig. 11.

We designate by:

- $P_G$ = the power available from source $Y_G$.
- $P_L$ = the power delivered to load $Y_L$.
- $P_{i2}$ = the power delivered by $T_1$ to $T_2$.
- $G_{T,\text{total}}$ = the transducer gain of the two-stage amplifier.
- $G_{T1}$ = the transducer gain of $T_1$.
- $G_{P2}$ = the actual power gain of $T_2$ (= power delivered to $Y_L$ divided by input power of $T_2$).

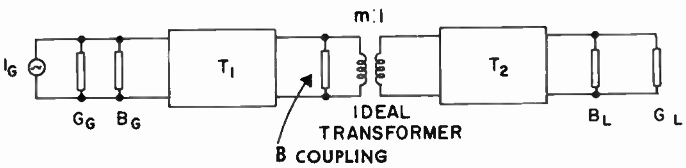


Fig. 11—Two stage amplifier with interstage impedance transformation (in practice the ideal transformer would, of course, be part of the coupling susceptance).

Then:

$$G_{T.\text{total}} = \frac{P_L}{P_G} = \frac{P_{i2}}{P_G} \cdot \frac{P_L}{P_{i2}} = G_{T1} \cdot G_{P2}. \tag{66}$$

Defining $G_1$ and $G_2$ in the sense used throughout the body of this paper and indicating by a second subscript whether the quantity considered belongs to $T_1$ or $T_2$, one has according to (10) and (30):

$$G_{T1} = \frac{4 | y_{21} |^2}{D_0} (G_{11} - g_{11}) \left[ \frac{k(L + M)}{2G_{11}} - g_{22} \right]. \tag{67}$$

The power gain of $T_2$ is:

$$G_{P2} = \frac{| y_{21} |^2 (G_{22} - g_{22})}{g_{11}[G_{22}^2 + B_{22}^2] - MG_{22} - NB_{22}}. \tag{68}$$

For any given value of $G_{22}$, $G_{P2}$ will be maximum if

$$B_{22} = N/2g_{11}. \tag{69}$$

Substituting (69) into (68), $G_{P2}$ can then be written:

$$G_{P2} = \frac{4g_{11} | y_{21} |^2 (G_{22} - g_{22})}{(2g_{11}G_{22} - M)^2 - L^2}. \tag{70}$$

The transducer gain of the complete amplifier is:

$$G_{T.\text{total}} = \frac{16g_{11} | y_{21} |^4}{D_0} (G_{11} - g_{11}) \left[ \frac{k(L + M)}{2G_{11}} - g_{22} \right] \cdot \frac{G_{22} - g_{22}}{(2g_{11}G_{22} - M)^2 - L^2}. \tag{71}$$

The load admittance of $T_1$ is $(1/m^2)$ times the input admittance of $T_2$ and the generator admittance of $T_2$ is $m^2$ times the output admittance of $T_1$. Furthermore, the same stability factor is required for both stages and (10) applies to both stages. In view of (69) and (46) one can write:

$$G_{21} = \frac{k(L + M)}{2G_{11}} = g_{22} + \frac{1}{m^2} \left( g_{11} - \frac{MG_{22} + N^2/2g_{11}}{G_{22}^2 + N^2/4g_{11}^2} \right) \tag{72}$$

and

$$G_{12} = \frac{k(L + M)}{2G_{22}} = g_{11} + m^2 \left( g_{22} - \frac{W}{G_{11}} \right). \tag{73}$$

These two equations establish relationships between $G_L(=G_{22}-g_{22})$, $G_G(=G_{11}-g_{11})$ and $m$. $m$ can be eliminated by multiplying (72) by (73). One finds:

$$\left[ \frac{k(L + M)}{2G_{22}} - g_{11} \right] \left[ \frac{k(L + M)}{2G_{11}} - g_{22} \right] = \left( g_{22} - \frac{W}{G_{11}} \right) \left( g_{11} - \frac{MG_{22} + N^2/2g_{11}}{G_{22}^2 + N^2/4g_{11}^2} \right). \tag{74}$$

Eq. (74) relates $G_{11}$ to $G_{22}$ (or $G_L$ to $G_G$). One of these quantities can be expressed as a function of the other and substituted into the gain expression (71). Then, in principle, the maximum over-all transducer gain can be determined. It is, however, obvious that this procedure involves computational difficulties due to the complicated nature of the resulting gain expression.

## Acknowledgment

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# Ferrod Radiator System*

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*Summary*—**Commercially available ferrites with low dielectric and magnetic losses at microwave frequencies have brought about the development of ferrod radiator systems for 3-cm wavelength. These ferrites are, in general, characterized by high dielectric constants (approximately 13) and initial permeabilities slightly less than unity. The use of these ferrimagnetic materials as ferrite rod (ferrod) radiators considerably reduces the physical size of microwave antenna systems.**

**The high dielectric constant of the ferrite allows the diameter of the radiating elements of circular cross section to be approximately $\frac{1}{4}$ inch for 3-cm wavelength use. This makes possible magnetic coupling of the ferrod radiator from resonant cavities or from the narrow side of standard rectangular waveguides to obtain various types of linear arrays. A short feeding section ($\frac{5}{16}$ inch long) is used to support and lock the position of the ferrod. This device simplifies adjustments and reduces the mechanical tolerances necessary for the ferrod radiators.**

**Microwave ferrites are also characterized by a tensor permeability[1] which changes with applied magnetic field. This phenomenon is used to change the magnetic properties of the ferrod radiator to obtain electrical switching, scanning, lobing, changes in polarization, and amplitude or phase modulation of the radiated energy.**

**Phase shifters, switches, and circulators, which consist of short sections of ferrite-filled cylindrical waveguide, have also been developed for use with the ferrod radiator system.**

## Introduction

THE BASIC IDEAS of radiating short wavelength radio waves from dielectric media date back to the work done by Lord Rayleigh[2] in 1895. These ideas consist of shaping a length of dielectric material to control the phase and amplitude of the radiated energy.

Essentially, this is what was done to the ferrods which were used as radiating elements of 3-cm wavelength antenna systems developed by this laboratory.

An exact analysis of the operation of these ferrod radiators involves the solution of Maxwell's equations, subject to all the boundary conditions and parameters describing the ferrod. Its mathematical complexity is beyond the scope of this paper. For most engineering applications, however, a satisfactory explanation of its operation can be obtained by establishing analogies with array theory and with existing theories of transmission through uniform dielectric rods of circular cross section.

A good treatment of the theory of electromagnetic wave propagation along dielectric rods is given by Kiely[3] and McKinney.[4] These two surveys have been of particular help in determining the parameters necessary for obtaining good radiation characteristics. They also contain complete bibliographies of the literature available on dielectric-rod radiators.

Treatments of electromagnetic wave propagation in magnetically anisotropic media are given by Kales[5] and Gintsburg.[6] A similar treatment with the addition of graphical solutions of the transcendental equations involved in certain cases is given by Suhl and Walker.[7] Wave propagation in an infinite, magnetized ferrod is also being studied at this laboratory.

## Ferrod Radiator System

A basic ferrod antenna system for 3-cm wavelength is shown in Fig. 1. It consists of a ferrite radiating element, a low current modulating coil, a short section of cylindrical waveguide for supporting the ferrod, a simple wire mode filter, and a short section of rectangular waveguide used to excite the ferrod radiator. The length of the entire system is approximately 5 inches.

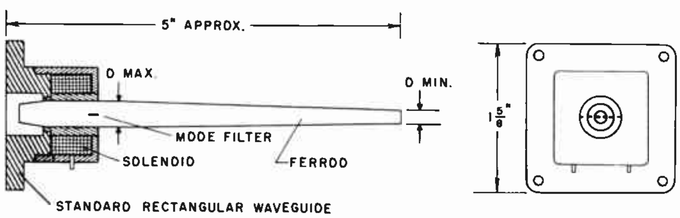


Fig. 1—Ferrod radiator system for 3-cm wavelength.

The radiating element used is a commercially available low-loss ferrite[8] which is properly shaped to give maximum radiation along its axis and minimum radiated power in the side lobes. The total length of the ferrod is 4.5 inches, and its diameter is 0.25 inch.

The ferrite-filled section of waveguide has an inside diameter of $\frac{1}{4}$ inch, and is approximately $\frac{3}{4}$ inch long.

* Original manuscript received by the IRE, December 7, 1956.

† Diamond Ordnance Fuze Labs., Washington, D. C.

[1] E. G. Spencer, R. C. LeCraw, and F. Reggia, "Measurement of microwave dielectric constants and tensor permeabilities of ferrite spheres," IRE Proc., vol. 44, pp. 790–800; June, 1956.

[2] Lord Rayleigh, "On the passage of electrical waves through tubes or the vibrations of dielectric cylinder," *Phil. Mag.*, vol. 43, pp. 125–132; February, 1897.

[3] D. G. Kiely, "Dielectric Aerials (Methuen's Monograph)," John Wiley and Sons, Inc., New York, N. Y.; 1953.

[4] C. M. McKinney, Jr., "Dielectric Waveguides and Radiators," Bumblebee Rep. No. 138, Defense Res. Lab., Univ. of Texas, Austin, Texas; 1950.

[5] M. L. Kales, "Modes in waveguides containing ferrites," *J. Appl. Phys.*, vol. 24, pp. 604–608; May, 1953.

[6] M. A. Gintsburg, "On waves in a gyrotropic medium," *C. R. Acad. Sci. (URSS)*, vol. 18; 1954. For English translation see Columbia Technical Translations No. 4, vol. 18, pp. 136–154; 1950.

[7] H. Suhl and L. R. Walker, *Bell Sys. Tech. J.*, vol. 33, pp. 579, 939, and 1133; 1954.

[8] General Ceramics R-1 Ferrite. A complete description of its microwave properties is given by Spencer, LeCraw, and Reggia, *op. cit.* See also R. C. LeCraw and E. G. Spencer, "Tensor permeabilities of ferrites below magnetic saturation," 1956 IRE Convention Record, Part 5, pp. 66–74.

The mode filter, 0.020 inch thick and approximately 0.060 inch wide, goes through the ferrite and makes electrical contact to both sides of the circular waveguide. The low current solenoid,[9] 9/16 inch wide and $1\frac{3}{8}$ inch in diameter, required about 10 ma to operate. These three components make up the "ferrite mode switch" or modulating section of the system.

Operation of the mode switch depends upon the ferromagnetic Faraday effect[10] of the ferrite which causes a rotation of the plane of polarization of the microwave energy incident upon the mode filter when a longitudinal magnetic field is applied. When the polarization of the incident wave is perpendicular to the length of the mode filter, the wave propagates essentially unchanged. Rotating the polarization parallel to the length of the mode filter short-circuits the electric field, thus rejecting the microwave energy incident upon it. Greater than 20 db switching is obtained with this modulation section with control currents of 10 ma ($H_0 \approx 40$ oersteds).

Modulating frequencies in the kilocycle range have been used with the system shown in Fig. 1. It is possible to extend this frequency range up to a few megacycles by techniques reported by R. C. LeCraw[11] of this laboratory.

## Design Data

Kiely[3] has shown the radiation pattern of a uniform dielectric rod of circular cross section excited in the $HE_{11}$ mode to be given approximately by

$$E_\theta = \frac{(K-1)\sin\frac{\pi L}{\lambda_0}(K-\cos\theta)}{(K-\cos\theta)\sin\frac{\pi L}{\lambda_0}(K-1)} \cos\left(\frac{\pi d}{\lambda_0}\sin\theta\right)$$

where

$E_\theta$ = normalized electric field intensity
$\theta$ = angle between the axis of the dielectric rod and the position where the electric field is being determined
$\lambda_0$ = free space wavelength
$\lambda$ = wavelength in an infinite medium of the dielectric
$L$ = length of the dielectric rod
$d$ = diameter of the dielectric rod.

The value of $K = \lambda_0/\lambda$ which gave optimum radiation characteristics was determined analytically to be 1.1, assuming no standing waves exist along the rod. To make this assumption valid, it was necessary to taper the rod to match the free space impedance. The calculated value of $d_{max}$ for the tapered rod (see Fig. 1) was found to be $\lambda_0/\sqrt{\pi(\epsilon-1)}$, and the value of $d_{min}$ was found to be 0.63 $d_{max}$. The magnetic permeability $\mu$ was assumed to be unity in this analysis.

The cutoff frequency ($f_c$) of a dielectric-filled waveguide of circular cross section operating in the $TE_{11}$ mode is given by $f = 0.293\ c/a\sqrt{\mu\epsilon}$ where

$c$ = velocity of light
$a$ = radius of circular waveguide.

Using a ferrite with a relative dielectric constant $\epsilon$ of 13.6 and a permeability $\mu$ of 0.76 as the dielectric medium requires that a diameter no smaller than 0.240 inch be used for $d_{max}$ in the 9–10 kmc frequency range. This is slightly larger than the diameter obtained from Kiely's equation. Using this value for $d_{max}$, the value of $d_{min}$ was found to be 0.152 inch. Other parameters of the ferrod radiator were determined experimentally, such as the length of the taper which is a function of the ferrod length and the length of the feeding section.

A comparison between the calculated and measured values of the radiation pattern resulted in only approximate agreement. The largest discrepancies occurred in the amplitude and position of the side lobes. This is not surprising since existing theories are themselves only approximate. Also, because of the high dielectric constant of the ferrod, it is possible that some other mode may be excited.

The small diameter required for the ferrod radiator at 3-cm wavelength makes possible direct coupling from the narrow side of a standard rectangular waveguide or from a small resonant cavity. This simplifies the construction of various types of arrays for obtaining radiation patterns with more narrow beam widths.

A single ferrod radiator, properly shaped for optimum radiation characteristics, is shown in Fig. 2. It is coupled from the end of a rectangular waveguide (1.0×0.5 inch) by a small locking device, also shown in the foreground.

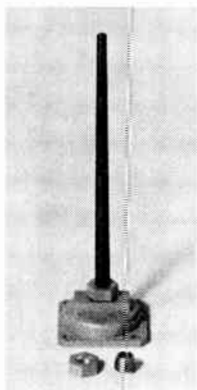

Fig. 2—Ferrod radiator with optimum taper.

This short feeding section (0.32 inch long) is designed to support the ferrod radiator and to lock its position after adjustment. It also reduces the machining tolerances necessary for the ferrod and minimizes the time necessary to make final adjustments in the various types of arrays. Impedance matching is accomplished

[9] Advance Electric and Relay Co., Solenoid SEIC-1600 D. The magnetic field is given by H (oersteds) = 3.9 I(ma).
[10] A. G. Fox, S. E. Miller, and M. T. Weiss, "Behavior and applications of ferrites in the microwave region," *Bell Sys. Tech. J.*, vol. 34, pp. 5–103; January, 1955.
[11] R. C. LeCraw, "High speed magnetic pulsing of ferrites," *J. Appl. Phys.*, vol. 25, pp. 678–679; May, 1954.

by adjusting the length of the ferrod inside the waveguide, this adjustment being made less critical by a short taper at the input end of the rod. Input vswr's less than 1.2, over a 5 per cent frequency band, and half-power beamwidths of less than 28° were not difficult to obtain with this single radiator.

A conservative power rating for the ferrod antenna is 20 watts of average power per rod. The maximum rating depends upon the rf losses in the ferrite and its environment. With presently available ferrites, it is believed that this power handling capability can be extended to 50 w. When only the dielectric property of a ferrite, which is insensitive to temperature changes, is made use of in the antenna array, it is expected that an average power of 100 w per rod would have little effect on the performance of the antenna.

## Linear Ferrod Arrays

Several types of linear arrays with ferrods coupled from the narrow side of rectangular waveguide have been constructed. Two of these antennas are shown in Fig. 3. This type of end-fed array is easy to construct and its length can be extended to give radiation patterns with more narrow beams. The individual elements are spaced $\lambda_g$ apart, resulting in maximum radiation along the axis of the ferrods. The radiation pattern of this broad-side antenna is a narrow fan-shaped beam. Small matched loads or adjustable shorts are used to terminate the section of the waveguide making up the array, depending upon its length and application.

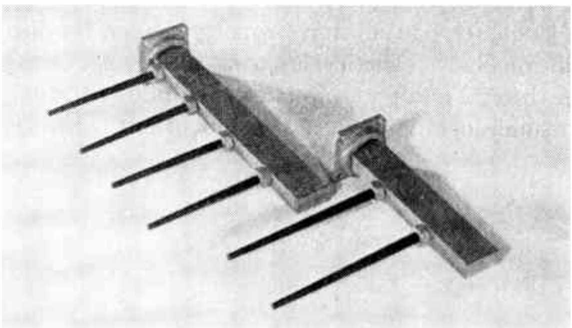

Fig. 3—Two linear ferrod arrays showing one method of coupling.

Another similar type of linear array, but center-fed, is shown in Fig. 4. The microwave energy is coupled into the waveguide from the back of the array through a small ferrite rotating-type switch. Thirty-db switching and a half-power beamwidth of 7° were obtained with this antenna system. Short matched loads are used at each end of the waveguide section to minimize internal reflections.

The positions of the ferrods with respect to the field distribution in several linear arrays are shown in Fig. 5. The ferrods are placed at positions of maximum magnetic field and minimum electric field. The positions of electric field are indicated by crosses and dots at the center of the magnetic loops ($H_{min}$). Fig. 5(a) shows the field distribution inside a $TE_{207}$ rectangular cavity-fed array. The ferrod radiators can be coupled from either the top (as shown) or the side of this low-$Q$ cavity.

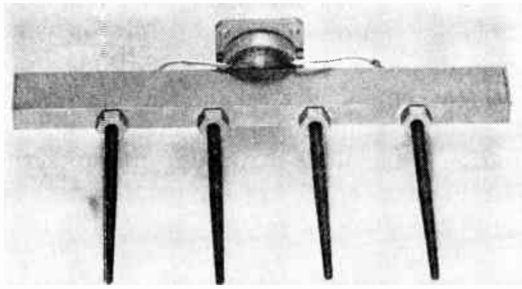

Fig. 4—Linear ferrod array with ferrite switch (or modulator) at input.

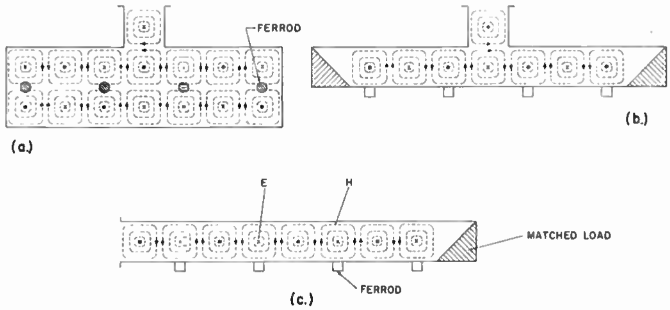


Fig. 5—Linear arrays showing magnetic coupling of ferrod radiators.

Stacking one linear array above another narrows the beamwidth of the radiated energy in the vertical plane and thus increases the gain of the system. Several arrangements that were used are shown in Fig. 6. Phase shifters may be used between the individual arrays to insure in-phase excitation of all the radiating elements, and matched loads are used where necessary.

Consideration has also been given to the construction of long linear arrays for obtaining very narrow fan-shaped beams. One of these arrays, shown in Fig. 7, has a beamwidth of approximately 4°, and its side lobe amplitudes are down at least 18 db from that of the main beam. Its measured gain is approximately 140. This center fed array has matched loads at both ends and it is approximately 22 inches long. The work on these long arrays is being continued.

## Cavity-Fed Arrays

Another type of antenna array has been designed with ferrod radiators magnetically coupled from resonant cavities. These small cavity arrays are easy to construct and in general give higher gains than the linear arrays previously described.

The field distribution and position of the ferrods inside a $TE_{304}$ mode cavity, are shown in Fig. 8. The magnitudes, directions, and positions of the electric and

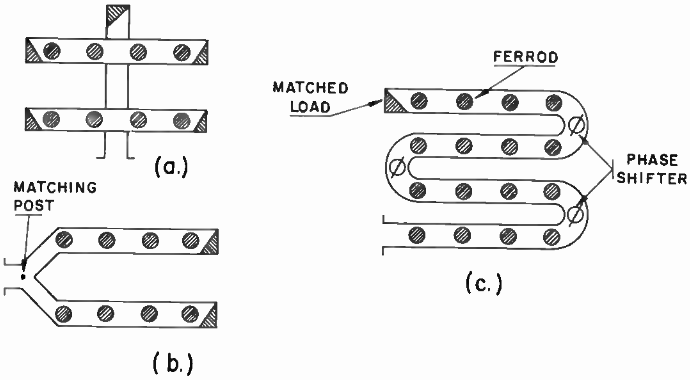


Fig. 6—Stacking of linear arrays.

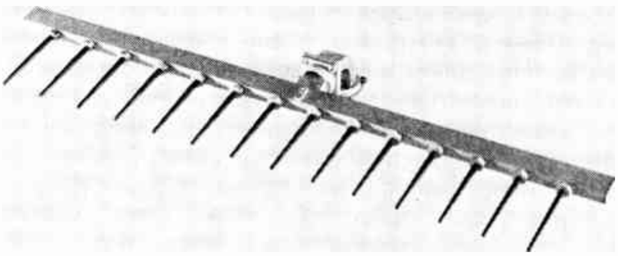

Fig. 7—Long linear ferrod array.

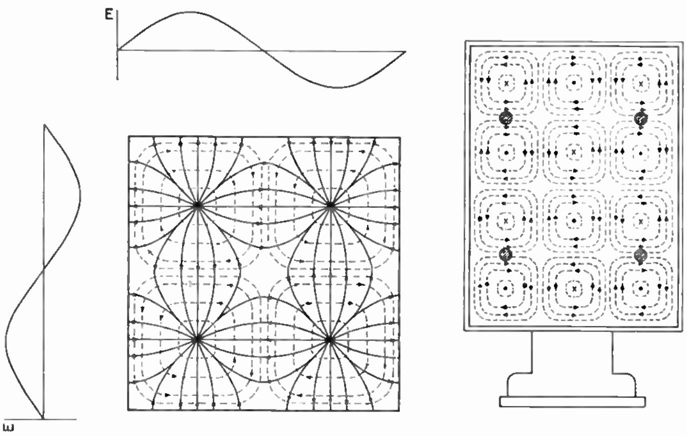


Fig. 8—Field distribution inside a $TE_{304}$ mode cavity feeding a 4-ferrod square array.

magnetic fields can be seen in the drawing at the right, as well as the positions of the ferrods (shaded circles) at points of maximum magnetic field. Positions of maximum electric field are indicated by the crosses and dots at the center of the magnetic loops.

An expanded view of a section of this rectangular cavity, also showing current lines, can be seen at the left of the figure. The position of zero electric and zero magnetic field can be seen to exist at the exact center of this diagram. Shorting posts, 1/16-inch diameter, are placed at these positions to insure the proper mode of oscillation of the cavity (suppress lower order modes) and to insure in-phase excitation of the ferrod radiators.

A photograph of this cavity and a smaller one which is resonant in the $TE_{203}$ mode are shown in Fig. 9. The positions of the 6 shorting posts of the $TE_{304}$ cavity and the 2 used in the $TE_{203}$ mode cavity can be seen in this figure. The dimensions of the $TE_{203}$ cavity are $1.75 \times 2.75 \times 0.50$ inches, and the $TE_{304}$ mode cavity, $3.75 \times 2.75 \times 0.50$ inches.

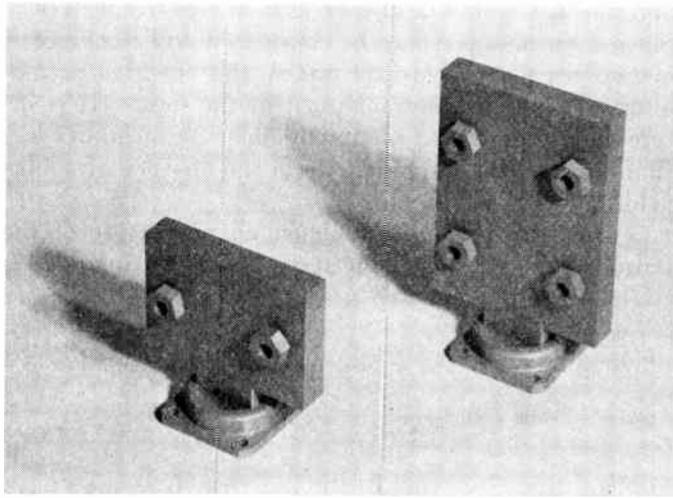

Fig. 9—Linear and square cavity feed arrays.

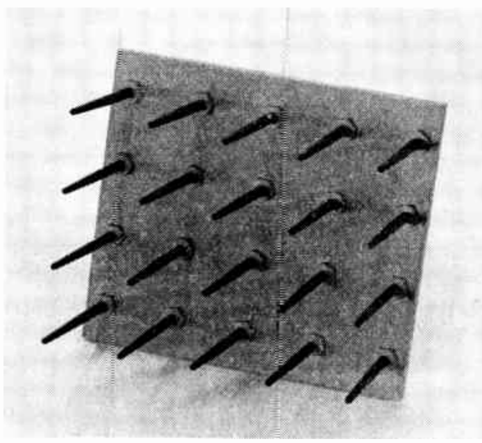

Fig. 10—$TEM_{809}$ mode cavity-feed array.

This cavity technique has been extended to obtain even sharper radiating beams and higher gains. An example of this is shown in Fig. 10. The waveguide feed enters the back at the center of the cavity. The shorting pins (or mode filters) at positions of zero electric and magnetic fields can be seen as small circles between the ferrods. The dimensions of this $TE_{809}$ mode cavity-feed array are $7.0 \times 7.75 \times 0.5$ inches and its measured beamwidth is approximately $6 \times 7°$.

### Ferrite-Filled Cylindrical Waveguide Devices

In order to utilize the small size of the ferrod antenna system, compact microwave switches and circulators were designed for array switching and beam lobing.

These devices use a longitudinal magnetic field applied to a ferrite-filled section of cylindrical waveguide, similar to those shown in Figs. 1 and 4.

Operation of these devices depends upon the variable magnetic properties of the ferrite as shown in Fig. 11. These curves[8] show the magnitude of the dispersive components of the scalar permeability ($\mu' \pm K'$) of the R-1 ferrite below magnetic saturation for the positive and negative circularly-polarized waves respectively. Since a linear wave may be considered to be composed of two circularly-polarized waves, the effective permeability ($\mu'_{eff}$) has been included in the figure. It is interesting to note that the ferrite has a permeability less than unity at these frequencies in the absence of a bias field.

A small microwave switch which makes use of the tensor characteristic of the ferrite is shown in Fig. 12.

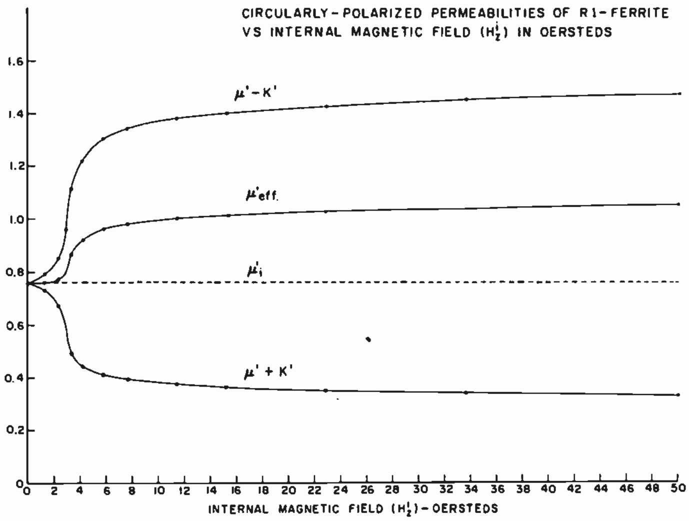


Fig. 11—Real components of the circularly-polarized permeabilitie of R-1 ferrite below magnetic saturation.

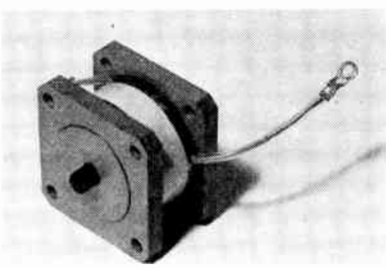

Fig. 12—Ferrite-filled cylindrical waveguide switch.

It consists of a $\frac{1}{4}$-inch diameter ferrite-filled cylindrical waveguide section inside a low-field solenoid. The ferrite protrudes approximately $\frac{1}{8}$ inch into the rectangular waveguide which is used with the switch. When no current flows through the solenoid, the switch is ON, and the microwave energy propagates through the device with essentially no change. With approximately 10 ma of solenoid current, the microwave energy is rotated 90° by the magnetized ferrite and is rejected by the cross-polarized rectangular waveguide at the output. This switch has an insertion loss of about 0.6 db when the switch is ON, and an isolation of greater than 30 db when a magnetic field is applied.

By using a slightly smaller ferrite-filled cylindrical waveguide section, this same switch can be made to operate as a waveguide-below-cutoff switch.[12] When current flows in the solenoid, the effective permeability of the ferrite is increased sufficiently to allow the ferrite-filled section of waveguide to be near, but above cutoff. With no current in the solenoid, the effective permeability is decreased and the waveguide is operating below cutoff. Greater than 50 db switching is obtained with this device when used in this manner.

A rotating flange is used at one end of this switch to allow the position of the output waveguide to be adjusted. This permits higher isolations in the OFF position and lower insertion losses in the ON position when the device is used as a rotation type switch. The switch is $1\frac{1}{8}$ inches long and requires less than 20 ma of current in the solenoid (approx. 80 oersteds) for operation at 3-cm wavelength.

Another ferrite device making use of the Faraday effect is shown in Fig. 13. Its construction is essentially

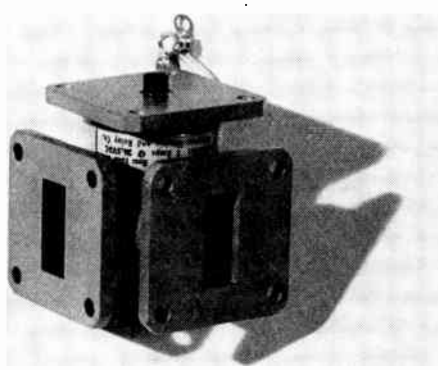

Fig. 13—Three-port microwave switch with ferrite-filled cylindrical waveguide input.

the switch shown in Fig. 12, inserted into a right-angle waveguide bend. The microwave energy enters the ferrite-filled section of waveguide where its plane of polarization is acted upon by an external magnetic field. By rotating the plane of polarization through 90°, energy propagating through one arm can be switched to the other. Greater than 20 db switching has been obtained with this circulator with less than 20 ma of solenoid current. Several of these units in tandem could be used to perform more complex functions in other antenna systems.

[12] F. Reggia, "Magnetically-controlled Microwave Switch with High Isolation," Diamond Ordnance Fuze Labs., TM 41.04-16, unpublished report; May, 1955.

### Array Switching and Lobing

The small ferrite switch of Fig. 1 can be used to control the power transmitted from each radiating element of a ferrod antenna system. The switches shown in Figs. 4 and 12 can be used to control (or modulate) the power transmitted by all radiating elements simultaneously.

The 3-port microwave switch of Fig. 13 has been used effectively in beam lobing systems such as that shown in Fig. 14. The two linear ferrod arrays, rotated azimuthally from each other, are alternately excited by the ferrite switch. Any angle of rotation between the radiating beams can be designed into this antenna system, depending upon its use. The advantage of this system over others tried is the requirement of only one control solenoid.

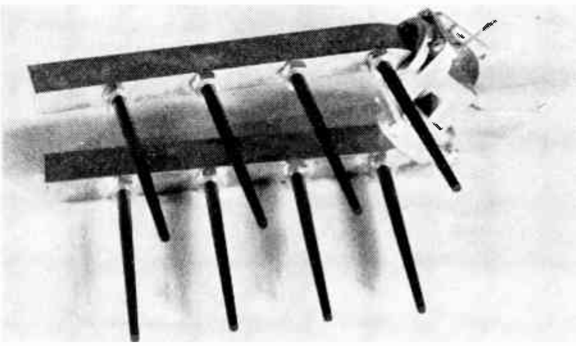

Fig. 14—Beam lobing ferrod antenna system incorporating three-port microwave switch.

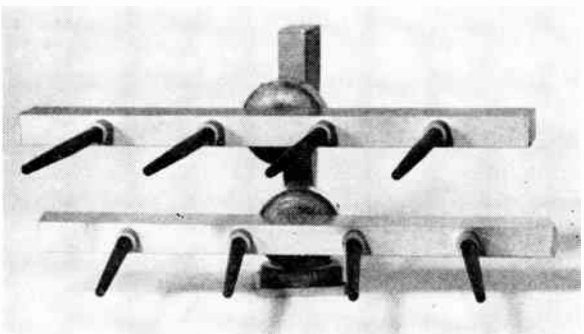

Fig. 15—Beam lobing ferrod array, showing magnetic lobing switches.

Placing one linear array above another on a section of twisted waveguide gives another type of beam lobing system as shown in Fig. 15. A small rotation type ferrite switch in each array provides the necessary beam switching from one array to the other. Any number of these linear arrays can be stacked to obtain greater azimuthal range. When using this antenna system in direction-finding radars, out-of-phase square waves are applied to the two solenoids. Lobing rates of a few kilocycles have been used with the solenoids shown.

### Array Scanning

Several methods for obtaining electrical beam scanning have been tried. One method is to apply a longitudinal magnetic field directly to the feed section of a ferrod radiator. Phase shifts up to 50° have been obtained by this technique with little change in the radiated power or plane of polarization. Another method is to apply a transverse magnetic field to a ferrite slab placed in the position of a circularly-polarized rf magnetic field inside the rectangular waveguide section making up the array as shown in Fig. 16. A phase shift of 90° is not difficult to obtain with this technique. Adjustable ferrite-loaded cavities along the narrow side of the waveguide and inside larger cavity-fed arrays have also been used to obtain a scanning angle up to 10°. A phase shift of 90° between the individual ferrod radiators, spaced $\lambda_g$ (approximately 3/2 $\lambda_0$) apart, is necessary for a 10° shift of the radiated beam.

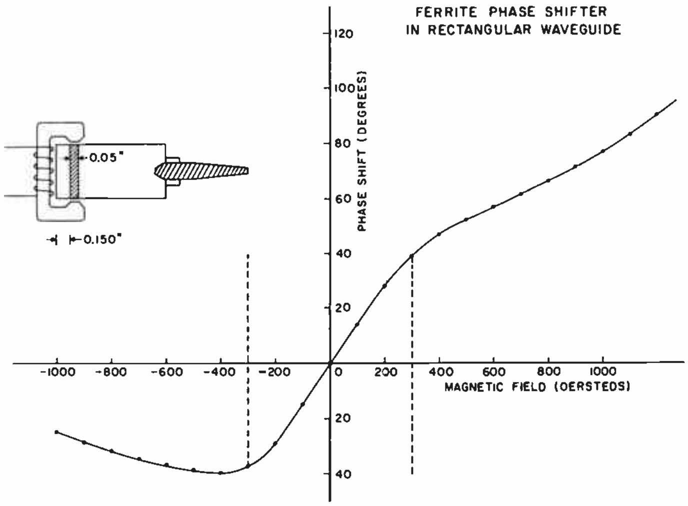


Fig. 16—Nonreciprocal phase shifter in rectangular waveguide.

Circular scanning can be obtained by combining a horizontal and a vertical linear scanning array. By applying an ac voltage to the horizontal scan winding 90° out of time phase with the voltage on the windings for the vertical scan, a nearly circular scan will result. This ac source could also be modulated to produce a spiral scan.

### Rotation of Plane of Polarization

Radiation, first in one plane of polarization and then in another, can be had simply by applying a longitudinal magnetic field to the individual ferrod radiators. An example of this could be the antenna system shown in Fig. 1, without the wire mode filter. This change of polarization by a magnetized ferrod radiator vs the dc control current is shown in Fig. 17. A rotation of 180° is obtained with less than 0.2 db change in the power transmitted with 12 ma of solenoid current ($H_0 \approx 47$ oersteds).

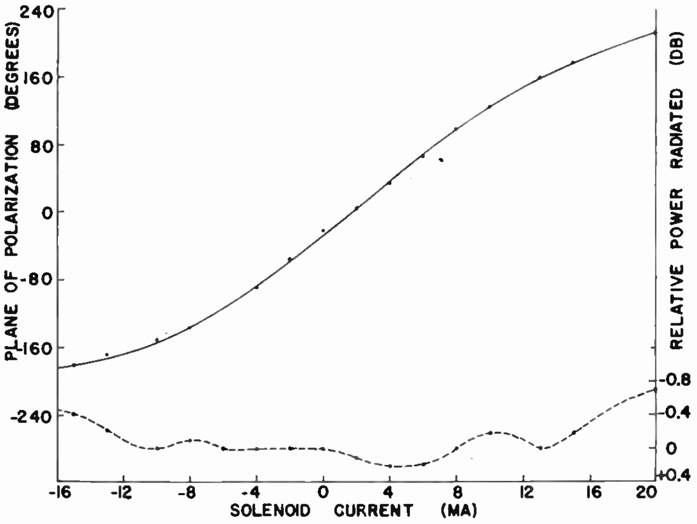


Fig. 17—Rotation of the plane of polarization of radiated energy from ferrod radiator at 9430 mc.

## Circularly-Polarized Radiators

A circularly-polarized radiation field can be obtained by exciting the ferrod from the broad side of a properly terminated waveguide. This position of circularly-polarized field, where the $x$ and $z$ components of the rf magnetic field are in time quadrature and of equal magnitude, is located along a plane approximately one-fourth the distance across the waveguide. Either sense of rotation can be had by choosing the correct side of the waveguide.

## Radiation Patterns

The radiation characteristics of single and multi-element ferrod radiators of different lengths, diameters, and shapes have been obtained in the frequency range between 9000 and 10,000 mc. The single ferrod radiators were excited from the end of rectangular waveguides, as seen in Fig. 2, and the multi-element arrays were coupled from the narrow side of rectangular waveguides or from resonant cavities as previously described.

A simplified circuit arrangement used for the radiaton pattern measurements is shown in the block diagram of Fig. 18. A carefully selected waveguide rotating-joint, calibrated at the operating frequencies, was used to rotate the ferrod antennas in the azimuthal plane.

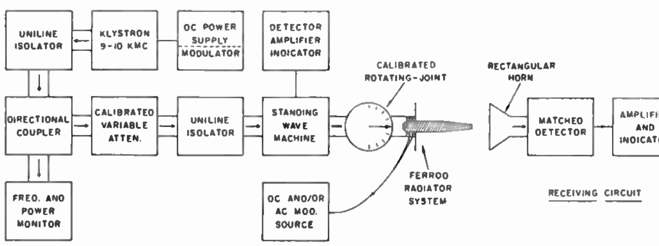


Fig. 18—Simplified circuit arrangement for radiation pattern measurements.

Another rotating mechanical device was used in the receiving circuit to rotate both the rectangular horn and the matched detector for determining the plane of polarization of the radiated energy. A more accurate antenna measuring system was also used over a much longer path, free from all obstructions. In the latter system, the entire transmitting circuit was rotated and an automatic recorder was used in the receiving circuit.

The directivity of a single untapered ferrod radiator (0.240-inch diameter) fed from the end of a rectangular waveguide vs its length, at 9500 mc, is shown at the left in Fig. 19. The minor lobes have been left out to simplify the drawing. The decreasing width of the main beam with increasing length of the ferrod can be seen in this figure. The gain of a single untapered ferrod radiator 4.5 inches long was found to be approximately 20 at this frequency.

A comparison between the radiation patterns of a tapered and an untapered ferrod radiator, 3.5 inches long, is shown at the right of this figure. Although the untapered ferrod gives a more narrow beamwidth, the power radiated in its side lobes is excessive. This side lobe radiation is almost completely eliminated by the tapered ferrod but not without an increase in the width of the main beam. By increasing the length of the tapered ferrod, the original value of the beamwidth can be restored.

The radiation characteristics of single ferrod radiators 5.5 inches long vs the length of tapers at 9300 mc (at right) and 9600 mc (at left) are shown in Fig. 20. Their diameters were 0.240 inch at the input end and 0.152 inch at the output end.

The length of taper giving the best results was approximately 4 inches. The beamwidth can be seen to increase with increasing length of taper and with decreasing frequency. The gain of the ferrod with the 4-inch taper was compared to that of a standard horn and found to be approximately 35. Impedance matching resulting in vswr's less than 1.2 is not difficult to achieve with these ferrods when coupled from the end of a waveguide section.

The radiation pattern of another single ferrod 5.4 inches long, but with a slightly larger diameter (0.270 inch), over the frequency range of $9400 \pm 200$ mc is shown in Fig. 21. The length of taper which gave the best results was found to be 5 inches. The input impedance for this diameter ferrod, when excited from the end of a waveguide section, remained relatively constant over the 400-mc bandwidth. A beamwidth of 30° and a measured gain of approximately 35 were obtained with this radiator.

The radiation characteristic in the azimuthal plane of a $4 \times 5$ ferrod array excited from the $TE_{809}$ resonant cavity at 9500 mc is shown in Fig. 22. A beamwidth of approximately 7° and a gain of approximately 130 were obtained for this array which had a somewhat square-shaped beam. A photograph of this antenna is shown in Fig. 10.

The radiation patterns, in the plane of the ferrods,

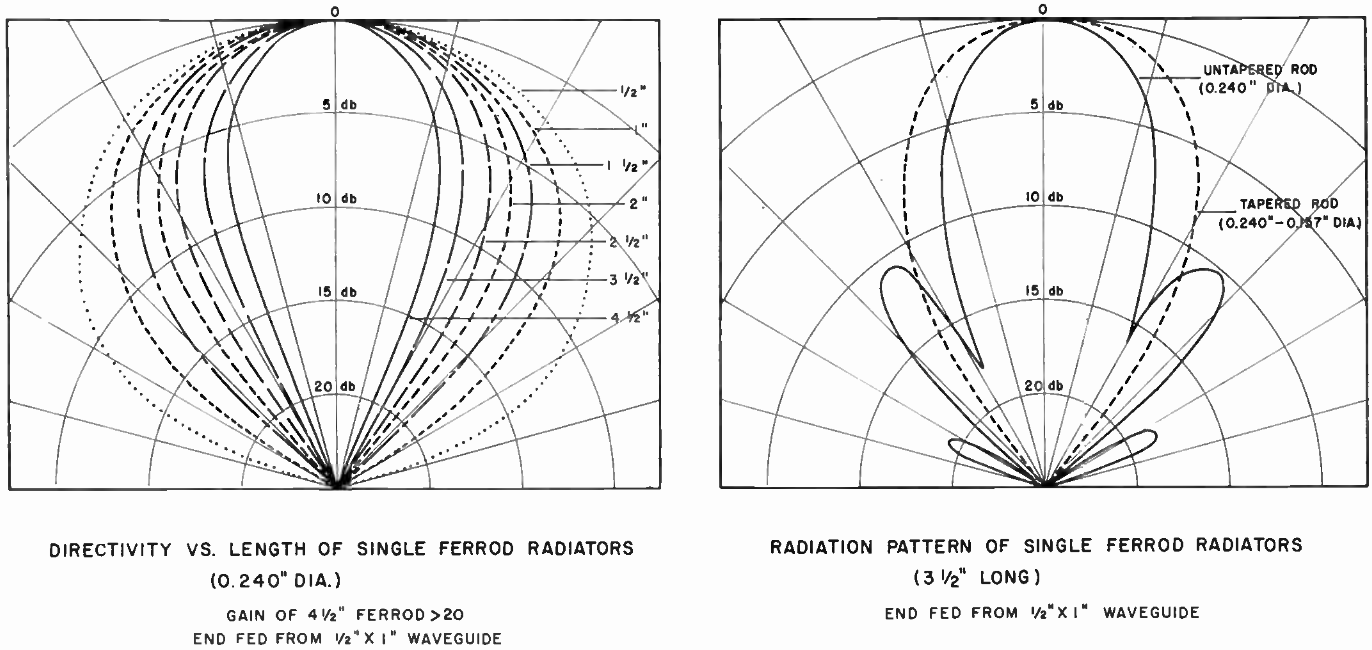


Fig. 19—Radiation patterns of single ferrod radiators.

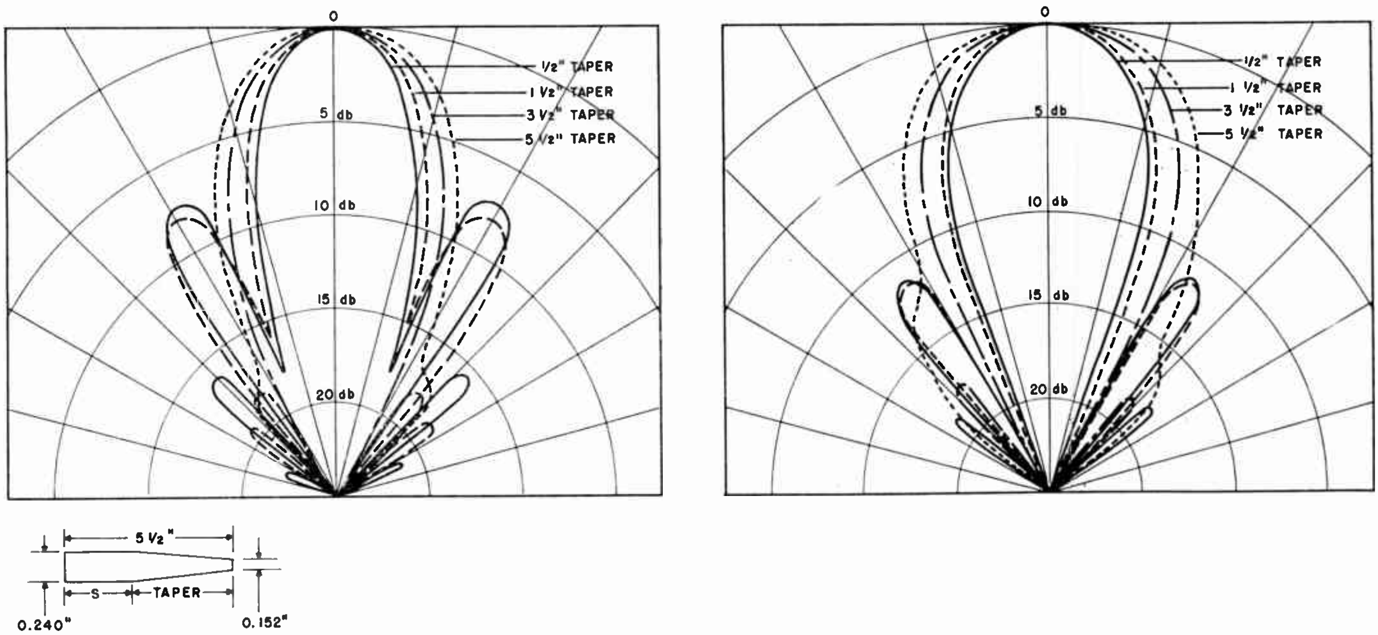


Fig. 20—Radiation patterns of single ferrod radiators 5.5 inches long. Parameter: Length of taper. End fed from $\frac{1}{2}\times 1$-inch waveguide. Gain of ferrod with $3\frac{1}{2}$-inch taper approx. 40.

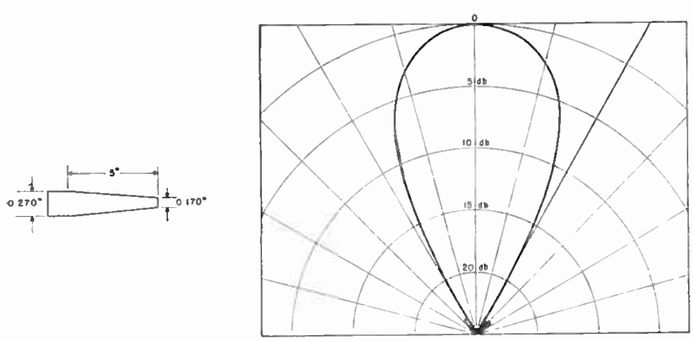


Fig. 21—Radiation pattern of a single ferrod radiator 5.4 inches long.

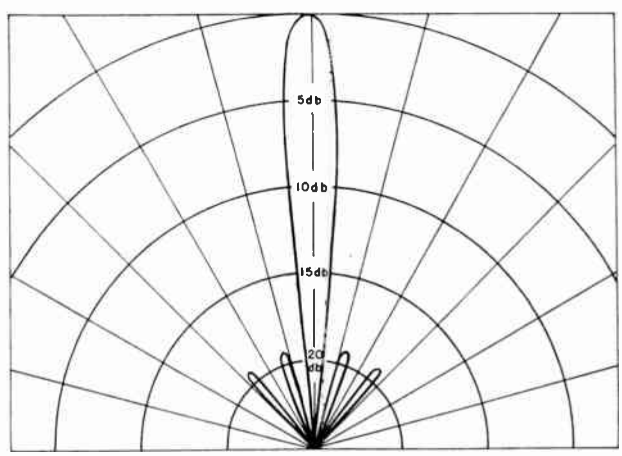


Fig. 22—Azimuthal radiation pattern of cavity-fed ($TE_{809}$) ferrod antenna.

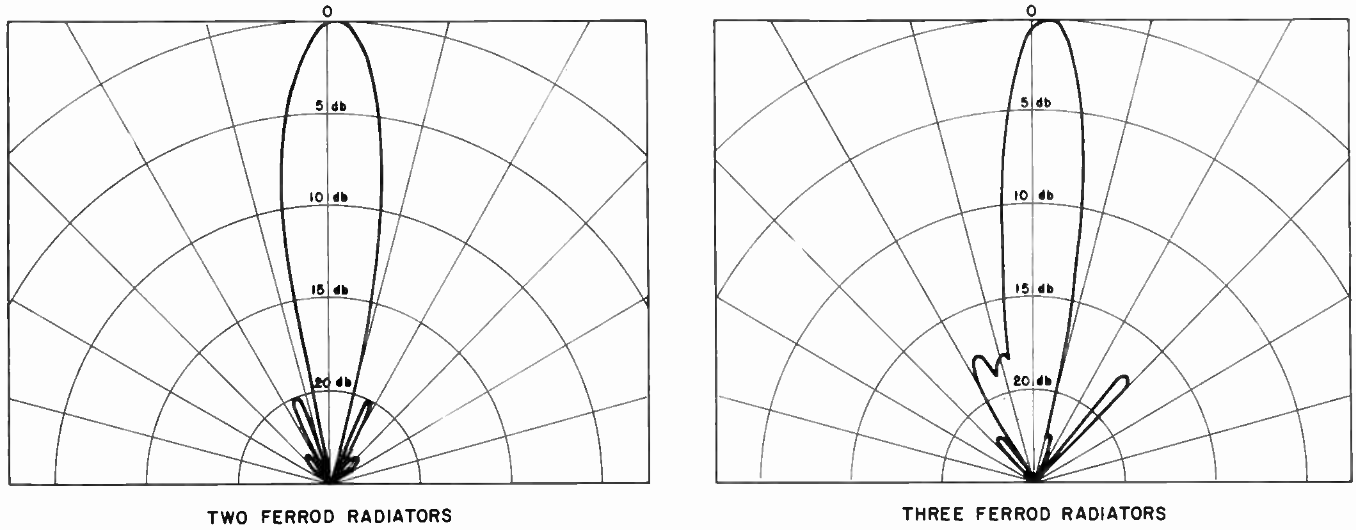


Fig. 23—Radiation pattern of two linear ferrod arrays. Radiating elements spaced approx. $3/2\lambda_0$ apart.

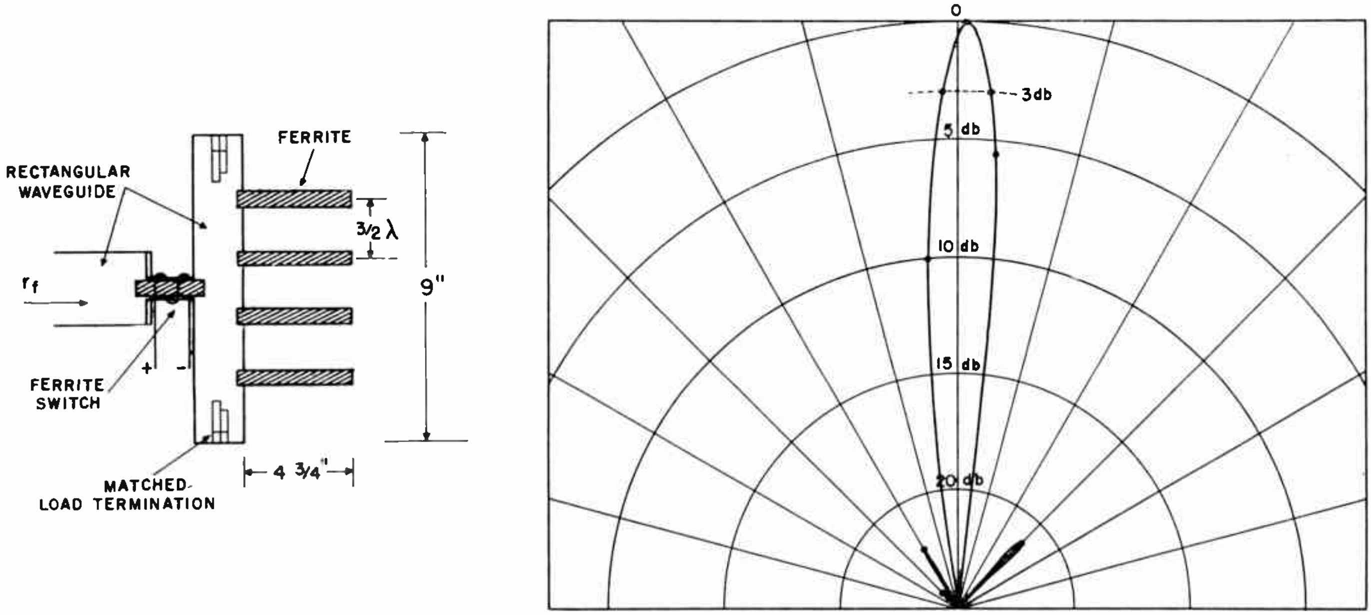


Fig. 24—Radiation pattern of a linear array of four ferrods.

or two linear arrays are shown in Fig. 23. The ferrods were 5 inches long and spaced $\lambda_0$ apart along the narrow side of a section of rectangular waveguide. The beamwidth for the arrays can be seen to decrease with increasing number of ferrods, resulting in higher gain antennas. The vertical beamwidth of both arrays was approximately that of a single rod, resulting in a fan-shaped beam. A photograph of this type antenna is shown in Fig. 3.

A simplified drawing of a 4-rod linear array and its radiation characteristics in the plane of the ferrods is given in Fig. 24. The one-half power beamwidth in this plane was approximately 7° and in the vertical plane was that of a single ferrod. Details of the ferrod radiators and the associated modulating system are shown at the left. Short matched loads are used at each end of the waveguide section to minimize internal reflections. A photograph of this antenna is shown in Fig. 4.

## Conclusion

The ferrod radiator systems described above are especially useful in the field of portable radars and have good possibilities in the miniaturization of fire-control antenna systems requiring high-speed electrical scanning or lobing. Advantages of these antenna systems include simplicity, low construction costs, no moveable parts, microsecond response to an applied magnetic field, small control currents, and the possibility of relatively broad-band operation at microwave frequencies.

## Acknowledgment

We should like to thank Mary Ann Armistead for her assistance in setting up various microwave measurements and for taking some of the data, Mary M. Borda for making the many detailed drawings, and A. J. Pannone of the National Bureau of Standards Grinding Shop for preparing the ferrods.

# IRE Standards on Piezoelectric Crystals—The Piezoelectric Vibrator: Definitions and Methods of Measurement, 1957*

**57 IRE 14. S1**

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## Introduction

In 1949, a standard on piezoelectric crystals[1] was issued which covered the definition of axes for piezoelectric crystals and their relation to orthogonal axes, standards for specifying crystal plate orientation, and nomenclature referring to the piezoelectric relations, symbols, and units. It is the object of this present standard to specify the nomenclature and a practical method of measuring the various quantities associated with piezoelectric vibrators.

## 1.0 The Piezoelectric Vibrator and Its Equivalent Circuit

### 1.1 The Piezoelectric Vibrator

A piezoelectric vibrator consists of an element cut from piezoelectric material usually in the form of a plate, bar, or ring, and with electrodes attached to or supported near the element to excite one of its resonance frequencies.

### 1.2 The Equivalent Circuit

The properties of any mode of a lightly-damped mechanical vibrating system, piezoelectrically excited through electrodes which form a two-terminal network, can be represented near resonance by an equivalent electric circuit (Fig. 1) which consists of a capacitance

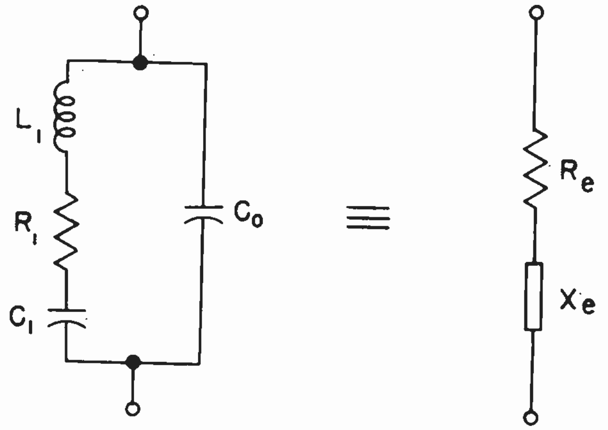



Fig. 1—Equivalent electric circuits of a piezoelectric vibrator.

$C_1$, inductance $L_1$, and resistance $R_1$, in series, shunted by a second capacitance, the parallel capacitance $C_0$.

These are the fundamental parameters of the piezoelectric vibrator. The representation of a piezoelectric vibrator by the circuit of Fig. 1 is useful only if the four parameters are constant and independent of frequency and amplitude.

The parameters are independent of frequency if the vibrator has no other mode of motion near its resonance. Generally, the mode in question is sufficiently isolated from other modes to permit this assumption. *When this is not true, the equations and measuring methods outlined herein do not apply.* The validity of the circuit representation can be determined by measuring and plotting the impedance or admittance of the vibrator as a function of frequency.

At a given frequency the parameters of the equivalent circuit generally approach constant values as the amplitude of vibration approaches zero. The amplitude which can be tolerated before the parameters are appreciably affected varies widely between vibrators of various types and can only be determined by experiment.

### 1.3 Parameters of Piezoelectric Vibrators

The four fundamental parameters $C_1$, $L_1$, $R_1$, and $C_0$ define the network completely and all other parameters may be derived from them. The important parameters of a piezoelectric vibrator are listed in Table I. The equation for the impedance $Z$ (or admittance $Y$)

$$Z = \frac{1}{Y} = \frac{1}{j}\cdot\frac{\omega_s}{\omega}\cdot\frac{R_1 Q}{r} \cdot \frac{1 + jQ\left(\frac{\omega}{\omega_s} - \frac{\omega_s}{\omega}\right)}{1 + jQ\sqrt{\frac{1+r}{r}}\left(\frac{\omega}{\omega_p} - \frac{\omega_p}{\omega}\right)} \cdots \quad (1)$$

of the equivalent electric circuit of the piezoelectric vibrator is the basic equation for deriving the relations of the various parameters. For the symbols used in (1) see Table I.

For the purpose of defining the different characteristic frequencies, the impedance $Z$ of the equivalent network, its resistive component $R_e$, its reactive component $X_e$, and the reactance of $X_1$ of the $L_1$, $C_1$, $R_1$ branch are plotted as functions of frequency in Fig. 2. These curves,

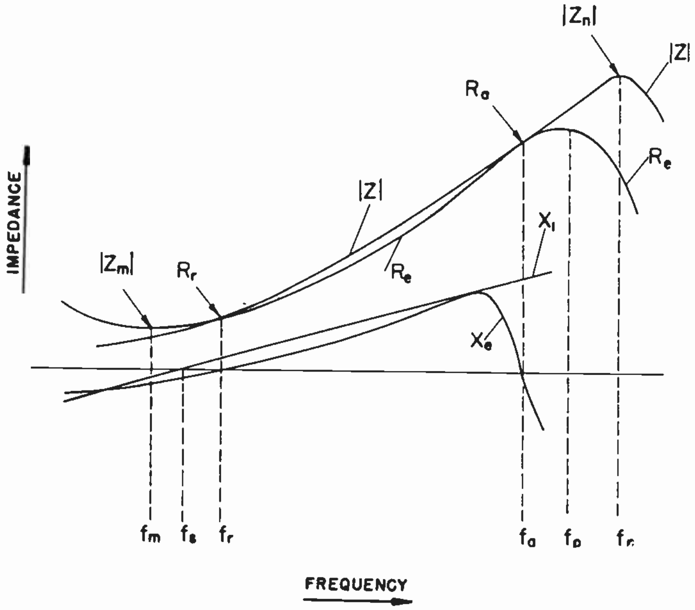



Fig. 2—Impedance $|Z|$, resistance $R_e$, reactance $X_e$, and series arm reactance $X_1$ of a piezoelectric vibrator as a function of frequency. For the meaning of the different symbols, see Table I.

[1] "Standards on Piezoelectric Crystals," Proc. IRE, vol. 37, pp. 1378–1395; December, 1949.

TABLE I

PARAMETERS OF THE EQUIVALENT CIRCUIT OF A PIEZOELECTRIC VIBRATOR

$C_1$ = Motional Capacitance
$L_1$ = Motional Inductance
$R_1$ = Motional Resistance
$C_0$ = Shunt (Parallel) Capacitance
$r = \dfrac{C_0}{C_1}$ Capacitance Ratio

Characteristic Frequencies:

$f_m$ = Frequency at Maximum Admittance (Minimum Impedance)
$f_s$ = Motional (Series) Resonance Frequency $= \dfrac{1}{2\pi\sqrt{L_1C_1}}$
$f_r$ = Resonance Frequency (Reactance $X_e=0$)
$f_a$ = Antiresonance Frequency (Reactance $X_e=0$)
$f_p$ = Parallel Resonance Frequency $= \dfrac{1}{2\pi}\sqrt{\dfrac{1}{L_1C_1}\left(1+\dfrac{1}{r}\right)}$
$f_n$ = Frequency at Minimum Admittance (Maximum Impedance)

$Z = \dfrac{1}{Y} = R_e + jX_e$ = Impedance of Vibrator
$Z_m$ = Impedance at $f_m$ (Minimum Impedance)
$Z_n$ = Impedance at $f_n$ (Maximum Impedance)
$Y_m$ = Admittance at $f_m$ (Macimum Admittance)
$Y_n$ = Admittance at $f_n$ (Minimum Admittance)
$R_r$ = Impedance at $f_r$, Resistive (Resonance Resistance)
$R_a$ = Impedance at $f_a$, Resistive (Antiresonance Resistance)

$Q = \dfrac{\omega_s L_1}{R_1}$ = Quality Factor
$M = \dfrac{Q}{r} = \dfrac{1}{\omega_s C_0 R_1}$ = Figure of Merit
$\omega = 2\pi f$
$L_0$ = Inductance in Parallel with Vibrator
$C_L$ = Load Capacitance

$R_{st}$ = Standard Resistor
$f_{m\pi}$ = Frequency of Maximum Transmission
$f_{sL}$ = Motional Resonance Frequency of Combination of Vibrator and $C_L$
$f_{mTL}$ = Frequency of Maximum Transmission of Combination Vibrator and $C_L$
$R_{1L}$ = Resulting Motional Resistance of Combination DF Vibrator and $C_L$

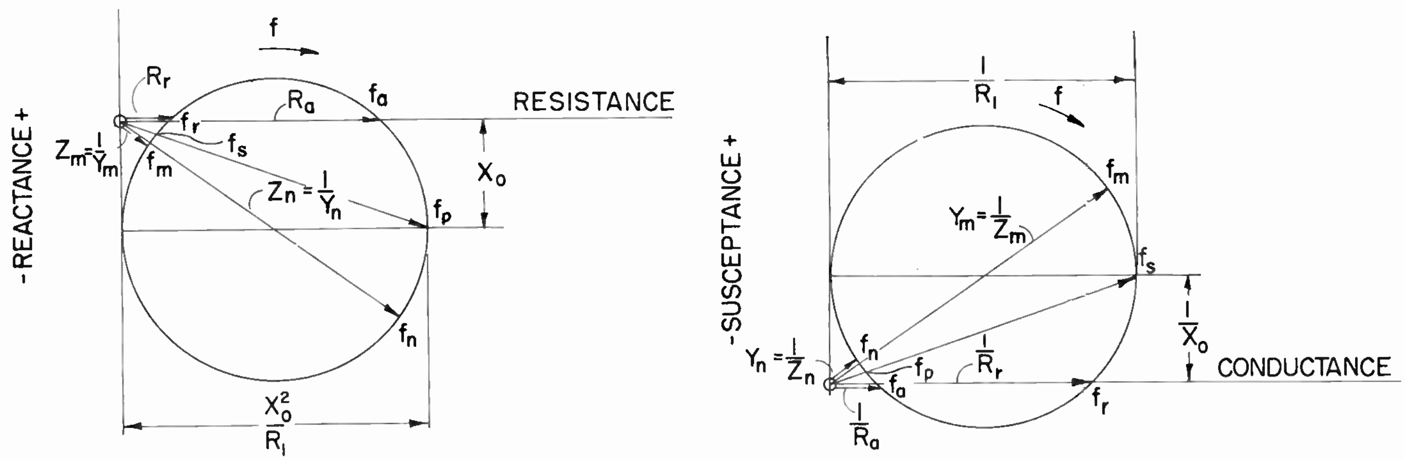



Fig. 3—Impedance and admittance diagram of a piezoelectric vibrator. The symbols conform with those used in Table I and Fig. 2.

however, have only qualitative character and do not represent a particular piezoelectric vibrator. Further, to clarify the situation, the impedance and admittance circles of a piezoelectric vibrator are reproduced in Fig. 3. The circle representation of the impedance or admittance of a piezoelectric vibrator is valid only if, in the admittance diagram, the circle diameter is large in comparison with the change of $\omega C_0$ in the resonance range. If the latter condition is not fulfilled, the admittance curve shows a more complicated character.

TABLE II

RELATIONS BETWEEN THE CHARACTERISTIC FREQUENCIES OF A PIEZOELECTRIC VIBRATOR

| CHARACTERISTIC FREQUENCY | 1ST APPROXIMATION | | 2ND APPROXIMATION | |
|---|---|---|---|---|
| | $\frac{f}{f_s}$ | DEVIATION $\frac{\Delta f}{f_s}$ FROM MORE PRECISE VALUE | $\frac{f}{f_s}$ | DEVIATION $\frac{\Delta f}{f_s}$ FROM MORE PRECISE VALUE |
| $f_m$ | $\frac{f_m}{f_s}=1$ | $-\frac{1}{2M^2r}$ | $\frac{f_m}{f_s}=\sqrt{1+\frac{1}{2r}\left[1-\sqrt{1+\frac{4}{M^2}}\right]}$ | $\frac{1}{2M^4r^2}$ |
| $f_r$ | $\frac{f_r}{f_s}=1$ | $\frac{1}{2M^2r}$ | $\frac{f_r}{f_s}=\sqrt{1+\frac{1}{2r}\left[1-\sqrt{1-\frac{4}{M^2}}\right]}$ | $\frac{1}{2M^4r^2}$ |
| | $\frac{f_a}{f_s}=1+\frac{1}{2r}$ | $-\frac{1}{2M^2r}\left(\frac{1}{r}+1\right)$ | $\frac{f_a}{f_s}=\sqrt{1+\frac{1}{2r}\left[1+\sqrt{1-\frac{4}{M^2}}\right]}$ | $-\frac{1}{2M^2r}\cdot\frac{1}{r}$ |
| $f_n$ | $\frac{f_n}{f_s}=1+\frac{1}{2r}$ | $\frac{1}{2M^2r}\left(\frac{1}{r}+1\right)$ | $\frac{f_n}{f_s}=\sqrt{1+\frac{1}{2r}\left[1+\sqrt{1+\frac{4}{M^2}}\right]}$ | $\frac{1}{2M^2r}\cdot\frac{1}{r}$ |
| $p$ | $\frac{f_p}{f_s}=1+\frac{1}{2r}$ | $-\frac{1}{8r^2}$ | $\frac{f_p}{f_s}=\sqrt{1+\frac{1}{r}}$ | 0 |

Since (1) is nonlinear in frequency, it is convenient to make certain approximations in deriving practical equations for general use. It is the error of these approximations, in addition to the instrumentation errors, that governs the over-all accuracy of the derived parameters. Table II gives relations between the characteristic frequencies, $f_m$, $f_r$, $f_a$, $f_p$, $f_n$ and the series resonance frequency $f_s$ of a vibrator for two different degrees of approximation, and particularly the deviation of the characteristic frequencies from $f_s$. The first approximation is commonly used in the case of vibrators with high $M$ and $r$; *i.e.*, high $Q$, as for instance in the case of quartz vibrators. More exact equations than those of the first approximation have been derived by various authors.[2,3] They are shown in Table II as second approximation. The accuracy of the two approximations is given in columns 3 and 5 in Table II. For a vibrator with $M^2r=1000$, for instance, the error in determining $f_s$ from $f_m$ would be $-5\cdot10^{-4}$ for the first approximation and $-5\cdot10^{-7}$ for the second approximation.

## 2.0 METHODS OF MEASURING THE PARAMETERS OF THE EQUIVALENT CIRCUIT

### *2.1 Measurements—General*

For production tests, where only frequency and a quality factor are of concern, oscillators are generally used. In this method, the vibrator is part of the oscillator circuit and controls the frequency and magnitude of the output voltage. It is one of the most convenient methods of measurement, particularly for determining effects of temperature variation, and with proper precautions can be used for the determination of all the parameters. The errors due to instrumentation, however, are not as readily specified or controlled as for some of the methods where the vibrator is used as a passive element, and, therefore, oscillators are not as suitable for standard test methods.

Three of the most commonly used methods of the passive type are 1) bridge,[4] 2) Q meter,[4] and 3) transmission circuit[5] methods.

The bridge method may be extended to the measurement of impedance (or admittance) as a function of frequency in the most extreme conditions. This procedure is laborious and generally unwarranted when dealing with an isolated mode. For measurement of critical frequencies and impedances, bridges are generally limited with respect to frequency and impedance range. The transmission circuit is recommended for use in laboratory tests because it is simple, convenient, and involves equipment normally available. It is also used extensively for production tests. With proper consideration of errors of instrumentation and suitable equations, this method is the best suited for the ranges of $r$ and $M$ generally encountered.

[2] G. E. Martin, "Determination of equivalent-circuit constants of piezoelectric resonators of moderately low Q by absolute-admittance measurements," *J. Acoust. Soc. Amer.*, vol. 26, pp. 413–420; May, 1954.

[3] E. A. Gerber, "A review of methods for measuring the constants of piezoelectric vibrators," PROC. IRE, vol. 41, pp. 1103–1112; September, 1953.

[4] W. D. George, M. C. Selby, and R. Scolnik, "Precision measurement of electrical characteristics of quartz crystal units," PROC. IRE, vol. 36, pp. 1122–1131; September, 1948.

[5] L. F. Koerner, "Progress in development of test oscillators for crystal units," PROC. IRE, vol. 39, pp. 16–26; January, 1951.

### 2.2 Transmission Circuit Method

*2.2.1 Type of Measurement:* This method of measurement consists of determining the frequency and the impedance at maximum transmission of a $\pi$ network in which the vibrator under test is the series branch. The frequency at maximum transmission is measured with and without a capacitance $C_L$ in series with the vibrator. From these measurements, $f_s$, $C_1$, and $R_1$ can be determined.

*2.2.2 Circuit:* The measuring circuit consists of a variable-frequency oscillator, detector, measuring network, and frequency measuring equipment connected in accordance with Fig. 4. The oscillator must have a high degree of purity of output waveform.

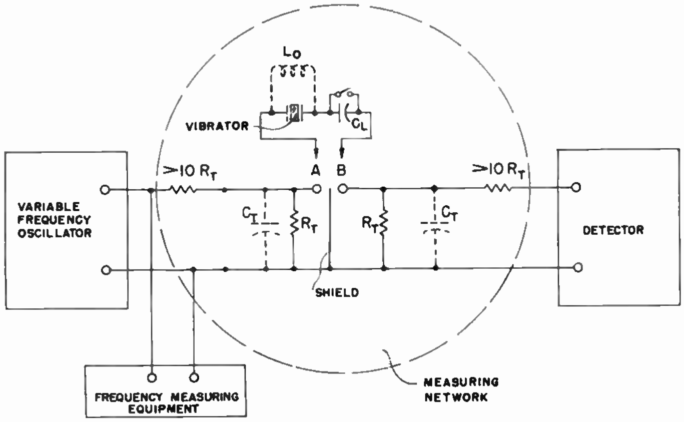



Fig. 4—Transmission measuring circuit.

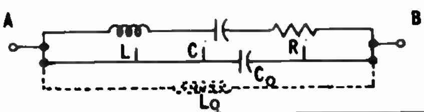
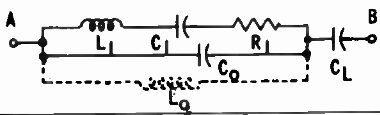
*2.2.3 Procedure*

*2.2.3.1 Motional Resonance Frequency $f_s$:* The frequency of the oscillator is adjusted for maximum transmission with the piezoelectric vibrator connected to terminals $A$ and $B$ of the network. This is the frequency of maximum transmission $f_{mT}$ of the network. At the first approximation, $f_{mT}$ is equal to the frequency of minimum impedance $f_m$ and the motional resonance frequency $f_s$ of the vibrator. If higher accuracy for $f_s$ is required, Table III should be consulted which gives

TABLE III

FREQUENCY ERROR $\dfrac{\Delta f}{f}=\dfrac{f_{mT}-f_s}{f_s}$ AND RESISTANCE ERROR $\dfrac{\Delta R}{R}=\dfrac{R_{st}-R_1}{R_1}$ FOR SERIES RESONANCE OPERATION, AND FREQUENCY ERROR $\dfrac{\Delta f}{f}=\dfrac{f_{mTL}-f_{sL}}{f_{sL}}$ AND RESISTANCE ERROR $\dfrac{\Delta R}{R}=\dfrac{R_{st}-R_{1L}}{R_{1L}}$ FOR POSITIVE REACTANCE OPERATION.

THE SYMBOLS USED IN THIS TABLE ARE EXPLAINED IN TABLE I AND FIG. 4

| | | SERIES RESONANCE OPERATION | | POSITIVE REACTANCE OPERATION | |
|---|---|---|---|---|---|
| 1) | |  | |  $\frac{C_L+C_o}{C_L}=u$ | |
| 2) | | $\frac{\Delta f}{f}$ ; $\frac{\Delta R}{R}$ | DEVIATION OF $\frac{\Delta f}{f}$ FROM MORE PRECISE VALUE FOR FREQUENCY ERROR | $\frac{\Delta f}{f}$ ; $\frac{\Delta R}{R}$ | DEVIATION OF $\frac{\Delta f}{f}$ FROM MORE PRECISE VALUE FOR FREQUENCY ERROR |
| 3) | FREQUENCY ERROR $\frac{\Delta f}{f}$ DUE TO $C_o$ AND $R_T$ | $-2\frac{R_T}{R_1}\frac{1}{M^2r}$ | $-\frac{1}{2M^2r}\frac{1}{1+\frac{4}{M^2}\frac{R_T}{R_1}}$ | $-2\frac{R_T}{R_1}\frac{1}{M^2r}\frac{1}{u}$ | $-\frac{u}{2M^2r}\frac{1}{1+\frac{4}{M^2}\frac{R_T}{R_1}}$ |
| 4) | FREQUENCY ERROR $\frac{\Delta f}{f}$ DUE TO $C_o$ AND $R_T$ WITH PARALLEL $L_o$ | $-4\frac{R_T}{R_1}\cdot\frac{1}{M^2r}\beta$ $\beta=\frac{1}{2}\left(1-\frac{L_oC_o}{L_1C_1}\right)$ | $-\frac{1}{M^2r}\beta\frac{1}{1+\frac{8}{M^2}\frac{R_T}{R_1}\beta^2}$ | $-4\frac{R_T}{R_1}\frac{1}{M^2r}\beta\frac{1}{u}$ | $-\frac{u}{M^2r}\beta\frac{1}{1+\frac{8}{M^2}\frac{R_T}{R_1}\beta^2}$ |
| 5) | FREQUENCY ERROR $\frac{\Delta f}{f}$ DUE TO $C_T$ | $\frac{R_T^2C_T}{R_1^2C_o}\cdot\frac{1}{M^2r}$ | $-\left(4\frac{R_T}{R_1}+1\right)\frac{1}{2M^2r}$ | $\frac{R_T^2C_T}{R_1^2C_o}\frac{1}{M^2r}\frac{1}{u^2}$ | $-\left(4\frac{R_T}{R_1u^2}+1\right)\frac{u}{2M^2r}$ |
| 6) | FREQUENCY ERROR $\frac{\Delta f}{f}$ DUE TO DETECTOR SENSITIVITY S $S=\frac{\text{SMALLEST DETECT. CURR. CHG}}{\text{CURRENT}}$ | $\left(2\frac{R_T}{R_1}+1\right)\frac{.707\sqrt{S}}{Mr}$ | $-\left(2\frac{R_T}{R_1}+1\right)^2\frac{S}{M^2r}$ | $\left(2\frac{R_T}{R_1u^2}+1\right)\frac{.707\sqrt{S}}{Mr}$ | $-\left(2\frac{R_T}{R_1u^2}+1\right)^2\frac{u}{M^2r}S$ |
| 7) | FREQUENCY ERROR $\frac{\Delta f}{f}$ DUE TO ERROR IN $C_L$ EQUAL TO $\Delta C_L$ | ——— | ——— | $-\frac{1}{2}\frac{\Delta C_L}{C_L}\frac{C_o}{C_L}\frac{1}{u^2r}$ | ——— |
| 8) | RESISTANCE ERROR $\frac{\Delta R}{R}$ DUE TO $R_T$ & $C_T$ | $\left(\frac{1}{2}\frac{R_TC_T}{R_1C_o}-1\right)\cdot\frac{8R_T}{R_1M^2}$ | ——— | $\frac{8R_T}{R_1M^2}\left(\frac{1}{2u}\frac{R_TC_T}{R_1C_o}-1\right)$ | ——— |

under the heading, "Series Resonance Operation," in lines 3 and 5, the relationship between $f_{mT}$ and $f_s$, as the ratio

$$\frac{\Delta f}{f} = \frac{f_{mT} - f_s}{f_s},$$

for two different degrees of approximation, as a function of the network parameters.

*2.2.3.2 Motional Capacitance $C_1$:* The measurement of 2.2.3.1 is repeated, but with a particular $C_L$ in series with the vibrator. The motional resonance frequency for this combination, $f_{sL}$, can again be derived from the measured frequency at maximum transmission $f_{mTL}$, with the help of lines 3, 5, and 7 of Table III, under the heading, "Positive Reactance Operation." $C_1$ is then obtained from $C_L$, $f_s$, $f_{sL}$, and $C_0$:

$$C_1 = 2\frac{f_{sL} - f_s}{f_s}(C_0 + C_L). \qquad (2)$$

$C_0$ can be eliminated and the accuracy of measurement is increased by using two or more different values of $C_L$.

*2.2.3.3 Motional Resistance $R_1$:* The detector reading is adjusted to some convenient reference point and noted. The vibrator is replaced by a standard resistor $R_{st}$ (variable or selected value) to give the same reading as the maximum transmission value for the vibrator. This value of resistance gives approximately the magnitude of minimum impedance $|Z_m|$ of the vibrator. At a first approximation, $R_{st} = |Z_m|$ is equal to $R_1$. Table III furnishes in line 8 the necessary correction formulas for better approximations. For the sake of completeness, correction data for the resulting motional resistance $R_{1L}$ of the combination vibrator-$C_L$ are also given.

*2.2.4 Network Requirements:* The accuracy of the results increases as the following conditions are approached:

1) Reactance $X_T$ of terminating capacitance $C_T$ much greater than terminating resistance $R_T$, ($X_T \gg R_T$).
2) Stray capacitance $C_{A-B}$ between terminals $A$ and $B$ low compared to vibrator capacitance $C_0$, ($C_0 \gg C_{A-B}$).
3) Reactance of stray capacitance $C_{A-B}$ high compared to series resistance $R_1$, ($X_{A-B} \gg R_1$).
4) Inductance of leads connecting vibrator low compared to inductance of vibrator.

In the case of vibrators with low figure of merit $M$, it is advisable to use a shunting coil $L_0$ connected in parallel with the vibrator. If the combination $L_0C_0$ is tuned close to the motional resonance frequency $f_s$ of the vibrator, $f_s$ and $R_1$ will be measured instead of $f_{mT}$ and $|Z_m|$. Table III, line 4, supplies information as to the accuracy of the determination of $f_s$ as a function of $L_0C_0/L_1C_1$.

Table III supplies additional information in line 6 as to the influence of the detector sensitivity (smallest detectable current-change/current) on the determination of $f_s$ and $f_{sL}$.

*2.3 Measurement of $C_0$*

$C_0$ of the equivalent vibrator circuit is slightly less than the measured value for a free piezoelectric element and slightly greater than the measured value of the clamped condition of the piezoelectric element. There is, therefore, no simple direct method of measuring $C_0$ exactly. This difference, however, is generally sufficiently small to be neglected unless the vibrator has a small capacitance ratio; *i.e.*, $r < 50$.

$C_0$ is usually determined in a capacitance bridge or in the $Q$-meter at a frequency sufficiently lower than that of all modes of vibration of the vibrator to reduce their effect as far as possible. Where greater accuracy is required, use must be made of the fact that the dielectric permittivity depends upon the mode of vibration. This will be discussed in a subsequent standard.

# Correction

The following correction to "IRE Standards on Methods of Measurement of the Conducted Interference Output of Broadcast and Television Receivers in the Range of 300 KC to 25 MC, 1956 (56 IRE 27 S1)," which appeared on pages 1040–1043 in the August, 1956, issue of PROCEEDINGS OF THE IRE, has been brought to the attention of the editors.

On page 1041, the last sentence of paragraph 2.2.2 should read:

(For this requirement the power plug is short-circuited.)

# Correspondence

## Microwave Ferrite Phase Shifter*

A ferrite phase shift device which appears to have some promise as a sideband modulator or as part of a high speed switch has been tested, and some of the results are reported here.

In one form the arrangement consists of a helical transmission line with a ferrite tube mounted inside the helix. Not only does the ferrite-loaded helix act as a transmission system whose phase shift (or insertion loss) can be varied by varying the permeability of the ferrite, but also as a solenoid which, when properly excited, supplies the low-frequency field that varies this permeability.

When tested as a sideband modulator, the particular arrangement used generated first-order sidebands which were traced continuously, with the exception noted below, from 70 kc to 50 mc with the carrier at 9 kmc. The lower limit was a result of the resolution of the available spectrum analyzer; it is believed that much lower frequency sidebands can be as easily, if not more easily, generated. The upper limit was a result of the high-frequency limit of readily available oscillators; sideband generation was still quite strong at 50 mc. The sidebands were considerably reduced in magnitude and essentially disappeared in a region about 2 mc wide centered at a modulating frequency of 17 mc. This was possibly a domain wall resonance phenomena since the frequency at dropout was somewhat influenced by the dc axial biasing field.

Some understanding of the manner in which the device works can be obtained from an examination of the static transmission characteristics shown in Fig. 1. As might be expected, these transmission characteristics were essentially the same for either direction of axial magnetic field, except for slight hysteresis effects near zero field. In order to test the device as a sideband modulator, the helix was excited at one end (end A—Fig. 2) by a low-frequency oscillator, while the other end (end B—Fig. 2) was terminated in a 50-ohm resistor. By-pass chokes of the usual variety were built into the coaxial low-frequency input and output lines, as shown in Fig. 2, to minimize the microwave energy leakage. Energy at the carrier frequency of 9 kmc was fed into one of the waveguides, and the other waveguide was connected to a spectrum analyzer. If one assumes the phase shift per unit field is linear over a small range of magnetic field, it can be readily shown that for sinusoidal low-frequency modulation the amplitude of the $n$th order side bands is given by $J_n(sA)$, the $n$th order Bessel function of argument $sA$, where

$s$ = total radian phase shift per unit axial modulating field,
$A$ = amplitude of axial modulating field.

From the data in Fig. 1, using the catalog quoted power output of the modulating frequency sources, and assuming the helix performs as an ideal sheath solenoid, it was estimated that in the presence of a small axial dc biasing field the modulation index $sA$ would be approximately 0.01. This leads to first-order sidebands about 46 db down from the carrier. The measured magnitudes of these sidebands were indeed of this order. They were essentially equal, and the variation in their amplitude as a function of a superimposed axial dc field was exactly as would be predicted from the static data.

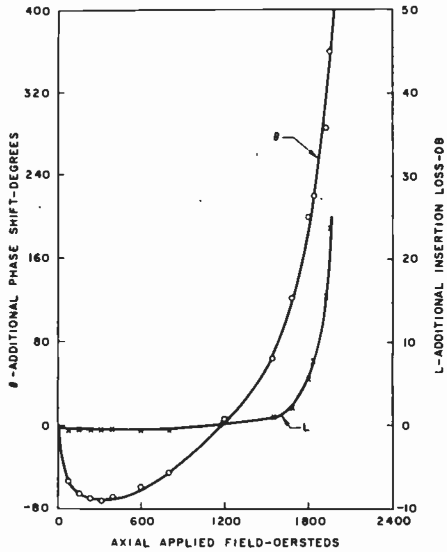



Fig. 1—Additional phase shift and insertion loss of the ferrite-loaded helix as a function of an applied axial magnetic field.

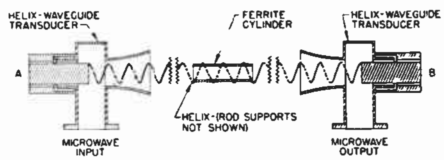



Fig. 2—Cross section of ferrite-loaded helix and end transducers. Helix has an inner diameter of 0.125 inch and is wound of 0.22-inch diameter molybdenum wire with 30 turns per inch. Ferrite cylinder is made of Ferramic R-1 material and is 1 inch long with 0.124-inch outer diameter and 0.094-inch inner diameter. The helix is supported by three sapphire rods (not shown) spaced equally around helix circumference. Waveguide is standard RG-52/U or WR90 with dimensions 1 × ½ × 0.050 (wall) inches.

In summary, the possibilities available from a ferrite-loaded helix as a phase shifter have been demonstrated. The frequency response is believed to be several orders of magnitude higher than any presently available similar devices, with the upper limit probably determined primarily by the ferrite material and not the physical arrangement. Single-sideband generation would appear possible by the use of sawtooth-like modulation on the helix. To increase the phase shift per unit modulating field it is evident that more ferrite loading is required, which can be accomplished, for example, by increasing the length of the ferrite tube or increasing its wall thickness. It also appears clear that the helix parameters as well as the type of ferrite will also influence the modulation index.

The writer would like to acknowledge the helpful discussions held with J. T. Mendel as well as the assistance of G. Lee and A. Iversen, who supplied the rod-supported helix.

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* Received by the IRE, November 19, 1956.

## Stability Requirements and Calibration of Radiometers when Measuring Small Noise Powers*

A radiometer is a receiving equipment designed to measure noise power, usually to an accuracy of better than 1°K in effective noise temperature. In a simple radiometer, where the output is a function of the total noise power at the input, excellent receiver stability is required since equivalent receiver input temperatures of 500° to 3000°K are ordinarily encountered. Dicke modified the simple radiometer such that it periodically measured the difference in noise temperature between a reference termination and the antenna, and thus partially eliminated the need for excellent stability.[1,2] However, measurement of noise temperatures as low as a fraction of 1°K imposes severe requirements on receiver stability even for an improved Dicke system. This note discusses the stability requirements for radiometers when measuring small noise temperatures, and proposes a modification of the Dicke system to alleviate the severe stability requirement. Using this modified Dicke system, it also becomes possible to calibrate small noise temperatures using ordinary noise sources.

In the radiometer proposed by Dicke, an attenuator at temperature $T_0$ is periodically inserted between the antenna and receiver terminals. The system is designed to measure the difference between the detector output voltage with the receiver connected to the antenna and that with the receiver connected to the attenuator. The following analysis assumes sufficient smoothing in the radiometer (a large bandwidth-time constant product) so that residual noise fluctuations do not limit receiver sensitivity.

* Received by the IRE, November 9, 1956.

[1] R. H. Dicke, "The measurement of thermal radiation at microwave frequencies," *Rev. Sci. Instr.*, vol. 17, pp. 268–275; July, 1946.

[2] C. H. Mayer, "Improved microwave noise measurements using ferrites," IRE Trans., vol. MTT-4, pp. 24–28; January, 1956.

The difference, or output, voltage, $V_R$, from a Dicke radiometer with a square-law detector (assuming receiver noise factor and gain do not vary appreciably during a sampling period) can be written as:

$$V_R = CG[(F-1)T_0 + T_A] - CGFT_0 \quad (1)$$

$T_A$ = effective antenna temperature in degrees K

$G$ = receiver power gain

$F$ = receiver noise factor

$T_0$ = attenuator temperature in degrees K

$C$ = a constant.

If the radiometer characteristics drift, $V_R$ will also drift. To measure $T_A$ to an accuracy $\Delta T$, it is necessary that the change in $V_R$ due to radiometer drift be small compared with the change in $V_R$ due to a variation of $\Delta T$ in $T_A$.

This stability criterion can be expressed as:

$$\frac{\partial V_R}{\partial F}\Delta F + \frac{\partial V_R}{\partial G}\Delta G \ll \frac{V_R}{\partial T_A}\Delta T. \quad (2)$$

Substituting (1) in (2) we obtain the stability criterion:

$$(T_A - T_0)\frac{\Delta G}{G} \ll \Delta T. \quad (3)$$

Eq. (3) indicates the accuracy of temperature measurement to be independent of drifts in receiver noise factor, which is not true when a linear detector is used. It also indicates that for a given accuracy the allowable gain drift is a function of $(T_A - T_0)$, and for values of $T_A$ near $T_0$ the system is most insensitive to gain drift. [It is interesting to note that (3) for a simple radiometer with a square-law detector becomes:

$$\frac{\Delta G}{G}[T_A + (F-1)T_0] + T_0\Delta F \ll \Delta T \quad (4)$$

indicating that the accuracy is dependent on receiver gain and noise factor drifts, and that for small values of $T_A$ the simple radiometer is considerably more sensitive to receiver drifts than the Dicke system.]

In practice it is difficult to keep gain drifts below about 0.1 per cent. Thus for a Dicke system using a square-law detector it becomes difficult to measure noise temperature below about 0.3°K when $T_0$ is 300°K. If it were practical to reduce $T_0$ to near 0°K, this minimum detectable temperature could be reduced considerably.

Another method for reducing the minimum detectable temperature is as follows: A variable noise source is coupled to one arm of a directional coupler which is inserted between the antenna and attenuator terminals. The coupler is arranged so that the noise introduced by the noise source is coupled into the receiver, and using a well-designed coupler there is negligible change in receiver input impedance or antenna noise power transferred to the receiver. If the noise source is adjusted so that the additional noise temperature coupled into the receiver is fixed at $T_0$, the system becomes very insensitive to receiver drifts when measuring small values of $T_A$. By varying the amount of additional noise temperature about $T_0$ so to keep the receiver output voltage at zero as $T_A$ varies (while recording the amount of noise source variation required to do this) a system similar to that of Machin, Ryle, and Vonberg is obtained except that it is now possible to calibrate values of $T_A$ below room temperature.[3] Here the excess noise temperature of the noise source coupled into the receiver (temperature relative to $T_0$) is a direct measure of $T_A$.

The basic problem involved in determining the absolute value of $T_A$ with this modified Dicke system is essentially the same as that with the original Dicke system or any comparison system. It is the problem of determining reference temperatures to an accuracy compatible with the desired radiometer accuracy. This problem appears to be the remaining limitation on absolute measurement of small noise powers.

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[3] K. E. Machin, M. Ryle, and D. D. Vonberg, "The design of an equipment for measuring small radio-frequency noise powers," *Proc. IEE*, vol. 99, part III, pp. 127–134; May, 1952.

## Low Frequency Dispersion of $\rho$ and $\epsilon$ in Ferrites*

In a recent integrating paper on dielectric properties and conductivity in ferrites, Van Uitert[1] referred to the fundamental work of Koops[2] on Ni/Zn ferrites and his two-layer model. He pointed out—as was formerly suggested by others[3]—the critical dependence on firing temperature of this model and, interpreting the data published by Koops, brought some evidence of a dependence on firing conditions of the thickness of the superficial layer.

We think that this important point requires a more careful examination. If the notations lf and hf are connected respectively with low and high frequency measurements and the notations 1 and 2 with bulk and surface layer of the grain, the ratio $x$ of the layer thickness to the grain diameter is easily obtained from the expressions given by Koops for $\rho$ and $\epsilon$:

$$x = \frac{(\rho_{lf} - \rho_{hf})^2}{\frac{\epsilon_{lf}}{\epsilon_{hf}}\rho_{lf}^2 - \rho_{hf}^2} \qquad \frac{\epsilon_{hf}}{\epsilon_{lf}}\left(1 - \frac{\rho_{hf}}{\rho_{lf}}\right)^2.$$

By plotting against frequency $\rho$ and $\epsilon$ experimental data given by him for $Ni_{0.4}Zn_{0.6}Fe_2O_4$ at various sintering atmospheres and maximum sintering temperatures, and by extrapolating these curves for low and high frequencies, $\rho$ and $\epsilon$ values are obtained. Taking an average grain diameter of 1 micron and a bulk permittivity $\epsilon_{hf}$ of the order of 20, the thickness is given by Table I.

TABLE I

| Sintering at 1300°C | Cooling | Per cent $Fe^{2+}$ after sintering | Thickness in angströms |
|---|---|---|---|
| air | fast in air | 0.42 | 1.0 |
| oxygen | fast in air | 0.38 | 5.4 |
| air | slow in air | 0.07 | 20 |
| oxygen | slow in oxygen | 0.1 | 32 |

The thickness values give evidence of two different oxidation mechanisms. The two first data (some angströms) are in good agreement with the idea of oxygen chemisorption during cooling, more important, of course, in oxygen than in air, but which affects only the surface of crystals. The thickness found for the two other data, ten times larger, suggests on the other hand a more complete oxidation mechanism, with an important decrease of the concentration in the ferrous ions formed during the firing at 1300°C.

In 1955 we studied the occurrence of this second mechanism in hematite and in Mn/Zn ferrite and considered the hypothesis of the existence of an interstitial lattice of oxygen ions on a thickness of several crystalline cells. Its formation, by diffusion from the chemisorbed layer, would require, of course, a high number of defects in the crystals and would give, therefore, a good account of the dependence on the maximum firing temperature and the cooling conditions. As a matter of fact, the curves' resistivity vs firing temperature gave us a lot of information about the grain structure of the ceramic material.

We thought that structure-sensitive phenomena like valency induction and ferrimagnetism itself must be suppressed by the second mechanism. We were able to verify this point of view on hematite and on Mn/Zn ferrite and could understand a lot of formerly unexplained facts about thermistor materials and high permeability ferrites. We were surprised to find among the consequences of this oxidation mechanism the large time decrease of permeability reported by Snoek for the Mn/Zn ferrite.[4]

We already mentioned, in connection with Van Uitert's former work[5] that minor additions could prevent the formation of such oxidized layers.[6]

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* Received by the IRE, December 10, 1956.

[1] L. G. Van Uitert, "Dielectric properties of and conductivity in ferrites," Proc. IRE, vol. 44, pp. 1294–1303; October, 1956.

[2] G. C. Koops, "On the dispersion of resistivity and dielectric constant of some semiconductors at audio frequencies," *Phys. Rev.*, vol. 83, pp. 121–125; July, 1951.

[3] L. C. F. Blackman and N. P. R. Sherby, "Note on an investigation of the anomalous time-constant of certain iron deficient magnesium manganese ferrites," *J. Electronic (Gr. Brit.)*, vol. 1, pp. 385–388; January, 1956.

[4] J. P. Suchet, "Structures granulaires à couche superficielle dans les céramiques à base d'oxyde de fer," *J. Phys. Radium*, to be published.

[5] L. G. Van Uitert, "High resistivity nickel ferrites. The effect of minor additions of Mn or Co," *J. Chem. Phys.*, vol. 24, pp. 306–314; February, 1956.

[6] J. P. Suchet, "Effect of minor additions on the resistivity of ferrites," *J. Chem. Phys.*, vol. 25, p. 368; August, 1956.

## Graphical Method of Determining the Efficiency of Two-Port Networks*

In a recent contribution to this column it was shown[1] how a method, introduced by Altschuler,[2] for finding the maximum efficiency of an arbitrary two-port network, which is terminated in its conjugate-image impedance match, could be given a simple geometric explanation. We shall now show how the efficiency of a two-port network, terminated in an arbitrary load, can be obtained by similar graphical constructions.

The common method in analyzing a two-port network is to perform three measurements (immittance or reflection-coefficient measurements), having the output terminated in pure reactances. If we assume that the complex impedance plane is stereographically mapped on the unit sphere (Fig. 1), three points on the great circle in the $yz$-plane are transformed into three points on the right hemisphere. Through the latter points a circle, the image circle of the great circle, can be drawn.

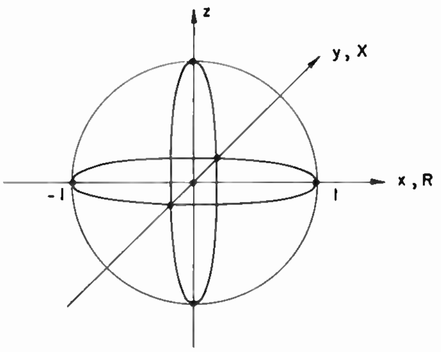

Fig. 1—The unit sphere.

First, the image circle is moved until it is symmetric with both the $xz$ plane and the $xy$ plane. This can be done in two steps, each corresponding to a lossless network. In the first step, the image circle is transformed so that it is symmetric with the $xz$ plane. Several methods are possible, the simplest being the ones of Weissfloch[3] and Wheeler-Dettinger.[4] Weissfloch extracts a series reactance, corresponding to a parabolic transformation that has its fixed point at the top of the sphere; Wheeler-Dettinger extracts a piece of uniform transmission line, corresponding to an elliptic transformation that has its axis of rotation coalescing with the $x$ axis. In Fig. 2 the transformed image circle is shown in the $xz$ plane as a straight line $C$. The symmetry to the $xy$ plane is now easily obtained by a hyperbolic transformation along the $z$ axis; this corresponds to an ideal transformer. The projection $C$ is transformed into $C'$ (see Fig. 2).

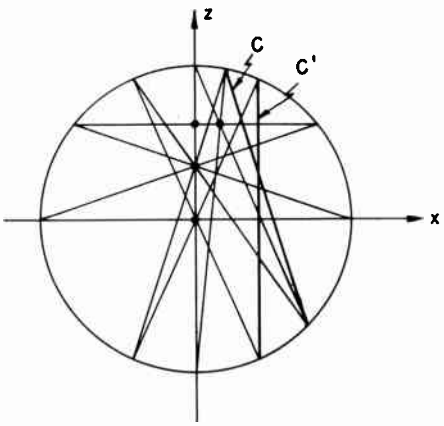

Fig. 2—Graphical construction $C \to C'$, which corresponds to an ideal transformer.

In a basic research work, Deschamps[5] has shown that the efficiency of a two-port network can be expressed as the ratio of two "pseudo distances" from a point $P$ on the sphere to two circles $C_1$ and $C_2$. This ratio is invariant by inversion; therefore it is not influenced by transformations like the ones already performed above. In Fig. 3 we interpret the circle $C_1$ as $C'$ and $C_2$ as the great circle in the $yz$ plane. An arbitrary load will, after transformation through the network, correspond to a point on the sphere to the right of $C'$. For simplicity, let us select the point $P$ shown in Fig. 3. Using the notations of the figure, we can write the efficiency $\eta$ of the two-port network:

$$\eta = \frac{h_1}{d_1} \cdot \frac{d_2}{h_2}, \tag{1}$$

where $d_2 = 1$, because it is the radius of the unit circle. The use of similar triangles immediately yields a distance $OP_1 = 1/\eta$ on the $z$-axis. The polar to $P_1$ cuts the $z$ axis at $P_2$, so that $OP_2 = \eta$. Fig. 4 shows a simple construction for obtaining $P_2$ without any constructions outside the unit circle.

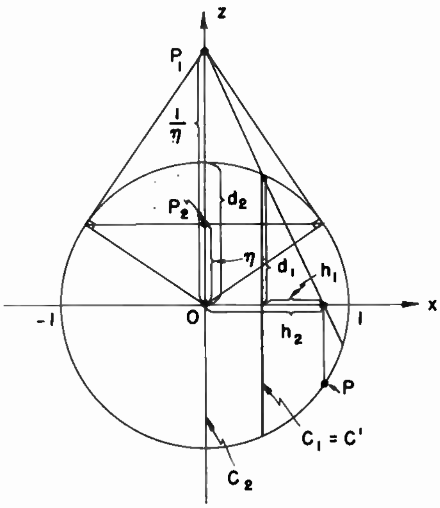

Fig. 3—Graphical determination of the efficiency $\eta$.

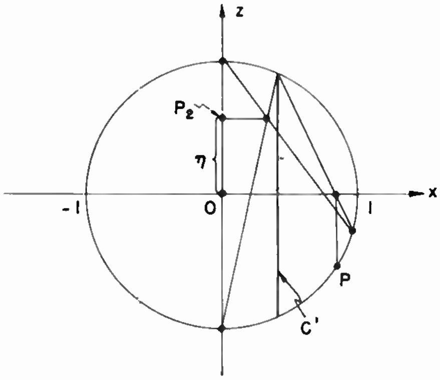

Fig. 4—Graphical determination of the efficiency $\eta$.

It is interesting to check that if the point $P$ is situated in $(1, 0)$, $\eta = \eta_{max}$; this is evident from Fig. 5. $P_2$ is simply the crossover point between the $z$ axis and a line joining $(-1, 0)$ and the upper point of $C'$. Therefore

$$\eta = \eta_{max} = \frac{h_1}{d_1}. \tag{2}$$

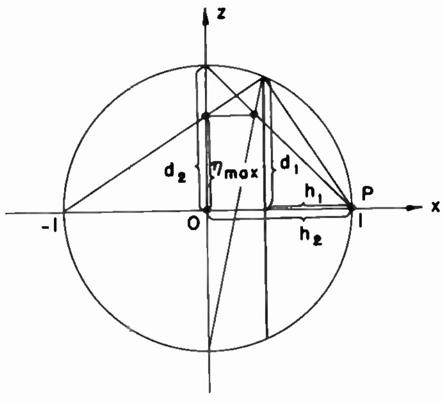

Fig. 5—Graphical determination of the maximum efficiency $\eta_{max}$.

The $yz$ plane contains the complex reflection coefficient plane. Fig. 5 reveals that $\eta_{max}$ is the radius of a circle which is $C'$ stereographically mapped on the $yz$ plane from the point $(-1, 0)$. This checks nicely with Wheeler and Dettinger.[4]

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This work was supported in part by the Army (Signal Corps), the Air Force (Office of Scientific Research, Air Research and Development Command), and the Navy (Office of Naval Research).

[1] E. F. Bolinder, "Maximum efficiency of four-terminal networks," Proc. IRE, vol. 44, p. 941; July, 1956.

[2] H. M. Altschuler, "Maximum efficiency of four-terminal networks," Proc. IRE, vol. 43, p. 1016; August, 1955.

[3] A. Weissfloch, "Die wirkleistungsverluste in linearen vierpolen in abhängigkeit vom wert des transformierten scheinwiderstandes," *E.N.T.*, vol. 19, pp. 259–265; December, 1942.

[4] H. A. Wheeler and D. Dettinger, "Measuring the Efficiency of a Superheterodyne Converter by the Input Impedance Circle Diagram," Wheeler Monograph No. 9, March, 1949.

[5] G. A. Deschamps, "Geometric viewpoints in the representation of wave guides and waveguide junctions," *Proc. Symposium on Modern Network Synthesis*, Polytech. Inst. of Bklyn., pp. 277–295; April, 1952.

## TV Sweep Generation with Resonant Networks and Lines*

The principal object of this communication is to extend the discussion in the above article by Schlesinger.[1] At the same time we wish to direct the attention of the reader to a correction of the appendix of the same article.

Eq. (14) of the appendix of Schlesinger's article gives as the error squared

$$G = \int_{\tau/2}^{H-(\tau/2)} \left[ f(t) - \sum_{n=1}^{N} a_n \sin (n\omega t) \right]^2 dt. \quad (14)$$

$G$ becomes an "extremum," if all the $\delta G/\delta a_k$ vanish where $k$ runs from 1 to $N$. Recognizing that $n$ is a dummy index of summation in (14) we obtain

$$\partial G/\partial a_k = \int_{\tau/2}^{H-(\tau/2)} 2 \left[ f(t) - \sum_{n=1}^{N} a_n \sin (n\omega t) \right] \cdot \sin (k\omega t) dt$$

which results in a system of simultaneous equations of the form

$$\sum_{-1}^{N} A_{kn} a_n = \int_{\tau/2}^{H-(\tau/2)} f(t) \sin (k\omega t) dt, \quad (k = 1, 2, \cdots N)$$

where

$$A_{kn} = \int_{\tau/2}^{H-(\tau/2)} \sin (n\omega t) \sin (k\omega t) dt, \quad (k, n = 1, 2, \cdots N).$$

Since the functions $\sin (k\omega t)$ are not orthogonal over the interval of integration, the matrix $A_{kn}$ is not a diagonal one and Schlesinger's (15) and (16) and Tables I and II are incorrect. The error in (15) is further illustrated by the example $f(t) = \sin \omega t$, and by noting that in Fig. 3 of Schlesinger's article the modified approximation has a larger error over the reduced interval than that of the normal approximation. (At 20° the respective errors are 40 per cent and 22 per cent.)

Referring to the resonant type of sweep Dr. Schlesinger concludes that "... it can meet but not beat the performance standards set by the power feedback type of circuits now in use." By paying attention to the details of design a relatively simple resonant type of circuit will yield greatly improved performance from the standpoint of circuit efficiency without any sacrifice of retrace or linearity characteristics.

I have had a resonant type of sweep in operation since 1950. The circuit shown in Fig. 1 employs a 6AU5GT driver and a 6W4GT diode to deflect a 70° tube at 15 kv anode voltage. The total input to the circuit is 30 ma at 360 v; *i.e.*, an input power of 11 watts; in addition to the sweep, screen supply and high voltage are obtained as outputs. This high efficiency is obtained by the addition of two inductances and capacitors to the conventional circuit. It may be mentioned that the 6AU5GT has seen about 6000 hours of service.

* Received by the IRE, July 3, 1956.
[1] K. Schlesinger, PROC. IRE, vol. 44, pp. 768–775; June, 1956.

It does not seem that ac-dc conversion efficiency should be defined as the ratio of $(\frac{1}{2}LI^2f)$ over power input, because the expression $(\frac{1}{2}LI^2)$ is merely the peak energy stored in the yoke, whereas the true value of the circulated volt-amperes is the summation of the products of the frequency components of yoke voltage and current; this will be approximately $(\frac{2}{3}LI^2f)$. Of course, the actual power dissipated in the yoke depends upon the yoke losses at the various harmonics of sweep frequency. In the conventional circuit the reactive power is handled by the driver and diode; the tube losses will be roughly proportional to the square of the current components of the yoke, and it may be observed that 72.5 per cent of this loss is included in the fundamental and 89 per cent in the fundamental and second harmonic. In an efficient resonant type of sweep circuit, employing resonance at fundamental and second harmonic, the tube losses at these frequencies may be eliminated, so that if the higher harmonic components of current *are not seriously increased*, the tube dissipation may be reduced to 11 per cent of its normal value. The principal losses will then be those in the yoke, the transformer, and the reactive elements which are added to the circuit.

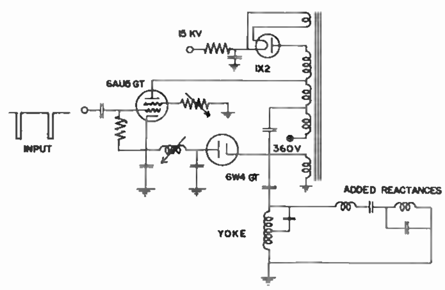


Fig. 1—Practical resonant sweep deflection circuit.

While the proper admittance may be realized in the form of a (nonuniform) transmission line which resonates the yoke at the sweep frequency and many of its harmonics, the proper admittance function should be determined from the condition that the higher order harmonic power not be increased.

The determination of the admittance function will be explained by reference to Fig. 2. Fig. 2(b) and 2(c) shows the yoke voltage and current. Vacuum tubes are employed to control the yoke voltage during the trace portion of the cycle; during the retrace portion, the vacuum tubes are open circuits. Fig. 2(d) and 2(e) shows the currents which would flow in $L_1$ and $L_2$ if they are respectively series tuned to frequencies equal to the reciprocal of the trace period and twice the reciprocal of the trace period. The magnitudes of $I_1$ and $I_2$ may be controlled by the magnitudes of $L_1$ and $L_2$, so that with a proper choice (based upon Fourier's series) the sum current is a minimum in rms value and is shown in Fig. 2(f). The vacuum tube current is shown in Fig. 2(g).

For the details of design we refer again to Fig. 2. Let the sweep period be $2\pi$ and the retrace period be $2b$ so that the trace period is $2(\pi - b)$. We shall write the yoke voltage as

$$E = \cos \omega t \qquad -b < t < b$$
$$E = \cos \omega b \qquad b < t < 2\pi - b. \quad (1)$$

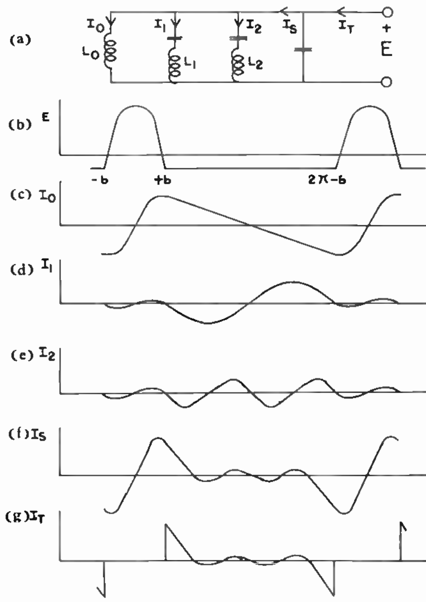


Fig. 2—Determination of correct admittance function.

Since the integral of the yoke voltage over one full period must vanish we have the relation

$$\tan \omega b = -\omega(\pi - b). \quad (2)$$

If the yoke voltage is impressed across a unit inductance in series with a capacitor equal to $(1/\alpha)^2$, the current in the inductance is a solution of the differential equation

$$(p^2 + \alpha^2)I = pE \quad (3)$$

where $p$ denotes differentiation with respect to time. Because of the differential equation and the continuity of the impressed voltage the current and its first derivative are both continuous. We choose $\alpha$ so that

$$\alpha(\pi - b) = k\pi \quad (4)$$

where $k$ is an integer. As a result of (3) and (4) and the conditions of continuity, the current is

$$I = \frac{\omega}{\omega^2 - \alpha^2} \left[ \sin \omega t - \frac{\sin \omega b}{\sin \alpha b} \sin \alpha t \right] \qquad -b < t < b$$
$$I = A \sin \alpha(t - \pi) \qquad b < t < 2\pi - b \quad (5)$$

where

$$(-1)A\alpha = \frac{\omega}{\omega^2 - \alpha^2} [\omega \cos \omega b - \alpha \sin \omega b \cot \alpha b]. \quad (6)$$

Reference to (2) permits one to write

$$(-1)^k A = \frac{\omega^2}{\omega^2 - \alpha^2} (1 + k\pi \cot \alpha b) \cos \omega b. \quad (7)$$

The yoke current in a unit inductance during the trace period is $(\pi - t) \cos \omega b$, and since $A$ is the magnitude of the current in $L_k$, the proper value of $L_k$ is

$$L_k = \frac{\omega^2}{2(\omega^2 - \alpha^2)}(1 + k\pi \cot \alpha b)L_0,$$

$$\alpha = \frac{k\pi}{\pi - b} \tag{8}$$

where $L_0$ is the yoke inductance. The capacitances are determined from

$$C_k = 1/\alpha^2 L_k. \tag{9}$$

By elementary synthesis other equivalent circuit configurations may now be derived which have a more convenient practical form.

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## *Author's Comment*[2]

In this author's opinion, Mr. Wagner's letter does not invalidate any of the statements made in the paper.

1) The set of coefficients derived by (16) in the appendix of the paper, is plotted graphically in curve (b) of Fig. 3. This curve approaches the ideal sawtooth wave 3 times better in the sweep-interval ($40° \leqq \phi \leqq 320°$), than does the conventional Fourier series, graph (a) in Fig. 3. The interval $-40° < \phi < +40°$, where curve (b) is twice as much in error as curve (a), is not sweeptime, but flyback-time. This is so stated on page 770, lines 5–8. of the paper. By citing this statement in reverse, as he does, Mr. Wagner makes a supporting evidence look like a discrepancy.

2) The expression $\frac{1}{2}LI^2$ for circulating energy, as used in the paper, is a logical choice, since it can be measured readily and is independent of waveform. Both advantages are lost if this expression is replaced by the sum of sideband energies, as suggested by Wagner. Besides, the end result of his summation is in error by an order of magnitude.

3) The circuit shown in Wagner's Fig. 1 does not belong in the group of circuits for "synthetic sweep," as discussed in the paper. Rather, his is a conventional power feedback circuit, with the "booster diode" shifted from the plate return to the cathode return.

In operation, Wagner's circuit will: a) stop output as soon as trigger pulses stop; b) deliver an approximate sawtooth-output, even with the reactance network disconnected.

Both performance features are the opposite of the typical behavior of sweep forming networks, as shown in Figs. 6 and 12 of the paper. None of the circuits shown there uses power feedback. Note that power-recovery by a booster-diode, and power-storage and recirculation in a resonant network, are two basically different approaches to the problem. The first is inherently a "one-shot" system; the second is a "ringing" system. Both methods are mutually exclusive!

Wagner's circuit is a hybrid between the two, trying to combine the advantages of both. However, it has been the writers experience that the introduction of power-feedback into resonant sweep causes the network $Q$ to drop at a rate which largely offsets the booster gain. Conversely, the addition of a sweep-forming network to a well-designed power feedback circuit may do more harm than good, since the network only adds its own losses to the system, without the returns from resonance.

A glance at Wagner's Fig. 2 confirms the suspicion that his network is heavily damped. His branch-currents $I_1$ and $I_2$ have lost all resemblance to pure harmonics. It requires some courage to try to analyze that situation!

Of course, this test cannot be used to prove or disprove my analysis of synthetic sweep, since there is no basis for comparison.

Specifically, I can see why the retrace in Wagner's circuit is as fast as that of a conventional switching circuit, and not as slow as would follow from my Fig. 2, for a two-pole resonant network. His network cannot "flywheel," since it is gated after each cycle.

Mr. Wagner's contention, that only two harmonics are enough to build a good sawtooth, is without any foundation in theory or practice. This may happen with distorted current components as shown in his Fig. 2. Such complex waveforms are, indeed, unusually accommodating at times.

But if Wagner ever tried to synthetize a saw from pure sinewaves, and had only the first two terms to work with, then I am sure that he would soon find himself in complete agreement with all statements made in my paper.

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## *Letter from Mr. White*[3]

The recent paper by Schlesinger[1] on resonant networks for sweep circuits was of especial interest inasmuch as a somewhat similar investigation some years ago indicated that remarkably good sweep waveforms could be synthesized from a finite number of harmonic components. Although this study never reached the experimental stage due to practical difficulties associated with the problem at hand, the results are presented here in the hope that they may be of interest to others wishing to pursue the subject further.

The modified Fourier series used by Schlesinger exhibits a kind of Gibbs' phenomenon which is typical of applications where a discontinuous function is approximated by a finite number of Fourier components. For example, it may be noted that the waveform of Fig. 3 of Schlesinger's paper has a peak slope variation of about 50 per cent whereas when the number of harmonics is increased from 4 to 8 as in Fig. 4 of Schlesinger's paper, the peak slope error is reduced only to about 38 per cent. To one whose primary interest is the minimizing of slope errors, these figures may be discouraging. As will be shown, however, using a different technique to determine the harmonic amplitudes permits one to synthesize sweep waveforms with the slope errors held to a specified tolerance.

Consider the time function

$$f(t) = T_{2r}\left[\frac{\cos\dfrac{\omega t}{2}}{\cos\dfrac{\omega\tau}{2}}\right]$$

where

$T_n(x)$ is the $n$th order Tchebycheff polynomial
$\tau$ = constant = $\frac{1}{2}$ the allowed total retrace time
$\omega = 2\pi \times$ the desired sweep repetition rate.

It is apparent that $f(t)$ is a polynomial in even powers of $\cos \omega t/2$ and that if we make use of the relation

$$\left\{\cos\left(\frac{\omega t}{2}\right)\right\}^{2s} = \frac{1}{2^{2s}}\left[\frac{(2s)!}{(s!)^2} + 2\sum_{k=1}^{s}\frac{(2s)!}{(s+k)!(s-k)!}\cos(k\omega t)\right]$$

then $f(t)$ can be expressed as a harmonic series in the form

$$f(t) = \bar{f} + \sum_{k=1}^{r} f_k \cos(k\omega t).$$

From the properties of the Tchebycheff polynomial, we may also deduce that

$$|f(t)| \leqq 1 \text{ when } \left|\cos\left(\frac{\omega t}{2}\right)\right| \leqq \cos\left(\frac{\omega\tau}{2}\right)$$

and

$$|f(t)| \geqq 1 \text{ when } \left|\cos\left(\frac{\omega t}{2}\right)\right| \geqq \cos\left(\frac{\omega\tau}{2}\right).$$

These properties are illustrated in Fig. 1

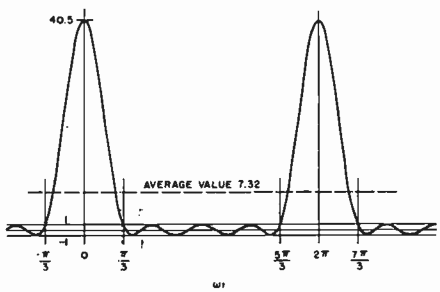


Fig. 1—The function $f(t)$: 4 harmonics, $\omega\tau = \pi/3$.

which is a plot of $f(t)$ for the case where $r = 4$ and $\omega\tau = \pi/3$.

Suppose now that we form the function

$$g(t) = \int_0^t \left\{\frac{f(t) - \bar{f}}{\bar{f}}\right\} d(\omega t) = \sum_1^r \frac{f_k \sin(k\omega t)}{k\bar{f}}$$

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$g(t)$ will be characterized by a steeply rising slope during the intervals $-\tau < t < \tau$ and an approximately uniform declining slope during the interval $\tau < t < (2\pi/\omega) - \tau$. The limits of variation of this slope are given by the relation

$$1 - \frac{1}{\bar{f}} \leq - \left\{ \frac{d}{d(\omega t)} g(t) \right\} \leq 1 + \frac{1}{\bar{f}}$$

or, in other words, the fractional variation in slope is $1/\bar{f}$. In Fig. 2, we have plotted both

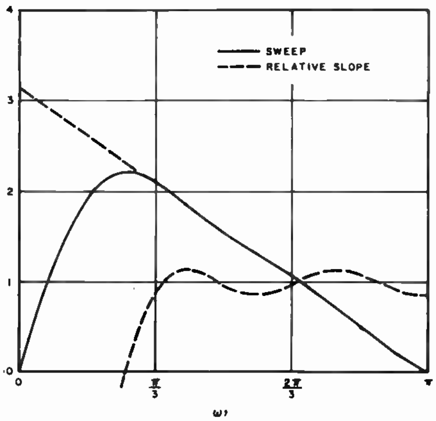


Fig. 2—The function $g(t)$ and its relative slope: 4 harmonics, $\omega\tau = \pi/3$.

$g(t)$ and its slope function for the four harmonic case where $\omega\tau = \pi/3$. The relation between maximum slope error, allowable retrace time, and the number of harmonics required is shown in Fig. 3. This is a plot of $1/\bar{f}$ vs $\omega\tau$ for the cases of 2, 4, 6, or 8 harmonics.

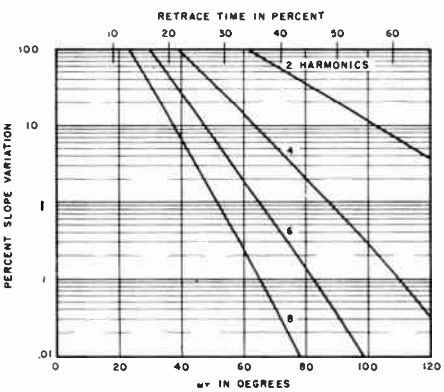


Fig. 3—Slope variation as a function of retrace time.

Fig. 4 shows the slope error for two cases which are roughly equivalent to the ones computed by Schlesinger. Curve no. 1 is the slope error for 4 harmonics and $\omega\tau = 53.3°$. It will be noted that the peak slope error is 25 per cent and that the duty cycle between waveform peaks, *i.e.*, between points where the slope is zero, is 77 per cent. This corresponds with Schlesinger's curve which gives about 50 per cent slope error over the same duty cycle. Curve no. 2 of the same figure is the slope error for 8 harmonics and $\omega\tau = 37°$. This curve gives a maximum slope

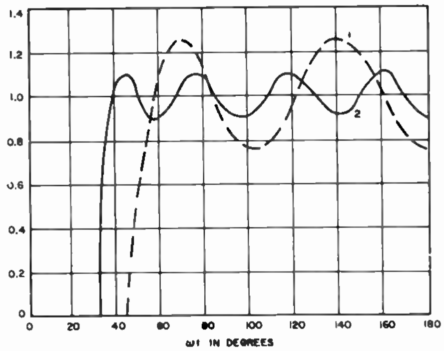


Fig. 4—Slope variation curves: curve 1, 4 harmonics, $\omega\tau = 53.3°$; curve 2, 8 harmonics, $\omega\tau = 37°$.

error of 10 per cent and a duty cycle between waveform peaks of about 83 per cent. This corresponds to Schlesinger's result for 8 harmonics which gave 38 per cent slope error with the same duty cycle.

From the oscilloscope photographs, it is apparent that Schlesinger in his experimental tests was obtaining better performance than his theory indicated. It is probable that the actual harmonic amplitudes were closer to the Tchebycheff values than to the Fourier values.

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### *Author's Comment*[4]

The correspondence by Mr. White is a valuable contribution to the problem of synthetic sweep. It is of particular interest that there seems to be general agreement between the Tchebycheff and the Fourier approach in evaluating retrace speed as a function of bandwidth. This point, as well as power efficiency, has been more of a problem, in our experimental work, than sweep linearity, which was generally satisfactory.

There may be room for two additional comments:

1) In the display of stationary television pictures, it appears that positional errors are more readily detected than slope errors of equal magnitude. Assuming, for example, a sinusoidal velocity modulation of $\pm\rho_v$ per cent with $N$ cycles per scan. The resulting positional error is $\rho_x = \rho_v/2\pi N$, where $\rho_x = \Delta x/W$, and $W$ the picture width. Hence, short ripples on the slope-curve may pass unnoticed in the display. This seems to be borne out by the photographs 7(b), 13(a), and 15(b) of the paper where small oscillations, discernible on the yoke-voltage wave, do not seem to appear on the sweep wave. For the same reason, the improvement in sweep linearity between Figs. 3(a) and 4 is better than the slope error reduction would indicate, because the "wavelength" of the slope function 4(b) has decreased over Fig. 3 in the process of doubling $N$.

2) In comparing theory with experience, it may be worth mentioning that all practical circuits for resonant sweep, as reported in the paper, used shock excited harmonics, rather than continuous waves. As a result, the amplitude of the $K$th harmonic decreased slightly during each scan, since $K-1$ cycles are ringing without external power supply. In the network of Fig. 6, for instance, a loss of 1.3 db per trace was observed for the 6th harmonic.

As a result, the spectral distribution varied slightly during a line scan, with a resultant small nonsymmetry discernible in Figs. 7 and 13. This is in violation of any theory dealing with synthesis from cw components. Hence, in order to study the merit of different spectra more accurately, it may be of interest to repeat the above tests with equipment using $N$ oscillators locked to line-sync, as sources for harmonic energy.

K. Schlesinger

[4] Received by the IRE, October 5, 1956.

## Simplification of Field Strength Computations for Shielded Enclosures*

The method developed by Haber[1] has made possible a considerable increase in the accuracy of sensitivity measurements on aircraft adf loop antennas in screen rooms, but the computations involved in arriving at the field strength become quite tedious. The following derivation results in a simplified formula allowing direct computation of the relation between the voltage output of the signal generator used to feed the transmission line, and the field strength in volts per meter at the loop. This relation, having the dimension of length, and usually referred to as the "attenuation constant" or "room factor" ($K_d$), is particularly useful in making sensitivity measurements as required by RTCA specifications, where acceptance criteria are based on the field strength necessary to obtain the desired performance.

Haber's original equation is

$$H_{\text{total}} = \frac{I}{w}\left[\frac{\sinh\alpha}{\cosh\beta - \cosh\alpha} - \frac{\sinh 2\alpha}{\cosh 2\beta - \cosh 2\alpha} + 4\sinh\alpha\cosh\beta\exp(-2\gamma)\right]. \quad (1)$$

This expression may be transformed by applying standard hyperbolic identities into

$$H_{\text{total}} = \frac{I}{w}\sinh\alpha\cosh\beta$$

* Received by the IRE, August 3, 1956; revised manuscript received, October 15, 1956.

[1] F. Haber, "Generation of standard fields in shielded enclosures," Proc. IRE, vol. 42, pp. 1693–1698; November, 1954.

$$\cdot\left[\frac{1}{(\cosh\beta+\cosh\alpha)(\cosh\beta-\cosh\alpha)}+4\exp(-2\gamma)\right]. \qquad (2)$$

This reduction gives an expression requiring the use of only three hyperbolic functions instead of six, with a quite appreciable saving in time and tempers when several computations are needed in quick succession.

As defined above, $K_d=E_L/E_{v/m}$, where $E_L$ is the signal generator output voltage, and $E_{v/m}$ is the field strength at the loop in volts per meter. Since $E_{v/m}=H_{\text{total}}Z_0$, and $E_L=IZ_L$, $Z_0$ being $120\pi$ ohms, the impedance of free space, and $Z_L$ the line terminating resistance, the final expression for $K_d$ becomes

$$\frac{1}{K_d}=\frac{14842\sinh\alpha\cosh\beta}{Z_{Lw}}\cdot\left[\frac{1}{(\cosh\beta-\cosh\alpha)(\cosh\beta+\cosh\alpha)}+4\exp(-2\gamma)\right]. \qquad (3)$$

The constant includes the conversion factor necessary to permit the use of the enclosure dimensions in inches.

This expression has been found convenient in the computation of room factors for a variety of screen rooms, and the sensitivity measurements obtained have shown much better agreement than those obtained with factors computed from older formulas.

It should also be noted that when operating at the low and medium adf frequencies (below 2 mc) in a screen room of average size (10–15 feet long), the transmission line length is small compared to the wavelength, and a considerable mismatch can be tolerated without seriously disturbing the field. Under these circumstances, adjustment of the room factor to a convenient value can often be made by changing the terminating resistance ($Z_L$) on the line without changing any other dimensions.

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## A Comparison of Two Radiometer Circuits*

*Letter from Mr. Tucker*

I feel that the above paper,[1] although interesting and valuable, fails to discuss several important points which are essential to a full understanding.

1) On examining the working for the Dicke system it becomes clear that the following assumptions are made which are not stated in the list of assumptions nor in the conclusion, although they are vital to the interpretation of the results:

$$\left.\begin{array}{l}\alpha\gg\beta\\ \alpha\gg q\end{array}\right\}\text{explicit assumptions}$$

$\beta\ll q$: an implicit assumption, left to the reader to puzzle out for himself.

The last assumption is implied by the author's approximation

$$P_2=[G_{13}(q)+G_{14}(q)]\beta.$$

In the working for the two-receiver system, the assumption that $\gamma\ll\alpha$ is implicit, but is not stated.

These assumptions are made only for the simplification of the analysis, and are not necessary for the physical realization of the systems, which would work perfectly well with a nearly sinusoidal input and $\alpha\ll\beta$, $\alpha\ll\gamma$, $q\ll\beta$. The conditions which *are* necessary for successful realization are not stated, but I think that in the Dicke system one is $\gamma<2q$, and in the two-receiver system one is $\gamma<2(f_1-\alpha)$. The two-receiver system is actually easily analyzed without any restriction other than $\gamma<2(f_1-\alpha)$,[2] and a recent paper of mine[3] shows that when there is no low-pass filter other than the one required to eliminate the sum frequencies from $2(f_1-\alpha)$ upwards, then the output signal /noise power ratio is

$$\left(\frac{s}{n}\right)_0=2\frac{\sigma_s^4}{\sigma_n^4}$$

in the author's terminology. The effect of the low-pass filter is easily allowed for, remembering that the output spectrum is triangular; thus there is no effect at all until $\gamma$ is reduced below $\alpha$, and then

$$\left(\frac{s}{n}\right)_0=2\frac{\sigma_s^4}{\sigma_n^4}\cdot\frac{\alpha^2}{\gamma(2\alpha-\gamma)}$$

and if $\gamma\ll\alpha$, this reduces to

$$\left(\frac{s}{n}\right)_0=\frac{\sigma_s^4}{\sigma_n^4}\cdot\frac{\alpha}{\gamma}$$

as obtained by the author.

2) In view of the approximations made, the statement of the improvement of the two-receiver system over the Dicke system to three significant figures (*i.e.*, 3.48) needs qualification.

3) The paper is concerned only with background noise arising in the receiving amplifier. If the noise arises (as in many practical applications, especially when the radiation being measured is acoustic and not electromagnetic) in the medium of propagation or in the receiver (*e.g.*, antenna) before the modulator, then the analysis of the two-receiver system is still valid; but that of the Dicke system is *not* valid, since this system was intended only to overcome amplifier noise.

4) It is concluded that the two-receiver system can detect a signal at a level 3.5 times lower than the Dicke system can. But, surely, if the two available receivers (*e.g.*, antennas) were merely joined in parallel and then applied to the Dicke system in place of the one receiver assumed in the paper, then the Dicke system would be improved by the doubling of signal power, and would be inferior to the other system by a factor of only 1.7.

5) It is not mentioned that the signals at the two receivers (*e.g.*, antennas) are fully correlated only if the line joining the two receivers is parallel to the signal wavefront. In typical applications there is smooth and continuous relative movement between the source of radiation (whether terrestrial or extraterrestrial) and the receiving system, as produced, for example, by the rotation of the earth, carrying the receivers, relative to a source of radiation such as a radio star. In fact such movement is usually essential to efficient detection of the signal. In such a case, the signal inputs of the two-receiver system are fully correlated only at one position, and the "dc" output, which traces the cross-correlation function of the two receiver inputs, is actually quasi-sinusoidal at a very low frequency of say $\gamma_0$ over a number of cycles which depends on the bandwidth of the receivers. It is clear, then, that there is a fairly definite lower limit to the low-pass cutoff frequency; when the filter has a rectangular band-response this limit is clearly $\gamma_0$ approximately, and this cutoff gives the lowest attainable value of just-detectable signal. This limitation does not apply to the single-receiver system, where the lowest value of the cutoff frequency ($\gamma$) is limited only by the time during which the signal is being received at all.

6) The assumption of a rectangular low-pass filter response is rather unrealistic, as in many practical cases the integration time required is very large, and RC-type networks have to be used. Using the results of a calculation by Griffiths[4] on the optimum use of RC networks for separating signal from noise, and assuming a quasi-sinusoidal output signal as described above, I have calculated the effect of replacing the filter in the two-receiver system by $N$ simple resistance-capacitance circuits in independent cascade (*i.e.*, separated from one another by an amplifier or attenuator); and on the assumption that the frequency of 3-db attenuation ($\gamma_1$) of a single RC section is very small compared with $\alpha$, the optimum condition for detection is

$$\frac{\gamma_1}{\gamma_0}=\sqrt{2N-1}.$$

Without this condition, the output signal/noise power ratio is

* Received by the IRE, January 9, 1956.

[1] S. J. Goldstein, Jr., Proc. IRE, vol. 43, pp. 1663–1666; November, 1955.

[2] At this stage the reader should be reminded that the two-receiver system merely multiplies together the two input signals and then smooths the product with a low-pass filter. The input bandwidth for signal and noise is $\alpha$, the center frequency of the band is $f_1$ and the cutoff frequency of the low-pass filter is $\gamma$. The input signal amplitude is $\sigma_s$ and the input rms noise amplitude is $\sigma_n$.

[3] D. G. Tucker, "Signal/Noise Performance of Multiplier (or Correlation) and Addition (or Integrating) Types of Detector," IEE Monograph No. 120R, London, England; 1955.

[4] J. W. R. Griffiths, "Optimum rc filters for separating sinusoidal signal from noise," *Wireless Engr.*, vol. 33, pp. 268–270; November, 1956.

$$s = \sigma + j\omega$$

$$\text{complex spinvel} = \text{logvel} + j(\text{spinvel}),$$

where "log velocity" is contracted to "logvel." The unit for "logvel" is nepers per second, the unit for "spinvel" is radians per second, and the unit for "complex spinvel" is the ideal unit "nerad," introduced by Buss.

In doing away with the standards we live by, is it not just as important that we consider "the thing" that spins around, with reference to $\omega$ in ac theory; voltages, currents, and charges. More plainly, let us *once and for all* do away with the spinning *vector* in ac theory, in favor of the *phasor* and its ill-treated step-child, the *sinor*. How can sound-thinking engineers continue year after year to call scalar quantities vectors? What about the hoax that $j\omega$ is an algebraic quantity (which would mean that it is the sum of $j$ pieces, each one being $\omega$), or the still bigger hoax that by rigorous mathematical thinking we obtain $-1$ because we can take the square (meaning $j^2$) of a square root, when in reality there is no square, since $j^2$ does not mean "$j$ squared" but merely indicates that the operation has been performed twice! As a world authority in this field, Emeritus Prof. J. Hollingworth of the University of Manchester, England, puts it: "It would have saved a lot of trouble if it had been written $j_2$" ("it" standing for $j^2$).[3] Adding more "what abouts," how come that "instantaneous complex quantities" still do not rate symbols of their own, such as $\boldsymbol{I}_t$ for $Ie^{j\omega t} = Ie^{j(\omega t+\alpha)}$? Returning to $s=\sigma+j\omega$, $\omega$ is called "real frequency," and still worse, $\sigma$ is sometimes called "imaginary frequency." In the first place, neither one is a frequency, and $\sigma$ is just as real as $\omega$, so "real frequency" fails to provide the proper distinction, and if anything should be associated with "real," it seems that it should be the first quantity $\sigma$ in $s=\sigma+j\omega$, not the second quantity $\omega$. The worst misnomer of them all, in this writer's opinion, is, however, *imaginary*, used by Buss and most everybody else. This unfortunate term, introduced some three hundred years ago by René Descartes (1596–1650), seems slated to becloud our minds for all future, giving young students the idea that they are playing around with fictitious things from an "imaginary world." This can be at least partly avoided in straight-forward presentation of $s=\sigma+j\omega$ if we say that $\sigma$ is the real (or reference, or prime-direction) part, that $j\omega$ is the $j$ term, and that $\omega$ is the $j$ part, or, if we so wish, the quadrature part.[4] This terminology, in which "imaginary axis" is replaced by "$j$ axis," also eliminates the controversy of $\omega=\text{Im}(s)$, the real $\omega$ being the imaginary part of something, and substitutes the well-known writing $\omega=Q(s)$, where $Q$ is inspired by "quadrature."

Many difficulties, such as the ones mentioned above, are still thrown in the path of the young electrical engineering student, and this fact is mentioned here as a serious suggestion that if we cleaned up the fundamentals, perhaps the pieces would all fit together when we would come to higher learning, so that this discussion, ably started by Buss, would not be necessary.

Harry Stockman
Neutronics Research Co.
Waltham, Mass.

[3] J. Hollingworth, "The symbolic method," *Bull. Elec. Engr. Education, England*, no. 10, pp. 22–23; June, 1953.
[4] H. Stockman, "On reciprocal inductance," Proc. IRE, vol. 43, p. 341; March, 1955.

## Information Rate of a Human Channel*

The rate at which a human being can respond to and reproduce information is important in communication and in properly relating machines to human beings. Previous studies[1,2] have indicated that reading aloud randomized lists of words leads to a higher information rate than, for instance, typing, playing randomized sequences of notes on a piano, or tracking (pointing). Licklider,[1] using words chosen randomly from a dictionary, found a higher information rate for tracking and reading simultaneously[3] than for reading alone.

We obtain our highest rates, 42–43 bits/second, by using a "code" consisting either of the 2500 most common monosyllables[4] (11.3 bits/word), or prose scrambled or randomized over long stretches (11.82 bits/word[5]). These rates are as high as those which Licklider obtained for reading and tracking simultaneously. With our code, reading and tracking simultaneously gives a lower information rate.

We find that reading rates vary among individuals and groups, being low (~2 words/second) among porters and high (~3 words/second) among engineers. For a given individual, reading rate varies with familiarity (among which thousandth in order of frequency of usage the words are chosen) and with number of syllables, as shown in Fig. 1. Nonsense words are at first read slowly, but with increased familiarity the rate approaches that for meaningful words.

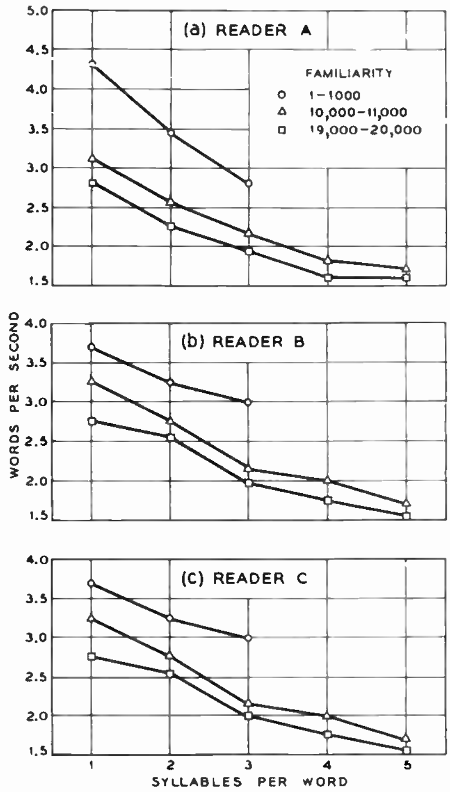


Fig. 1

Experiments indicate that the limitation on reading rate is a very fundamental one, mental rather than muscular (limitation on rate of utterance), as illustrated by the fact that a phrase can be repeated or prose read faster than randomized lists can be read (Fig. 2). Neither does the limitation seem to be one associated with our mode of writing, for two bilingual Chinese readers read randomized lists of common Chinese characters at almost the same rate at which they read randomized lists of the corresponding English words.

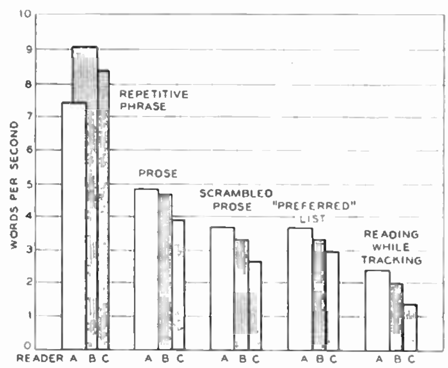


Fig. 2

In all, we have found a lower bound of some 43 bits/second to the rate at which information can be transmitted through a human channel. This rate can be reached by reading aloud randomized lists of the 2500 most familiar monosyllables. Longer or shorter lists (more or less complex tasks) give lower rates. The limitation on the rate appears to be mental rather than muscular. No higher rate has been achieved through multiple tasks. Because the limitation appears to be a mental one, it may imply that information rates of this order could provide satisfactory sensory inputs in sound or picture transmission. We plan to publish our results more comprehensively in an early issue of the *Bell System Technical Journal*.

J. R. Pierce and J. E. Karlin
Bell Telephone Labs., Inc.
Murray Hill, N. J.

* Received by the IRE, October 15, 1956.
[1] J. C. R. Licklider, K. N. Stevens, and J. R. M. Hayes, "Studies in Speech, Hearing and Communication," Mass. Inst. Tech., Acoustics Lab. Tech. Rep.; September 30, 1954.
[2] H. Quastler, *et al.*, "Human Performance in Information Transmission," University of Illinois, Control Systems Lab. Rep. no. R-62; March, 1955.
[3] In Licklider's case, "tracking" means touching the interior of a box drawn beside the word; in our case it means putting a dot close to a short vertical line to the right of a word.
[4] E. L. Thorndike, "A Teacher's Word Book of the Twenty Thousand Words Found Most Frequently . . . ," Columbia University Teachers College; 1932.
[5] C. E. Shannon, "Prediction and entropy of printed English," *Bell Sys. Tech. J.*, vol. 30, pp. 50–64; January, 1951.

# Contributors

Robert D. Hatcher was born at Watervliet Arsenal, N. Y. in 1914. He was educated at Virginia Polytechnic Institute, U.C.L.A., and Princeton. He has worked in radio since 1926.

R. D. HATCHER

Mr. Hatcher taught physics and electronics at the Naval Academy from 1944 to 1948. Since 1948 he has been a member of the staff of the National Bureau of Standards and the Diamond Ordnance Fuze Laboratories. He is Chief of the Microwave Development Section at DOFL.

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Robert A. Pucel (S'48–A'52–S'54–M'56) was born on December 27, 1926, in Ely, Minn. After serving in the U. S. Navy from 1945 to 1946, he entered the Massachusetts Institute of Technology and received the B.S. and M.S. degrees in electrical engineering in 1951. From 1948 to 1951 he was enrolled in a cooperative plan and worked for the General Electric Company.

R. A. PUCEL

Following his graduation in 1951, Dr. Pucel joined the Research Division of the Raytheon Manufacturing Company, Waltham, Mass., where he was a member of the Microwave Noise Study Group. In 1952, he returned to M.I.T. as a doctoral candidate, and was also a staff member of the Research Laboratory of Electronics. He received the degree of Doctor of Science in electrical engineering in 1955 from M.I.T.

In 1955, he rejoined the Research Division at Raytheon as a staff member of the Theoretical Physics Group, where he is principally engaged in the study and application of new solid-state device concepts.

Dr. Pucel is a member of Sigma Xi.

❖

Jan A. Rajchman (SM'46–F'53) was born in London, England, on August 10, 1911. He received his diploma in electrical engineering in 1934 and the degree of Doctor in technical sciences in 1938 from the Swiss Institute of Technology, Zurich, Switzerland. He started in the Summer of 1935 as a student engineer at RCA Manufacturing Co. in Camden. In 1936 he joined the staff of RCA Manufacturing Co. as a research engineer and in 1942 he was transferred to the RCA Laboratories in Princeton where he is a member of the research staff.

J. A. RAJCHMAN

At first Dr. Rajchman worked in electron optics. He is chiefly responsible for the development of the electron multiplier tube. During the war he was among the first to apply electronics to computers. Later he worked on the betatron for which he became a co-recipient of the 1947 Levy Medal of the Franklin Institute. After the war he resumed work on computing devices. He developed the selective electrostatic storage tube. Turning to the new field of magnetics he developed the magnetic core memory, magnetic switching circuits, and more recently the transfluxor. He is presently active in the field of magnetics and other solid state devices.

Dr. Rajchman is a member of the American Physical Society, the Council of the Association for Computing Machinery, and Sigma Xi. He has been granted more than 50 U. S. patents.

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Frank Reggia (A'55) was born in Northumberland, Pennsylvania on October 30, 1921. He attended George Washington University, and the University of Maryland, and is a graduate of Radio Matériel School at the Naval Research Laboratory in Washington, D. C. While a member of the Armed Forces, he served as an electronic specialist in both the United States and in the Pacific Theater. Following separation in 1945, he joined the staff of the Microwave Standards Section of the National Bureau of Standards, where he was engaged in research and development in a microwave standards program

F. REGGIA

Mr Reggia joined the technical staff of the Diamond Ordnance Fuze Laboratories, Department of the Army, Washington, D. C., in 1954 Since then he has been engaged in measurement techniques and applications of ferrites at microwave frequencies. He is a member of the Washington Society of Engineers. He received the Washington Society of Engineers' award for 1953.

❖

William Shockley (SM'51–F'55), one of three recipients of the 1956 Nobel Prize in Physics for his work on semiconductors and discovery of the transistor effect, was born in London, England, on February 13, 1910. He received the B.S. degree in 1932 from California Institute of Technology. Continuing his studies at Massachusetts Institute of Technology on a teaching fellowship, he received the Ph.D. degree in physics in 1936.

W. SHOCKLEY

In September of that year he joined Bell Telephone Laboratories, where his work included vacuum-tube and electron-multiplier design, radar development, solid-state physics, magnetism, and semiconductors.

In 1942, Dr. Shockley was assigned to Columbia University Division of War Research as Director of Research for the Anti-submarine Warfare Operations Research Group. In 1944 he became a consultant to the Office of the Secretary of War.

Returning to Bell Telephone Laboratories in 1945, he became director of transistor physics, and in collaboration with W. H. Brattain and John Bardeen, worked on the development of the transistor.

At present, Dr. Shockley is director of the Shockley Laboratory of Beckman Instruments, Inc., Mountain View, Calif.

He was awarded the IRE Medal for Merit in 1946 and the Morris Liebmann Award in 1952, and in 1951 was elected to the National Academy of Sciences. In June, 1956, Rutgers University awarded him the honorary degree of Doctor of Science. He is a member of the American Physical Society, Tau Beta Pi, and Sigma Xi.

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Edward G. Spencer was born on July 21, 1920, at Lynchburg, Va. He received the B.S.E. degree in physics from George Washington University and the M.A. degree in physics from Boston University. He did further graduate work at Massachusetts Institute of Technology and at the University of Maryland. From 1943 to 1946 he was engaged in microwave radar research at the Naval Research Laboratory.

E. G. SPENCER

From 1946 to 1949 Mr. Spencer was associated with the Cambridge Air Force Research Laboratory, and from 1949 to 1953 with the Naval Research Laboratory. During this time he worked in microwave spectroscopy of gases and paramagnetic and nuclear magnetic resonance of solids. In 1953, he joined the National Bureau of Standards, Ordnance Electronics Division, which is now the Diamond Ordnance Fuze Laboratories. He is at present engaged in microwave physics research.

Mr. Spencer is a member of the American Physical Society.

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Hermann Statz was born in Germany on January 9, 1928. He received the Diploma in Physics in 1949, and the Doktor der Naturwissenschaften degree in 1951, from the Technische Hochschule, Stuttgart. During 1949 to 1951, he was a research associate at the Max Planck Institut für Metallforschung Stuttgart. From 1951 to 1952, he held a research stipend from Deutsche Forschungsgemeinschaft; his work was concerned with problems in theoretical solid state physics. In 1952 he joined the staff of the Solid State and Molecular Theory Group at the Massachusetts Institute of Technology, where he worked on the theory of ferromagnetism. Since 1953 he has been with the Research Division of Raytheon Manufacturing Co., Waltham, Mass., doing research in solid state physics and related fields.

H. Statz

Dr. Statz is a member of the American Physical Society.

Arthur P. Stern (A'51–SM'55) was born on July 20, 1925 in Budapest, Hungary. He studied at the Universities of Budapest and Lausanne and at the Swiss Federal Institute of Technology in Zurich, where he acquired the Master's degree in electrical engineering in 1948.

A. P. Stern

From 1948 to 1951, Mr. Stern did research and development work in the field of gaseous discharges in Switzerland. In 1950, he became instructor for illumination engineering and photometry at the Swiss Federal Institute of Technology.

Mr. Stern came to the United States in 1951 and joined the staff of the General Electric Company's Electronics Laboratory in Syracuse, N. Y. He worked first in the field of color television systems and shifted his activities later to problems of solid state and, in particular, transistor circuitry. At the present time, Mr. Stern heads the Advanced Circuits development section of the Electronics Laboratory.

Mr. Stern is a member of the Scientific Research Society of America, and an associate member of AIEE.

❖

John E. Tompkins was born in Waterloo, Iowa, in 1922. He received the Bachelor's and Master's degrees in physics from George Washington University. During the war he worked in mass spectrometry at Oak Ridge.

Since 1949 Mr. Tompkins has been employed as a microwave physicist at the Bureau of Standards and the Diamond Ordnance Fuze Laboratories.

J. E. Tompkins

Mr. Tompkins is a member of Sigma Pi Sigma, The American Physical Society, The American Association of Physics Teachers, and an associate member of Sigma Xi.

❖

James P. Wittke was born in Westfield, N. J. on April 2, 1928. He received the M.E. degree from Stevens Institute of Technology in 1949 and the M.A. and Ph.D. degrees in physics from Princeton University in 1952 and 1955, respectively.

J. P. Wittke

After a year as instructor at Princeton University, Dr. Wittke joined the technical staff at RCA Laboratories. Since 1955, he has been engaged in an experimental and theoretical study of molecular amplification.

He is a member of the American Physical Society, Sigma Xi, and Tau Beta Pi.

# 1957 IRE National Convention Program

## Waldorf-Astoria Hotel, New York Coliseum March 18–21, New York, N.Y.

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## Convention Highlights

### Technical Program

A schedule of 55 technical sessions appears on the next page, followed by abstracts of the more than 280 papers to be presented.

### Radio Engineering Show

This year's exhibition will be held in a new convenient location, the New York Coliseum at 59th St. and 8th Ave. A list of the 840 exhibitors and their products appears in "Whom and What to See at the Radio Engineering Show" in the advertising section of this issue.

### Annual Meeting

Time: 10:30 **A.M.**, Monday, March 18.
Place: Grand Ballroom, Waldorf-Astoria Hotel.
Speaker: Donald G. Fink, Director of Research of Philco Corporation and Editor of IRE, "Electronics and the IRE-1967."

The special features of this opening meeting of the convention will be of particular interest to all IRE members.

### Annual IRE Banquet

Time: 6:45 P.M., Wednesday, March 21.
Place: Grand Ballroom Waldorf-Astoria Hotel.
Guest Speaker: Dr. John A. Hannah, President, Michigan State University.
Presentation of IRE Awards: John T. Henderson, IRE President.
Spokesman for IRE Fellows: Major General James D. O'Connell, U.S. Army Signal Corps.
Toastmaster: Rear Admiral Charles F. Horne, Convair.

### Cocktail Party

Time: 5:30–7:30 P.M., Monday, March 18.
Place: Grand Ballroom, Waldorf-Astoria Hotel.

### Women's Program

An entertaining program of tours and shows has been arranged for the wives of members. Women's headquarters will be located in the Regency Suite on the fourth floor of the Waldorf.

# SCHEDULE OF TECHNICAL SESSIONS

* Sessions Terminate at 12:00 Noon.

| | WALDORF-ASTORIA | | | | | NEW YORK COLISEUM | | |
|---|---|---|---|---|---|---|---|---|
| | **Starlight Roof** | **Astor Gallery** | **Jade Room** | **Sert Room** | **Grand Ballroom** | **Morse Hall** | **Marconi Hall** | **Faraday Hall** |
| **Monday March 18 2:30 P.M.-5:00 P.M.** | *Session 1*<br>NONLINEAR CONTROL SYSTEMS | *Session 2*<br>VEHICULAR COMMUNICATIONS | *Session 3*<br>PROPAGATION | *Session 4*<br>ULTRASONICS ENGINEERING I—An Educational Session | *Session 5*<br>AERONAUTICAL ELECTRONICS | *Session 6*<br>MULTIPLEX COMMUNICATIONS SYSTEMS | *Session 7*<br>INFORMATION THEORY—Coding and Detection | *Session 8*<br>SOLID-STATE DEVICES |
| **Tuesday March 19 10:00 A.M.-12:30 P.M.** | *Session 9*<br>AUTOMATIC CONTROL—General | *Session 10*<br>NAVIGATION | *Session 11*<br>NEW BROADCAST DEVELOPMENTS | *Session 12*<br>ULTRASONICS ENGINEERING II—Technical Session | *Session 13**<br>ENGINEERING MANAGEMENT VIEWPOINTS | *Session 14*<br>ANTENNAS I—General | *Session 15*<br>INFORMATION THEORY—Review and Recent Advances | *Session 16*<br>MICROWAVE TUBES |
| **Tuesday March 19 2:30 P.M.-5:00 P.M.** | *Session 17*<br>GENERAL COMMUNICATIONS SYSTEMS | *Session 18*<br>MEDICAL ELECTRONICS | *Session 19*<br>NEW OPERATIONAL TECHNIQUES CONCERNING VIDEO TEST SIGNALS (A Panel Discussion) | *Session 20*<br>HIGH FIDELITY AND HOME MEASUREMENTS | | *Session 21*<br>ANTENNAS II—Broadband Antennas | *Session 22*<br>INFORMATION THEORY—Applications | *Session 23 (Joint)*<br>TELEVISUAL SYSTEMS DEVICES |
| **Tuesday March 19 8:00 P.M.-10:30 P.M.** | *Session 24 (Joint)*<br>APPLICATIONS OF ELECTRONICS TO AIR TRAFFIC CONTROL | | | | | | | *Session 25 (Joint)*<br>MICROMINIATURIZATION—The Ultimate Technique |
| **Wednesday March 20 10:00 A.M.-12:30 P.M.** | *Session 26*<br>ELECTRONIC COMPUTERS I—Digital Computers | *Session 27*<br>MAGNETIC RECORDING | *Session 28*<br>NUCLEAR INSTRUMENTATION | *Session 29*<br>CIRCUIT THEORY I—Symposium: Modern Methods in Network Theory | *Session 30**<br>ENGINEERING MANAGEMENT TECHNIQUES | *Session 31*<br>TRANSISTOR APPLICATIONS | *Session 32 (Joint)*<br>MICROWAVE ANTENNAS | *Session 33*<br>ELECTRON TUBES—General |
| **Wednesday March 20 2:30 P.M.-5:00 P.M.** | *Session 34*<br>SYMPOSIUM: LONG RANGE TELEMETRY AND REMOTE CONTROL | *Session 35*<br>SPEECH ANALYSIS AND AUDIO AMPLIFIERS | *Session 36*<br>TRANSISTORIZING NUCLEAR INSTRUMENTATION | *Session 37 (Joint)*<br>SYMPOSIUM: APPLICATIONS OF COMPUTERS IN BIOLOGY AND MEDICINE | | *Session 38*<br>COLOR TELEVISION RECEIVERS | *Session 39*<br>MICROWAVES I—Components | *Session 40*<br>PRODUCTION TECHNIQUES |
| **Thursday March 21 10:00 A.M.-12:30 P.M.** | *Session 41*<br>ELECTRONIC COMPUTERS II—Symposium: Computers in Simulation, Data Reduction, and Control | *Session 42*<br>CIRCUIT THEORY II—Transistor and Amplifier Circuit Design | *Session 43*<br>COMPONENT PARTS I | *Session 44*<br>INDUSTRIAL ELECTRONICS | *Session 45**<br>RELIABILITY PROBLEMS | *Session 46*<br>SYMPOSIUM: DIGITAL TECHNIQUES FOR PROBLEMS IN TELEMETERING AND REMOTE CONTROL | *Session 47*<br>MILLIMICROSECOND INSTRUMENTATION—Special Topics | *Session 48*<br>MICROWAVES II—Switches |
| **Thursday March 21 2:30 P.M.-5:00 P.M.** | *Session 49*<br>ELECTRONIC COMPUTERS III—Mainly Analog | *Session 50*<br>CIRCUIT THEORY III—Network Design Techniques | *Session 51*<br>COMPONENT PARTS II | *Session 52*<br>ANALYSIS AND TECHNIQUES FOR IMPROVED RELIABILITY | | *Session 53*<br>SYMPOSIUM: LOW LEVEL MULTIPLEXING FOR TELEMETERING AND REMOTE CONTROL | *Session 54*<br>INSTRUMENTATION II | *Session 55*<br>MICROWAVES III—General |

# ABSTRACTS OF TECHNICAL PAPERS

## SESSION 1*

**MON. 2:30–5:00 P.M.**

**WALDORF-ASTORIA STARLIGHT ROOF**

### Nonlinear Control Systems

*Chairman:* M. R. Aaron, *Bell Telephone Labs., Murray Hill, N. J.*

### 1.1. Nonlinear Compensating Networks for Feedback Systems

E. Mishkin and J. G. Truxal, *Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

A class of nonlinear networks for the compensation of linear and nonlinear feedback systems is described. The nonlinear networks are restricted to tandem, parallel, or feedback networks involving diode realization of the frequency-independent, nonlinear portion of the network and active, RC realization of the linear, frequency-dependent portion. The entire compensating network is inserted in a low-power-level, subsidiary feedback path.

Analog-computer and actual-system studies are described to indicate the results in two systems, one involving backlash and the other the combination of saturation and dead zone separated by a linear, integrating system.

### 1.2. Direct Synthesis through Block Diagram Substitutions

O. J. M. Smith, *Elec. Eng. Div., University of California, Berkeley, Calif.*

A direct synthesis procedure is presented which starts with a block diagram representation of four statements. These statements are: the problem, the restrictions, the criteria for design, and the optimum possible mode of operation. Each of these is presented as a block diagram, and the four are combined for a resultant statement of the problem. By the application of certain types of block diagram substitutions, the resultant system design can be obtained directly. Examples are shown for the control of dead time and nonminimum phase zeros, lightly-damped oscillatory systems, nonlinear control of saturating systems, and statistical synthesis.

### 1.3. A Nonlinear Control System for Wide-Range Input Signals

J. Tou and Y. H. Ku, *Moore School of Elec. Eng., University of Pennsylvania, Philadelphia, Pa.*

This paper gives a nonlinear control system for improving servosystem performance when the input signal varies over a wide dynamic range, causing sensitive controllers and error-detectors to undergo saturation and lose accuracy. In parallel with the controller of a servosystem, a nonlinear device is connected which is inactive when the signal is below a certain preset range. When the signal gets beyond this range, the sensitive control becomes saturated and the nonlinear device forms a by-pass for sending adequate information to the controlled system. Thus a combination of precision and coarse controls is incorporated in the system. Several possible cases are investigated. Analysis and design of such nonlinear control systems are described in the paper.

### 1.4. Switching Discontinuities in Phase Space

J. C. Hung and S. S. L. Chang, *Dept. of Elec. Eng., New York University, New York, N. Y.*

In an "optimum" nonlinear servosystem, which may either be of the contactor type or of the saturated type, the error derivatives are not continuous at the moment of switching if the transfer function of the controlled system has a polynomial in $s$ as its numerator.

The above effect, which has hitherto been neglected in the literature, forms the basis for a general. modified phase-space switching criteria for nonlinear servos of this type. The theory is applied to the depth control problem of a submarine to determine its theoretical optimum switching boundaries.

### 1.5. Nonlinear Techniques Applied to the Analysis of Pilot-Induced Oscillations

I. Van Horn, *Convair Div., General Dynamics Corp., San Diego, Calif.*

The advent of high-speed aircraft has brought stability control problems which stem from the fact that the human pilot is a part of the closed loop causing oscillations. Two factors combine to make analysis of the system difficult. They are the unpredictability of the pilot transfer function and the nonlinearity of the pilot controls.

This paper uses describing-function techniques together with a simplified representation of the pilot transfer function to analyze pilot-induced oscillations. Comparisons are made with flight test results to verify the conclusions reached.

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## SESSION 2*

**MON. 2:30–5:00 P.M.**

**WALDORF-ASTORIA ASTOR GALLERY**

### Vehicular Communications

*Chairman:* C. M. Heiden, *General Electric Co., Syracuse, N. Y.*

### 2.1. How Far Can We Go in Narrowing Channels in the Land Mobile Radio Services?

C. B. Plummer, *Federal Communications Commission, Washington, D. C.*

The ratio of assigned channel bandwidths to intelligence bandwidths has been gradually reduced from about 60 to 1 to 10 to 1 in the last 10 years. Will equipment manufacturers 10 years hence approach a ratio of 1 to 1?

### 2.2. Practical Modern Network Theory Design Data for Crystal Filters

M. Dishal, *Federal Telecommunication Labs., Nutley, N. J.*

In many radio systems (for communication, navigation, counter-measures, etc.) it is often desired that the system band-pass response be supplied by a single lumped filter having these two characteristics: 1) A high rate of cutoff in db per octave outside the pass band, and 2) a high "fractional midfrequency," *i.e.*, a high ($f_0/BW_{3db}$).

For midfrequencies up to vhf it will be shown that when the required product of the above two characteristics is greater than approximately 800, it is impractical to satisfy the band-pass filter specification with "economical" LC resonators because of their limited unloaded Q.

Because quartz-crystal resonators have appreciably higher unloaded Q's, it is logical to consider their use when faced with the above specifications.

However, most of the information available on the design of crystal filters is based on the design procedures of image-parameter-theory. Because the generator and load impedances physically available can never satisfy the "characteristic impedance" requirement of this theory, this design information is only approximately "correct." In this present paper the design procedures of modern network theory are applied to some of the most useful crystal filter configurations, and the more accurate design data which results is presented.

### 2.3. Recent Developments in Mobile Radio in Britain

J. R. Brinkley, *Pye Telecommunications, Ltd., Cambridge, Eng.*

The paper outlines briefly the development of vhf mobile radio in Great Britain and describes why AM is dominantly used as opposed to fm.

The first ten years of development have been based on two frequency bands 71.5—88 and 156—184, and 50-kc channeling has been used in the lower and 100 kc in the higher bands.

A second phase of development is being entered. This has been precipitated by the loss of channels to commercial tv and to the pressure for more mobile channels. This has led in turn to the need for closer channel spacing and more stringent specifications. Fifty-kc channeling becomes mandatory in 1957 and already apparatus capable of 25-kc channel spacing is in operation in both bands.

The paper concludes with a reference to the standardization of fm for international marine purposes and considers the future of AM and fm systems in relation to even closer channel spacings.

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* Sponsored by the Professional Group on Automatic Control. To be published in Part 4 of the 1957 IRE Convention Record.

* Sponsored by the Professional Group on Vehicular Communications. To be published in Part 8 of the 1957 IRE Convention Record.

### 2.4. A Manually-Operated Demand Repeater for the 450–470-MC Band

S. F. MEYER, *Allen B. DuMont Labs., Inc., East Paterson, N. J.*

Channel assignments for land-mobile services in the 450 to 470-mc frequency band offer many advantages to various user groups. One disadvantage, however, is the relatively poor mobile-to-mobile coverage compared to the lower frequency bands. This paper describes a relatively inexpensive modification to existing base station designs thereby permitting it to operate as an "In-The-Band" repeater. The mode of operation (dispatch vs repeater) is selectable by the system dispatcher. The overall result is a land-mobile system in the uhf band with unusual flexibility available on an extremely economical basis.

### 2.5. Sinad Interference Evaluation by Vosim

N. H. SHEPHERD, *General Electric Co., Syracuse, N. Y.*

A new method of adjacent channel interference evaluation which closely simulates actual operating conditions will be described. Several methods have been used in the past to evaluate adjacent channel interference, such as tone and noise modulation on the interfering signal. Each of those methods of modulation have failed to reproduce levels of interference comparable to typical voice modulation. This paper will show the results of interference levels produced by typical voice modulation compared to Vosim modulation. Specifications for standard Vosim modulation will be presented to eliminate future guesswork in evaluating adjacent channel interference by laboratory measurements.

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## SESSION 3*

**MON. 2:30–5:00 P.M.**

**WALFORD-ASTORIA JADE ROOM**

**Propagation**

*Chairman:* P. NEWMAN, *Air Force Cambridge Research Lab., Cambridge, Mass.*

### 3.1. The Refractive Index of the Atmosphere as a Factor in Tropospheric Propagation Far Beyond the Horizon

R. E. GRAY, *Federal Telecommunication Labs., Nutley, N. J.*

Transmission loss measurements have shown that a close correlation exists, in tropospheric scatter propagation, between the index of refraction of the atmosphere and path loss. The exceptional values of transmission loss, which occur for, say, 10 per cent of the time, are shown to be chiefly due to abnormal variations with height of the refractive index; measurements indicate, however, that the monthly median value of transmission loss is a function of the average monthly value of the surface refractivity on the transmission path. Thus, with tropospheric scatter propagation, the transmission loss over a given path is a function of the surface value of refractive index, as measured on the transmission path, and of the variation of this index with height.

### 3.2. Attenuation and Fluctuation of Millimeter Radio Waves

C. W. TOLBERT AND A. W. STRAITON, *Elec. Eng. Research Lab., University of Texas, Austin, Texas*

Radio propagation measurements at a waveength of 4.3 mm over a 61-mile path between Pikes Peak and Mount Evans are reported. An estimate of the water vapor and oxygen absorption is made from the variation of the attenuation with moisture content. The spectra and rms of the fluctuations of the millimeter waves are compared to those of the 3.2-cm waves recorded simultaneously. Instantaneous signal levels of the two wavelengths during a shower are compared. The results of these measurements are interpreted in terms of various millimeter wavelength measurements previously made by this Laboratory and by others.

### 3.3. New Evidence of Anomalous Transequatorial Ionospheric Propagation

O. G. VILLARD, JR., S. STEIN, AND K. C. YEH, *Radio Propagation Lab., Stanford University, Stanford, Calif.*

The existence of anomalous north-south propagation across the equator occasionally reported by amateurs has recently been verified by means of a rotating-antenna oblique ionospheric sounding station using the ground back-scatter technique and located in the West Indies. During the afternoon and evening hours of virtually every day, echoes were observed which are believed to be a consequence of one-hop transequatorial transmission over distances varying between 6000 and 8000 km. The observed maximum usable frequency was appreciably higher than that which would be predicted on the basis of vertical sounding measurements.

An explanation of this effect is proposed, which takes into account the remarkable increase in virtual height reported around 1800 hours local time at equatorial sounding stations. It is suggested that the observed transequatorial propagation takes place by two successive ionspheric reflections at the northern and southern edges of this equatorial bulge in the ionosphere. The unusually high muf's observed may be explained in part by the operation of the secant law in the case of a warped ionosphere, and in part by the fact that the two reflections occur well to the north and to the south of the geomagnetic equator.

### 3.4. Foreground Terrain Effects on Overland Microwave Transmissions

L. G. TROLESE, *Smyth Research Associates, San Diego, Calif.*

Small terrain irregularities can produce large variations in field strength. Measurements taken at 3300 and 9375 mc on a 46.3-mile non-optical link demonstrate that proper antenna siting can increase the signal by 15 to 30 db. A method of making propagation measurements to determine the effect ot terrain irregularities on microwave links is described. The coupling between atmospheric refraction and irregular terrain is discussed.

### 3.5. Mountain Obstacle Measurements

R. E. LACY, *Signal Corps Eng. Labs., Fort Monmouth, N. J.*

During 1955 several hundred paths were tested for mountain obstacle transmission loss in approximately 40 different locations in California. Thorough investigations were made of representative frequencies from 50 to 500 mc. Limited measurements were made at 900 and 1800 mc. Almost without exception, all paths planned provided the anticipated obstacle-gain phenomena. Suitable techniques were evolved whereby these phenomena may be used to extend radio communication ranges in mountainous areas along non-line-of-sight radio paths. It has been shown that 20–30-db reduction in comparison to tropospheric scatter paths in path loss is readily obtainable in mountainous areas. In comparison with smooth earth paths, the obstacle gain figures are in the order to 70 to 80 db. A reliable and accurate means for computing the mountain obstacle transmission loss, and a simplified siting procedure were developed.

### 3.6. Passive Repeater Using Double Flat Reflectors

R. F. H. YANG, *Andrew Corp., Chicago, Ill.*

A means of transmitting a microwave signal over an obstruction such as a mountain by using double passive repeaters is reported. It consists of two large, flat, rectangular reflectors nearly parallel to each other with an angle of 45° or less between the beam and the normal to the surface. The propagation loss through the reflectors is given. Analysis of a typical 7-kmc system shows that this scheme is practical. Its advantages over other existing systems are discussed. A brief comparison between this system and one using forward-scatter propagation through the troposphere is presented.

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## SESSION 4*

**MON. 2:30–5:00 P.M**

**WALDORF-ASTORIA SERT ROOM**

**Ultrasonics Engineering I—An Educational Session**

*Chairman:* W. G. CADY *Pasadena, Calif.*

### 4.1. Training in Ultrasonics

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* Sponsored by the Professional Group on Antennas and Propagation. To be published in Part 1 of the 1957 IRE CONVENTION RECORD.

* Sponsored by the Professional Group on Ultrasonics Engineering in cooperation with the Acoustical Society of America. To be published in Part 9 of the 1957 IRE CONVENTION RECORD.

F. V. HUNT, *Div. of Eng. and Appl. Phys., Harvard University, Cambridge, Mass.*

W. P. RANEY, *Harvard University, Cambridge, Mass.*

### 4.2. Ultrasonic Analysis

T. F. HUETER, *Acoustics Lab., Massachusetts Institute of Technology, Cambridge, Mass.*

### 4.3. Some Fundamentals of Transducer Design for the Sonic and Ultrasonic Range

F. MASSA, *Massa Labs., Inc., Hingham, Mass.*

### 4.4. The Ethics, Standards, and Objectives of the Recently Formed Ultrasonics Manufacturers Association

S. R. RICH, *The General Ultrasonics Co., Hartford, Conn.*

This educational session on ultrasonics has been organized through the cooperation of the Acoustical Society of America and the Institute of Radio Engineers. A close liaison has been established between the Committee on Sonic and Ultrasonic Engineering of the Acoustical Society of America and the Professional Group on Ultrasonics Engineering of the IRE. The recently organized Ultrasonic Manufacturers Association has shown a fine spirit of friendliness and cooperation with both of these organizations.

The program for the educational session will emphasize four important aspects of the rapdily expanding field of ultrasonics: 1) The fundamental training for professional activity in this field, 2) a broad view of the applications, both academic and industrial, as shown in problems demanding ultrasonic analysis, 3) the chief component part (the transducer) of all ultrasonic equipment, and finally 4) the role which the manufacturer assumes in promoting the science of ultrasonics. Outstanding authorities will speak on each of the four aspects.

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## SESSION 5*

**MON. 2:30–5:00 P.M.**

**WALDORF-ASTORIA GRAND BALLROOM**

### Aeronautical Electronics

*Chairman:* J. V. N. GRANGER, *Granger Associates, Inc., Menlo Park, Calif.*

### 5.1. A New Aircraft Static Discharger

R. L. TANNER, *Stanford Research Institute, Menlo Park, Calif.*

Aircraft flying through ice clouds or snow are charged to high potentials by the impact of the ice particles. The charge thus deposited is relieved by corona discharges occurring at the aircraft extremities. These discharges produce the rf noise known as "precipitation static" which couples into and frequently disables aircraft radio systems.

The precipitation charging rate rises rapidly with speed, and wick dischargers, intended to discharge this current noiselessly, are of little value on modern aircraft because of their limited discharge capacity, their aerodynamic drag, and their vulnerability to deterioration in high-speed airstreams.

The present paper describes a discharger employing a new principle for decoupling the noise produced in the discharges. Dischargers using the principle have much increased current capacity and can be flush mounted, making them suitable for installation on jet aircraft.

### 5.2. Thermal Design of Commercial Airborne Electronic Equipment

H. M. PASSMAN, *Collins Radio Co., Cedar Rapids, Iowa*

This paper deals with the test techniques necessary to thermally evaluate commercial airborne electronic equipment. Special emphasis is placed on how to instrument equipments for temperature, air static pressure, and air flow quantity measurement, including instrumentation of vacuum tubes for bulb hot-spot temperature measurement.

As a result of the thermal test program outlined above, design improvements have been put into practice. These improvements are described. They include: 1) Dust cover ventilation design for both forced cooled and non-forced cooled equipments, 2) location of components, 3) use of improved tube shields, 4) new forced air cooling tube sockets, 5) effective forced air cooling of modulized equipments, and 6) other.

### 5.3. The New Look in Electronic Controls

R. J. MEYER, *Collins Radio Co., Cedar Rapids, Iowa*

With the coming of performance jet aircraft, the modern airplane has developed into an exceedingly complex machine. This complexity has imposed an exacting and difficult task upon its pilot.

To implement improved operation of electronic controls and reduce pilot errors, human engineering principles have been applied to design criteria. Horizontal digital type counter presentations have been utilized in all frequency and channel displays. Numerals and/or letters are simple in design, and employ large, easily-read horizontal characters. Location, shape, and type of movement of each control knob have been carefully chosen to accomplish the desired action with a minimum delay or chance of error, even under operating conditions in which a pilot is wearing a full pressure suit.

Application of human engineering principles to electronic controls has resulted in vastly improved performance and a corresponding reduction in pilot errors. Integration of several control panels into one central panel results in a reduction of critical panel area required and in addition retains the advantage of improved performance. Future applications will undoubtedly open new approaches to improve visual displays and error-free operation.

### 5.4. Field Test Equipment for Airborne Radar

W. W. KEITH AND F. E. SEARS, III, *Raytheon Manufacturing Co., Wayland, Mass.*

The use of skilled electronics personnel is minimized, equipment reliability is improved, and down time is reduced by means of an adequate field test equipment program. An example is given of a field test equipment for an airborne radar which features: 1) Runway equipment, for quick preflight tests at the aircraft without use of an oscilloscope, and for maintenance tests to localize trouble to a major unit of the radar, and 2) hangar equipment, for go-no go tests of hermetically sealed modules such as the IF amplifier, afc and servo-amplifier. Self-testing circuits detect failure of the test equipment.

### 5.5. Hazardous Environmental Factors and Effects Related to High-Supersonic-Speed Bomber Defense Problems

I. KATZ, *General Electric Co., Ithaca, N. Y.*

Postulation, design, and performance of defensive subsystems depends upon the functional and organic compatibilities of equipment with the prevailing physical environment. Flight at high supersonic speeds introduces many environmental factors and effects that decisively influence the choice of system and critically affect the performance and reliability of equipment. Extremes of temperature, pressure, radioactivity, acoustical noise, motion, vibration, shock, and certain chemical changes, as functions of both natural and induced physical conditions, constitute the hazardous environment.

The intriguing effects of flight operations and associated propulsive power levels on the physical environment are discussed with a view toward defining, in engineering terms and numbers, the challenging physical conditions that equipment will encounter and with which it must cope. Suggestions are given for achieving practical compatibility and for striking design compromises that implement attainment of realistic objectives.

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## SESSION 6*

**MON. 2:30–5:00 P.M.**

**NEW YORK COLISEUM MORSE HALL**

### Multiplex Communications System

*Chairman:* J. Z. MILLAR, *Western Union Telegraph Co., New York, N. Y.*

* Sponsored by the Professional Group on Aeronautical and Navigational Electronics. To be published in Part 8 of the 1957 IRE CONVENTION RECORD.

* Sponsored by the Professional Group on Communications Systems. To be published in Part 8 of the 1957 IRE CONVENTION RECORD.

### 6.1. Signal Mutilation and Error Prevention on Short-Wave Radio-Teleprinter Services

J. B. MOORE, *RCA Communications, Inc., New York, N. Y.*

Noise and fading conditions on such services differ markedly from those generally assumed in analyzing telegraph or data-handling services over wire lines; the problem being rendered particularly difficult by the wide range of types, durations, and rates of signal mutilation. The most successful system for mutilation detection and error prevention is the so-called "7-Unit" constant-ratio code and basic system plus Automatic-RQ. Observational data and theoretical analyses. from various sources, indicate improvement ratios—reduction in number of character transpositions—ranging from 20/1 to 1000/1 to 50,000/1 for signal-mutilation rates from 4 in 10 to 1 in 100 to 1 in 1000 characters respectively.

### 6.2. Time-Division Multiplex System with Addressed Information Packages

R. FILIPOWSKY AND E. SCHERER, *Westinghouse Electric Corp., Baltimore, Md.*

A digital multiplex communication system is described wherein the channels are combined in asynchronous time division; there is no fixed and predetermined sequence by which the individual transmission intervals would be allotted to the various channels. On the contrary, each transmission interval is electronically "auctioned" and is allotted to the highest bidder, *i.e.*, to the channel having instantaneously the most urgent need for forwarding its message or a fraction thereof. The receiver has no knowledge to which channel the information within one transmission interval should be directed, unless each interval will carry its own address. One interval may contain one quantized sample only or a larger amount of time compressed information, *i.e.*, one information package.

### 6.3. A New Time-Division Multiplex System

W. J. BIEGANSKI AND L. M. GLICKMAN, *Radio Corp. of America, Camden, N. J.*

A new time division, pulse position modulation, voice multiplex system has been developed. The system is completely transistorized. Transistor matrix techniques are used in the distributor, resulting in highly accurate timing without any adjustments. A phase lock loop serves to synchronize the demultiplexer to the multiplexer with greatly improved noise performance. Two versions of multiplex equipment, with different channel capacities, are described: 7-channel equipment, with provision for combining for 14-channel use, and 23-channel equipment, with provision for combining for 46-channel use. Drop and insert facilities are provided at reshapers. Transistor circuitry constituting the system is described in some detail.

### 6.4. 58-Channel PCM System

S. M. SCHREINER, *Federal Telecommunication Labs., Nutley, N. J.*, AND A. R. VALLARINO, *Precision Technology, Inc., Livermore, Calif.*

This paper describes a 48-channel pulse code modulation system (the AN/TCC-15) employing instantaneous compandors and 6-bit (64-level) cathode-ray-type coding tubes. The equipment accepts 47 telephone channel inputs and provides a binary-coded output signal capable of transmission over any 2.5-mc video circuit. The coding is accomplished by two 24-channel coders, each using a coding tube which generates a CP code and a high-speed flip-flop which translates the signal to a true binary code. The two-coder output signals are then combined before transmission. One hundred per cent standby of all common 48-channel equipment insures 24-channel operation in the event of any failure. The over-all system provides 47 audio channels with a 300-3600-cps pass band, an audio s/n greater than 60 db, and an interchannel crosstalk of better than $-53$ db. Recent field tests indicating a mean life of 2500 hours between failures attest to the high reliability designed into the equipment.

### 6.5. Portable Multichannel 11,000 MC Radio Link

H. ENGELMANN, H. A. FRENCH, M. W. GREEN, AND J. HARVEY, *Federal Telecommunication Labs., Nutley, N. J.*

A portable 7-channel telephone relay link operating in the 9 to 11-kmc region using time-division multiplex pulse position modulation is described. Printed circuitry and the use of miniature type components have been used to make possible the construction of an equipment $23 \times 21 \times 12\frac{1}{2}$ inches in size and weighing less than 125 pounds, containing multiplex equipment, transmitter, receiver, and antenna, all in one package.

The electrical design is described and the mechanical construction used to obtain a high space utilization factor is shown.

Applications and field experience obtained using the first developmental models is detailed.

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## SESSION 7*

**MON. 2:30–5:00 P.M.**

**NEW YORK COLISEUM MARCONI HALL**

### Information Theory—Coding and Detection

*Chairman:* W. B. DAVENPORT, JR., *Lincoln Labs., Massachusetts Institute of Technology, Lexington, Mass.*

* Sponsored by the Professional Group on Information Theory. To be published in Part 2 of the 1957 IRE CONVENTION RECORD.

### 7.1. Optimum Decision Feedback Systems

B. HARRIS, A. HAUPTSCHEIN, AND L. S. SCHWARTZ, *College of Eng., New York University, New York, N. Y.*

This paper studies the problem of reducing decision risk in the extraction of signals from noise for the case of decision feedback in which the transmitter is called upon to repeat signals corresponding to doubtful observations. The objectives of the paper are twofold: 1) To develop a theory of decision feedback systems by means of which the optimum conditions of operation in the least-cost sense can be specified, and 2) to demonstrate that integration or a simple parity check for interpreting doubtful observations is less efficient in its use of the fundamental communication parameters of power, bandwidth, and time than decision feedback.

### 7.2. Sequential Decoding for Reliable Communications

J. M. WOZENCRAFT, *Massachusetts Institute of Technology, Cambridge, Mass.*

A code for constant-rate communication over binary symmetric channels is presented. For coding, parity checks over subsets (selected by convolution) of preceding information digits are inserted between information digits. For decoding, the first information digit is determined by progressively discarding as improbable all possible messages inconsistent with its correct value. The discard probability threshold is varied to minimize both the error-probability and the average number of computations. The entire procedure is then iterated. The error probability is shown to decrease exponentially with code length, while the average number of binary computations per bit grows approximately linearly. Rates near one-half capacity appear feasible.

### 7.3. A Non-Mean-Square-Error Criterion for the Synthesis of Optimum Sampled-Data Filters

A. R. BERGEN, *Dept. of Elec. Eng., Columbia University, New York, N. Y.*

This paper presents a non-mean-square-error criterion for designing optimum linear finite memory filters for the continuous prediction and smoothing of sampled data. It is suggested by, and approximates, a generalized tolerance criterion whose use, while desirable in many applications, is excessively complicated. The new criterion simplifies the design considerably for inputs which are the sampled sum of Gaussian noise and a nonstationary message; the message consists of sections of finite duration having a polynomial variation with time alternating with arbitrary transitions. The design procedure is straightforward, requiring only a most elementary description of the inputs and results in fairly simple filter structures.

### 7.4. Message Redundancy vs Feedback for Reducing Message Uncertainty

W. B. BISHOP AND B. L. BUCHANAN, *Air Force Cambridge Research Center, Bedford, Mass.*

The use of redundancy in the coding of digital messages is generally directed toward overcoming the effect of a very small number of errors. The use of a small amount of feedback can be effective in reducing the uncertainty concerning the correctness of messages. Two assumptions normally made in these cases are: 1) The probability of error in any digit is independent of errors that occur elsewhere, and 2) no errors occur in the feedback path. There is a large class of digital communication systems for which these assumptions do not hold: the advantages of feedback for this class are compared with the advantages obtainable by using selective redundancy without feedback. For the class of communication systems in which the feedback channel is the same as the forward channel, it is proved that message uncertainty is reduced more by the transmission of $M+C$ digits in the forward direction (where $M$ represents the number of message digits and $C$ represents the number of redundant digits) than by the transmission of $M$ digits in the forward direction and $C$ in the reverse. The probability distributions of feedback termination and the message uncertainty at the time of termination are derived for systems in which feedback for any message is continued until the received feedback message agrees with the message sent. It is also shown that a new statistical error-checking procedure has advantages over other forms of redundancy when the probabilities of errors are within certain defined limits. A means of implementing this statistical error-checking procedure with simple electronic equipment is described.

### 7.5. The Analysis of Post-Detection Integration Systems by Monte Carlo Methods

R. DILWORTH, *California Institute of Technology, Pasadena, Calif.*, AND E. ACKERLIND, *Radio Corp. of America, Los Angeles, Calif.*

By means of a random sampling procedure the cumulative probability distribution of threshold crossings has been determined for the cases of: 1) IF amplifier, square-law detector, and post-detection integration, and 2) IF amplifier, linear detector, and post-detection integration. Various signal to noise ratios and various durations of signal are considered.

Random samples are selected from a population whose distributions are computed at the output of the detector for both cases and are processed in a manner determined by the post-detection integration filter characteristics.

Comparison is made with results obtained by other methods.

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## SESSION 8*

**MON. 2:30–5:00 P.M.**

## NEW YORK COLISEUM FARADAY HALL

### Solid State Devices

*Chairman:* W. J. PIETENPOL, *Bell Telephone Labs., Murray Hill, N. J.*

### 8.1. A New High-Frequency N-P-N Silicon Transistor

A. B. PHILLIPS AND A. M. INTRATOR, *General Electric Co., Syracuse, N. Y.*

This paper describes a newly-developed meltback diffused base transistor having a low series saturation resistance and a nominal alpha cutoff of 25 mc, which was designed for application in high-frequency linear amplifiers and general purpose switching circuits. A general description of the device is given and the theory of obtaining a $p$ base region by means of the diffusion-meltback technique is discussed.

Data are presented showing the variation of device parameters with temperature, frequency, current, and voltage. The large signal behavior of the device is shown and its performance in switching circuits is described.

### 8.2. Noise Figures in Semiconductor Dielectric Amplifiers

J. M. WALKER, R. E. SMITH, AND E. M. WILLIAMS, *Carnegie Institute of Technology, Pittsburgh, Pa.*

Noise measurements on amplifiers utilizing the voltage-dependent junction capacitance of germanium diodes as the nonlinear element in a series resonant circuit are described. Noise measurements on the diode amplifier are discussed in conjunction with an equivalent circuit which is used to explain the noise measurements and the noise figure. The important noise components are shown to be attributable to noise from the diode and noise amplitude and frequency modulation in the carrier. Low-temperature noise measurements on the amplifier and the diode parameters are presented together with a description of other methods of improving and optimizing the noise figure.

### 8.3. Determination of Thermal Resistance of Silicon Junction Devices

H. C. LIN AND R. E. CROSBY, JR., *CBS-Hytron, Semiconductor Operations, Lowell, Mass.*

The usual method of determining the thermal resistance of germanium junction devices is to measure the variation of the reverse saturation current which increases as the junction temperature rises. This method cannot be used with silicon junction devices, as the leakage current rather than the saturation current of these devices is the major portion of the reverse current. In the system described here the forward characteristics vs temperature rise are measured. This is possible as the forward voltage drop of a silicon junction device varies with temperature in a manner that is both predictable and relatively simple to measure. Theoretical results and experimental data are plotted and compared at various forward currents. There is a very good agreement. The operation of the circuit is also described.

### 8.4. An Alloy-Type Medium Power Silicon Transistor

H. G. RUDENBERG AND G. FRANZEN, *Transitron Electronic Corp., Wakefield, Mass.*

This silicon transistor was developed to evaluate an alloy fusion process for producing a medium power audio transistor for low voltage airborne use operating above 100°C. The 10-watt unit described has a power gain of 20 db, and weighs less than ½ ounce. Saturation voltages have been reduced below 2 volts. Internal heat dissipation is removed by thermal contact with the chassis heat sink. Some problems of design and processing affecting the ultimate characteristics will be discussed. Simplified test methods and a power converter utilizing this transistor will also be presented.

### 8.5. A New Semiconductor Device

C. A. ALDRIDGE, *General Electric Co., Syracuse, N. Y.*

This paper presents the basic biasing modes of operation for a new type of semiconductor amplifying device. In addition, the 24 basic signal-flow configurations are shown. Since the number of usable circuit configurations is quite large, a system is proposed for identifying the particular configuration in an organized manner. It is shown how each of the basic circuit forms can be realized by any one of several circuit connections, depending on the associated components used.

The mechanical and electrical characteristics of the new device are described. It is also shown that significant circuit economies are achieved.

### 8.6. Cadmium Sulfide Photocapacitors

F. GORDON, JR., P. A. NEWMAN, J. HANDEN, H. JACOBS, *Signal Corps Eng. Labs., Fort Monmouth, N. J.*, AND A. RAMSA, *CBS Labs., New York, N. Y.*

A description of the photocapacitor, including the preparation of the photosensitive materials and the assembly of the device, is given. Methods and equipment for evaluation of the device are described along with some results. A brief discussion of the possible mechanisms responsible for this change in capacitance with radiation is given. An attempt is made to correlate the theoretical and experimental results. Some of the possible applications are indicated along with the potentialities and limitations of the present device.

Changes in capacitance for a given dielectric material in the photocapacitor are dependent upon the wavelength and intensity of the light source and the frequency and magnitude of the applied voltage. The magnitude of these changes ($\Delta c/c$) and the variation in magnitude and response time with the variation of the above parameters is illustrated. Also shown is performance variation with different impurities and impurity concentrations and the method of preparation of these materials

* Sponsored by the Professional Group on Electron Devices. To be published in Part 3 of the 1957 IRE CONVENTION RECORD.

# SESSION 9*

## TUES. 10:00 A.M.–12:30 P.M.

## WALDORF-ASTORIA STARLIGHT ROOF

## Automatic Control—General

*Chairman:* G. S. AXELBY, *Westinghouse Electric Corp., Baltimore, Md.*

### 9.1. Digital Controllers for Feedback Systems

J. R. RAGAZZINI, *Columbia University, New York, N. Y.*

A digital controller for a feedback control system is a computer which accepts a sequence of numbers at its input and delivers a processed sequence of numbers at its output. The implementation of the computer may be completely analog or digital, or a combination thereof, as will be shown by typical designs. The output sequence of the controller may be generated at a rate equal to or at a multiple of the input sequence rate. For linear systems, it is possible to program the computer to cause the control system to have a desirable over-all prototype response subject to certain theoretical limitations. For computers in which the output data rate is a multiple of the input rate, faster response can be achieved. The theory of sampled-data systems is applied to the synthesis of the digital controller program which will produce these desirable prototype response functions. The theoretical restrictions which limit the choice of response functions are pointed out Examples of typical designs which illustrate the theory are given

### 9.2. Sampling in Linear and Nonlinear Feedback Control Systems

J. KUKEL, *Servomechanisms, Inc., Hawthorne, Calif.*

The $Z$ transform has been popularized by Ragazzini and Zadeh at Columbia University and Linvill and Salzer of M.I.T The usual approach to $Z$ transformation is through the Laplace transform. However the Laplace transform and its associated limitations can be completely by-passed. If the time domain differential equation is integrated, the appropriate numbers of times and the integrals changed to the convolution form, a time domain convolution equation with initial conditions results. The $Z$ transform of the convolution equation can be taken directly. Use of tabulated tables similar to the Laplace transform are most helpful. The convolution and the $Z$-transform operations allow the analysis or synthesis of continuous and sampled-data control systems which are linear or nonlinear. Nonlinear systems are included since the Laplace transform, which is limited to linear systems, is avoided.

The procedure can cover higher order and variable forcing function nonlinear systems. It does not have the limiting restrictions of the phase trajectory and the describing function technique. Although approximations are made in the substitution of continuous functions by their sampled form, the cumulative error in transient analysis can be made small by judicious choice of sampling interval. Steady-state analysis is treated as repeated transient and periodicity is assumed. Exactness occurs in period determination. Some error occurs in waveform amplitude. The error here, as before, is reduced by choice of smaller sampling interval.

### 9.3. The Role of Automatic Control in the Modification of Aircraft Dynamic Flight Characteristics

W. N. TURNER, *Hughes Aircraft Co., Culver City, Calif.*

The basic ideas and methods used in the determination of aircraft dynamic flight characteristics are described, and a physical description of the several typical primary dynamic flight modes of aircraft is given. The reasons for some of the undesirable features of these modes are indicated, and the manner in which automatic control techniques can be utilized to modify and improve the flying qualities is illustrated.

### 9.4. Solution of Statistical Problems by Automatic Control Techniques

R. L. COSGRIFF, *Dept. of Elec. Eng., Ohio State University, Columbus, Ohio*

It has been widely accepted that many processes in the broad field of science essentially can be classified as closed-loop systems. Many of the problems encountered in these fields are essentially statistical in nature and, as a result, have not been analyzed using the differential equation approach, however autocorrelation and spectral density approach have been used. It is demonstrated that the differential equation approach can be used to determine the expected behavior of many of the "statistical type systems."

### 9.5. The Design of Optimum Filters and Predictors

C. W. STEEG, *Dynamic Analysis and Control Lab., Massachusetts Institute of Technology, Cambridge, Mass.*

The solution of many automatic control problems depends upon obtaining some functional of a signal component of a measured quantity when the measurement contains both signal and noise. When the input to the system consists of a random signal plus a random noise, both derived from linear, time-varying, intergrodifferential operations on uncorrelated noise sources, a technique can be applied to simplify the design of filters and predictors. This technique involves the choice of a particular configuration for the optimum filter; a proof is given that knowledge of a certain operator in this configuration is sufficient to allow the determination of a significant class of optimum systems. Details of the method are presented to provide means for circumventing practical difficulties encountered in the actual design of optimum systems based on the earlier techniques devised by Weiner.

* Sponsored by the Professional Group on Automatic Control. To be published in Part 4 of the 1957 IRE CONVENTION RECORD.

# SESSION 10*

## TUES. 10:00 A.M.–12:30 P.M.

## WALDORF-ASTORIA ASTOR GALLERY

## Navigation

*Chairman:* R. C. NEWHOUSE, *Bell Telephone Labs., Whippany, N. J.*

### 10.1. A Realistic Radar Clutter Simulator

J. ATKIN, H. J. BICKEL, AND M. R. WEISS, *Dept. of Elec. Eng., Columbia University, New York, N. Y.*

The need for an entirely realistic search radar simulator arises in the design and testing of automatic detection, track-while-scan, beam-splitting, and other data-processing systems. Equipment capable of simulating the statistical and systematic characteristics of echoes received from isolated targets has been described in the literature. The authentic simulation of radar echoes from wooded terrain, meteorological phenomena etc., has been hampered by the difficulty associated with the reproduction of the particular statistical and spectral properties of reflections from this class of "extended targets." The device whose theory and implementation will be described, generates "sweeps" of Rayleigh distributed clutter having video bandwidth in range, but only audio (or sub-audio) bandwidth at constant range as a function of time. The audio fluctuation rate of the video at constant range duplicates the high correlation in real clutter amplitudes received during a number of consecutive sweeps; this differentiates such signals from ordinary noise. The key components in the clutter simulator are an ultrasonic delay-line memory and a 30-mc Gaussian noise source.

### 10.2. A Precision Multipurpose Radio Navigation System

W. N. DEAN, R. L. FRANK, AND W. P. FRANTZ, *Sperry Gyroscope Co., Div. of Sperry-Rand Corp., Great Neck, N. Y.*

This paper describes a pulsed hyperbolic navigation system similar to Loran, but achieving greater range, accuracy, and reliability over land and sea at all altitudes through the use of ground-wave cycle-phase-measuring techniques on a 100-kc carrier frequency.

Part I describes the general features of the system, analyzes the effect of station geometry on accuracy and interprets the results of extensive field tests. The system is shown to have sufficiently low altitude coverage and accuracy for point-to-point helicopter navigation coupled with range sufficient for transcontinental jets and transoceanic air and surface navigation. These characteristics can be obtained with very few transmitting stations, compared with other type systems

Part II discusses the choice of carrier frequency including calculations and measure-

* Sponsored by the Professional Group on Aeronautical and Navigational Electronics. To be published in Part 8 of the 1957 IRE CONVENTION RECORD.

ments of propagation parameter such as noise levels, groundwave attenuation, skywave attenuation and delay, groundwave phase stability, and dispersive effects. Observations of groundwave phase variations apparently due to weather are reported.

Part III outlines the instrumentation techniques utilized to achieve time-difference measurement to a few parts per million including automatic pulse envelope time difference measuring systems, and cycle-phase measuring systems incorporating cross-correlation detection and narrow banding. Transmitter and antenna designs to produce necessary sharp pulses are also described.

### 10.3. Some Final Approach System Requirements for High Landing Rates

F. H. BATTLE, JR., *Airborne Instruments Lab., Inc., Mineola, N. Y.*

The density and character of air traffic in prospect for the near future substantiate a goal of two instrument landings per minute, with a 4:1 range of approach speeds. Feasibility of scheduling landings at such a rate depends on a short final approach to avoid overtake, and on completion of turn-on during the final minute of flight. Horizontally-curved, rate-stabilized paths such as are obtained with flight directors, are analytically extended to their limits of usability with wide-sector final-approach aids. The practical maximum of horizontal coverage to be required of the final-approach system is derived.

### 10.4. General Characteristics of a 1000-MC Instrument Landing System

A. CASABONA, *Federal Telecommunication Labs., Nutley, N. J.*

By Air Force contract, the Federal Telecommunication Laboratories developed a cw localizer and glide slope, operating in the 960-mc to 1215-mc range, whose description and test forms the subject of this paper. The data indicate a substantial improvement in site freedom due mainly to the specular nature of 1000-mc reflections as compared to reflections at lower frequencies.

The cw localizer was later modified for pulsed operation which permitted integration of the landing signal with other navigation functions. The resulting equipment provides localizer guidance and distance to touchdown on a common transmission from the ground and utilizes a common airborne receiver.

The techniques discussed in the paper can be applied to the design of a landing system which offers great reduction in size of equipment, excellent mobility, increased freedom in siting, and integration with other navigation aids.

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## SESSION 11*

**TUES. 10:00 A.M.–12:30 P.M.**

**WALDORF-ASTORIA JADE ROOM**

### New Broadcast Developments

*Chairman:* C. E. SMITH, *Cleveland, Ohio*

### 11.1. An Analysis of Packing Density of Information in High-Velocity Transverse Video Magnetic Recording

W. SELSTED, *Ampex Corp., Redwood City, Calif.*

Coincidences exist between the nature of television synchronizing and video signals, and the peculiarities of fm systems employing high ratios of modulating frequency to carrier frequency, and high ratios of deviation to carrier These can be exploited to compress the television frequency range into a band which is compatible with the capabilities of a magnetic recorder using high tape-to-head relative velocities. The means by which this exploitation is accomplished also make possible desirable manufacturing and maintenance tolerances

### 11.2. High-Light Aperture Equalizer

M. V. SULLIVAN, *CBS Labs., New York, N. Y.*

The present method of aperture equalization of a television signal has one severe limitation. This is that the high-frequency noise in the picture signal is increased along with the high-frequency signal information. The amount that the high-frequency fine detail can be increased is limited by the undesired noise.

The "High-Light Aperture Equalizer" provides more equalization without increasing the noise in the same proportion. This feature is obtained by dividing the video signal into two parts with respect to amplitude, and equalizing the portion that is relatively free of noise.

The circuit has been successfully tested in monochrome and color television.

### 11.3. Single-Sideband Broadcast Developments

L. KAHN, *Kahn Research Labs., Freeport, N. Y.*

A new system for allowing an increase in the number of AM broadcast channels is described, which is called "Compatible Single-Sideband." This system would require use of a transmitter adapter but does not necessitate any change in the home receiver.

The advantages of this system may be outlined as follows: 1) The signal occupies approximately one-half the spectrum as does a conventional AM double sideband signal This reduction in bandwidth provides a means for reducing adjacent and cochannel interference; 2) reduction in fading distortion thus improving nighttime coverage; and 3) reduction in bandwidth also allows an improvement in signal-to-noise ratio of 3 db *or* approximately double the high-frequency response of typical home receivers.

A block diagram of an adapter suitable for use with any standard AM transmitter is discussed.

### 11.4. UHF High-Power Transmitting Developments

J. E. YOUNG, L. L. KOROS, AND I. MARTIN, *Radio Corp. of America, Camden, N. J.*

This paper treats development of transmitting equipment for the uhf portion of the spectrum involving high-power transmitters and high-gain antenna design. Information will be given concerning tubes having 25-kw power handling capabilities in this portion of the spectrum and trends in this area of development. A summary of power measurements is contained and experience with commercial application at an existing uhf station is treated.

### 11.5. A Dynamic Standard for Color Television Systems

R. C. KENNEDY, *National Broadcasting Co., New York, N. Y.*

This paper relates to the possible secondary usages which may be made of the time required for the beam in a television picture tube to return from the bottom to the top of the picture.

Heretofore, no application of this so-called retrace time has been attempted. However, it has been found that several different kinds of information may be transmitted while the actual television show is in progress.

One of the most important sets of signals which may be sent during this unused time enables the engineer to evaluate the performance of the complete television signal path from camera to the transmitter and receiver. These signals permit correct adjustment of signal strength, of the monochrome and color portions of a television program at successive points along the route of the show.

The result of having such information sent along with the program is to improve the received signal and enable the viewer to more accurately adjust his receiver.

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## SESSION 12*

**TUES. 10:00 A.M.–12:30 P.M.**

**WALDORF-ASTORIA SERT ROOM**

### Ultrasonics Engineering II—Technical Session

*Chairman:* J. W. HORTON, *U. S. Navy Underwater Sound Lab., New London, Conn.*

### 12.1. Sea Clutter in Radar and Sonar

R. M. HOOVER AND R. J. URICK *Ordnance Research Lab., Pennsylvania State University, University Park, Pa.*

The surface of the sea back scatters both electromagnetic radiation incident from above and acoustic radiation incident from below. Recent measurements of sound scattering by the sea surface make possible some interesting comparisons between radar sea-clutter and sonar surface reverberation at the same wave-

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* Sponsored by the Professional Group on Ultrasonics Engineering. To be published in Part 9 of the 1957 IRE CONVENTION RECORD.

length. The acoustic data provide the first systematic information about the sonar back scattering coefficient of the sea surface and give some hints as to the scattering processes that operate. A comparison of sonar with available radar data at approximately the same wavelength indicates that, although the backscattering cross sections are of the same order of magnitude, there appear to be differences in the effects of frequency, wind-direction, and grazing angle in the two cases.

### 12.2. Transducer Comparison Methods Based on the Electromechanical Coupling Coefficient Concept

R. S. Woollett, *U. S. Navy Underwater Sound Lab., New London, Conn.*

The electromechanical coupling coefficient ($k$) has proven to be a valuable unifying concept which facilitates preliminary comparison of different transduction methods, when a choice of transducer types has to be made for a particular application. The fundamental character of the electromechanical coupling coefficient is made evident by its energy definition, while equations relating it to equivalent circuit parameters and the electromechanical constants of the transducer facilitate its practical use.

The important role which the coupling coefficient plays in determining a transducer's performance is brought out by presenting equations for various performance properties, such as bandwidth, efficiency, transducer noise, electrical power factor, and power capacity. Factors influencing $k$ are discussed for two general categories of transducers: body-force and surface-force types; and coupling coefficient formulas are given for simple prototypes of the various transducer classes. Finally, an example is given of comparison of the different transduction methods for a particular application.

### 12.3. Polarization of Barium Titanate Ceramic

J. G. Froemel, *Gulton Industries, Inc., Metuchen, N. J.*

Polarization of barium titanate ceramic is one of the basic and fundamental properties of the materials and one which few have investigated. The following is a brief description of polarization, polarization procedures, and the methods of evaluating these polarization procedures. Various methods of polarization are presented along with their evaluation.

The resonance antiresonance method has been used by many to evaluate the various piezoelectric and elastic properties of barium titanate ceramics. However, this can lead to serious errors since various modes can add or subtract from the constant wanted. Polarization procedures are rarely given, which can account for the wide variation of such piezoelectric parameter as $d_{33}$, $d_{13}$, and others.

The $d_{33}$ can be shown to be nonlinear and of course frequency, field, and time dependent.

A direct method of measuring the piezoelectric coefficients is presented which eliminates or takes into consideration the above items.

The optimum polarization procedure for barium titanate is presented along with a discussion of the failures which make optimum polarization impossible in some cases. Reduction and diffusion in barium titanate have a pronounced effect on the polarization and are described in detail. Methods of eliminating or retarding these are presented.

A brief review of the chemical theory of polarization is also presented.

### 12.4. Ultrasonic Liquid Level Sensor

R. L. Rod, *Acoustica Associates, Inc., Glenwood Landing, N. Y.*

The ever-present need for a reliable yet simplified liquid level sensor is met using a newly-developed ultrasonic probe. Operating on the principle that an ultrasonic transducer in contact with a liquid presents a terminal impedance appreciably different than obtained when it is purely "air-loaded," the new sensor basically consists of a subminiaturized transducer appropriately mounted in a tank at a preset height. This change of impedance is not affected by characteristics of the liquid. A transistorized electronic control unit external to the tank energizes a relay when the level passes the sensitive face of the transducer with an accuracy of ±1/32 inch. The complete level sensor, including the transducer and the electronic control, weighs under 12 ounces and consumes only 100 milliwatts during operation. Performance is stable when sensing virtually all liquids at temperatures of from −270°F to 350°F. The sensor is impervious to foam and clinging droplets and is ideally suited to critical applications in airborne gauging and processing control systems requiring highest reliability and accuracy.

### 12.5. Precision Calibration of Ultrasonic Fields by Thermoelectric Probes

W. J. Fry and F. Dunn, *Bioacoustics Lab., University of Illinois, Urbana, Ill.*

The highly-stable, small, and readily-constructible ultrasonic probe developed and in use at this laboratory for the past five years will be discussed from the point of view of construction calibration, operation, and application.

This transient-type thermoelectric probe, which is currently being used in a number of other laboratories, yields information concerning the pressure amplitude, particle velocity amplitude, and intensity of the ultrasonic field in which it is placed. If the field characteristics are known, the principle of operation of the probe provides a method for determining the absorption coefficient of minute quantities of materials.

### 12.6. New Magnetostriction Filters for the MF Band

R. T. Adams, *Federal Telecommunication Labs., Nutley, N. J.*

A series of narrow-band filters has been constructed for carrier pick-off in a developmental telephone-carrier multiplex terminal. The filters are used to select clean single-frequency carriers from the harmonics of a 36-kc or 72-kc pulse train.

The resonator element is a simple one-piece $H$ structure, punched from Ni Span "C" sheet giving a double-tuned response shape. In the frequency range from 500 to 900 kc the filters provide 3 to 6-db voltage gain, operating from 30 ohms into a high impedance, with a pass band of 500 cycles and greater than 50-db rejection of adjacent 36-kc harmonics. The experimental units operate over −30 to +56°C. with a temperature coefficient of less than 2 ppm/°C. and withstand moderate shock and vibration. A hermetically-sealed case avoids formation of frost on the resonators at low temperature.

## SESSION 13*

**TUES. 10:00 A.M.–12:00 NOON**

**WALDORF-ASTORIA GRAND BALLROOM**

**Engineering Management Viewpoints**

*Chairman:* E. A. Walker, *Pennsylvania State University, University Park, Pa.*

### 13.1. Education: Academic Training for Engineering Management

E. Shapiro, *School of Industrial Management, Massachusetts Institute of Technology, Cambridge, Mass.*

The trilogy of science, engineering, and population point to an expanding economy in the generation ahead. To manage this expansion wisely requires the preparation of wise managers in increasing numbers. The management education program at the School of Industrial Management is designed to build on a base of science and/or engineering education. The object of the school is to prepare men to take wise and effective action in the management of industrial concerns. The education offered derives from the belief that a successful manager in the modern technological age should understand the principles of science and engineering upon which the application of industry rests. He should know the methods of production and distribution. He should understand as much as possible about himself, the nature of other human beings, and the environment—economic and political—in which he lives. In addition to this body of knowledge, a successful manager should possess the intellectual skill and resilience to deal with the novel situations presented to him by changes in the economic, social, and technological situation.

The philosophy and implementation of these objectives at the School of Industrial Management will be discussed as illustrative of an experiment in management education for engineering.

### 13.2. Finance: Wall Street Looks at Engineering Management

O. C. Roehl, *Keystone Custodian Funds, Boston, Mass.*

Management is a most important factor in security analysis. *Who* is behind a balance sheet is usually more important than *what*. The quality of a company's management is reflected in the price of its securities. In research and engineering based industries, such as the electronics industry, the quality of a company's engineering management is a most important determinant of a company's success. This fact is realized by investors and more and more attention is being given to evaluating the effectiveness of the engineering and research work of individual companies. While many factors account for the difference in performance of a security, yet the quality of a company's engineering management is often a primary reason why one company outperforms another in the security markets.

* Sponsored by the Professional Group on Engineering Management. To be published in Part 10 of the 1957 IRE Convention Record.

In these days of rapid technological changes, it is most important that the investor limit his investments to the companies with the best engineering and research management.

The factors that institutional investors consider in evaluating engineering management will be discussed and certain case histories will be reviewed to illustrate the role engineering management is playing and has played in determining the market value of the securities in the electronics industry.

### 13.3. Industry: Development of Engineering Managers Within Industry

E. I. GREENE, *Bell Telephone Labs., New York, N. Y.*

## SESSION 14*

**TUES. 10:00 A.M.–12:30 P.M.**

**NEW YORK COLISEUM MORSE HALL**

### Antennas I—General

*Chairman:* H. A. WHEELER, *Wheeler Labs., Inc., Great Neck, N. Y.*

### 14.1. On Ferrite Loop Antenna Measurements

J. L. STEWART, *California Institute of Technology, Pasadena, Calif.*

A general discussion relating to the application of small loop antennas with air and ferrite cores is given. A general procedure for simplified testing of ferrite-loaded magnetic-type small antennas is outlined in which radio-frequency radiation performance is expressed in terms of quantities easily measured at audio frequencies. Only a single measurement is needed to characterize the elementary-dipole-type ferrite-loaded antenna. Finally, a number of measurement results are given which apply to the usual rod-type ferrite-loaded loop antenna: measurement parameters cover a broad range of core lengths and diameters. It is found that typical ferrite-loaded loops have little electrical advantage over air loops although the packaging advantage of ferrite loops may be significant.

### 14.2. Ground Antenna Phase Behavior in a Differential Phase Measuring System

I. CARSWELL AND C. FLAMMER, *Stanford Research Institute, Menlo Park, Calif.*

A tracking system which measures distance by comparing the phase of signals arriving at a pair of spaced receiving antennas from a common source remains accurate when the signals have undergone identical phase distortions. The receiving antennas of such a system are normally mounted on a large ground plane to minimize multipath reflection errors at low elevation angles. This paper describes a study made of the effect of ground constants and ground plane size on the amplitude and phase of the electric field above a ground plane surrounded by lossy dielectric. Rapid phase shift near glancing incidence is found to be a function of elevation angle ground reflection coefficient and ground plane size. Results may be used to estimate tracking error in a differential phase measuring system.

### 14.3. Limits on the Information Obtainable from Antenna Systems

W. WHITE, *Airborne Instruments Lab., Mineola, N. Y.*

This paper starts with a review of the information obtainable from simple antennas such as radar dishes. The hopelessness of various super gain or super directivity proposals is demonstrated by an application of the sampling theorem. It is also shown that the information obtainable is profoundly affected by the *apriori* information available.

The paper continues to a study of more elaborate antenna systems. It is shown how the Mills Cross antenna used in radio astronomy is related to the separate beams of a V-beam radar or the use of separate search and height finding radars. Various interferometer techniques are also discussed and it is shown that there is an important difference between the radio astronomy case of incoherent radiation and the radar case with its coherent reflections.

### 14.4. Antenna Problems In Radio Astronomy

R. N. BRACEWELL, *Stanford University, Stanford, Calif.*

The design application of antennas with microsteradian beams such as the one under construction at Stanford University merit refined analysis of the type which has become familiar in the one-dimensional case in information theory.

Just as a waveform whose spectrum is band limited is fully represented by certain spaced samples, so an antenna array whose beam will be used to explore emitting or reflecting objects of limited extent may be whittled down to a certain lattice with no loss of information.

Array tapering raises questions analogous to optimum design of filters for fidelity and freedom from overshoot. A compromise between fineness of detail in the source or target and freedom from spurious effects due to side lobes is shown to depend on the spatial spectrum of the source and of the noise.

Beam-splitting techniques permit further whittling away of the antenna structure but introduce problems. Design details of the Stanford microwave spectroheliograph will be discussed.

### 14.5. High Altitude Breakdown Phenomena

J. ASHWELL, E. B. COLE, A. PRATT, AND D. SARTORIO, *Glenn L. Martin Co., Baltimore, Md.*

Breakdown problems associated with the operation of antennas at high altitudes and high speeds will be discussed, along with a presentation of some of the proposed solutions to these problems.

* Sponsored by the Professional Group on Antennas and Propagation. To be published in Part 1 of the 1957 IRE CONVENTION RECORD.

## SESSION 15*

**TUES. 10:00 A.M.–12:30 P.M.**

**New York Coliseum Marconi Hall**

### Information Theory—Review and Recent Advances

*Chairman:* D. SLEPIAN, *Bell Telephone Labs., Murray Hill, N. J.*

### 15.1. Cost of Transmission Reliability

R. M. FANO, *Massachusetts Institute of Technology, Cambridge, Mass.*

It has been shown that, contrary to earlier beliefs, the reliability of a communications system disturbed by noise could be made as perfect as desired for a fixed bandwidth, fixed signal power, and fixed transmission range; noise sets inherent limits to the transmission rate at which perfect reliability could be obtained, but not to the degree of the reliability. The cost of greater reliability is paid in terms of terminal equipment complexity and transmission delay. A digital communication system will be used to illustrate the dependence of reliability and its price on the rate of transmission in the channel characteristics. Possible future system development will be discussed, together with some of the theoretical and practical problems that must still be faced in that connection.

### 15.2. Channel Capacity without Coding

P. ELIAS, *Elec. Eng. Dept., Massachusetts Institute of Technology, Cambridge, Mass.*

Several communication systems employing feedback are discussed which illustrate an aspect of channel capacity and the trading relationship between the transmission rate, reliability, and delay without involving complicated coding and decoding procedure. These provide simple intuitive interpretations of some of the basic results in information theory.

### 15.3. Coding a Television Source

J. L. KELLY, JR., *Bell Telephone Labs., Inc., Murray Hill, N. J.*

The general problem of coding a continuous source for transmission over a discrete channel will be discussed. In particular, examples will be drawn from the field of television. Methods of estimating the rate of a continuous source will be outlined and the results of such estimates compared with known methods of transmission.

* Sponsored by the Professional Group on Information Theory. To be published in Part 2 of the 1957 IRE CONVENTION RECORD.

### 15.4. What Good Is Information Theory to Engineers?

J. R. Pierce, *Bell Telephone Labs., Inc., Murray Hill, N. J.*

Information theory has provided us with a new and precise language in which to talk about communication problems. Sometimes, fortunately, problems in this language give us a clearer understanding of them. It can scarcely be asserted, however, that information theory has led to startling advances in the way we do things.

From a broader point of view, information theory has given us a new way, full of inspiration and confusion, of guiding and evaluating experiments performed on human beings.

## SESSION 16*

**TUES. 10:00 A.M.–12:30 P.M.**

**NEW YORK COLISEUM FARADAY HALL**

**Microwave Tubes**

*Chairman:* V. Learned, *Sperry Gyroscope Co., Great Neck, N. Y.*

### 16.1. The Selection and Applications of Traveling-Wave Tubes

A. H. Nielsen and N. W. Hansen, *Federal Telecommunication Labs., Nutley, N. J.*

This paper studies the traveling-wave tube from the point of the user and describes the procedure in selecting the proper operating performance and circuit design most suited for his application.

The objectives of this paper are twofold: 1) To establish the requirements for the particular application in mind, the evaluation of traveling-wave tube characteristics as a function of this application, and to determine the tube characteristics to be utilized; and 2) to discuss the practical problems associated with designing and packaging a traveling-wave amplifier including power supply circuitry, solenoid design, and cooling.

Equipment will be demonstrated which utilizes the above techniques, showing how a systematic procedure in the selection and packaging of traveling-wave tubes is beneficial.

### 16.2. X-Band Traveling-Wave Tube Feedback Oscillator

V. G. Price, *General Electric Microwave Lab., Palo Alto, Calif.*

The principles of feedback amplifier design have been extended recently to the microwave region to provide stable oscillators capable of being tuned over a wide range. An $x$-band oscillator is described which utilizes a commercially-available traveling-wave tube amplifier with an external feedback circuit comprised of either a fixed or tunable two-part high-$Q$ cavity resonator. Experimental data on this oscillator are given which indicate the degree of frequency stability achievable on both a short- and long-term basis and, in addition, pushing and pulling information is given for various operating conditions. A comparison is made between the performance of an oscillator of this type constructed of commercially-available components and that of a similar type oscillator constructed with an ultra-stable power supply. Under ordinary laboratory conditions of vibration, ambient temperature, and pressure the first oscillator had a short time stability of one part in $10^6$ while the second had a stability of one part in $10^8$ (for periods of the order of 20 milliseconds).

### 16.3. A Light-Weight, Low-Level Traveling-Wave Tube Amplifier for S Band

J. O. Bramick, *Federal Telecommunication Labs., Nutley, N. J.*

The electrical and mechanical design of a low-power, all metal traveling-wave amplifier tube will be discussed.

The tube is capable of operating in the frequency range of 2 to 4 kmc at a power level of 100 milliwatts. The minimum gain at rated power output is 25 db. The tube has been miniaturized by locating the rf terminals at opposite ends of the tube, effecting a considerable reduction in the size and weight of both the tube and solenoid. Thus, by a rearranging of the rf terminals, a compact package is achieved in which the maximum diameter of the solenoid is 2 inches, and the combined weight of the solenoid and tube is 2 pounds, 10 ounces.

### 16.4. The Duplexer as a Means of Eliminating Interference from Nearby High-Power Radar Systems

I. Reingold and J. L. Carter, *Signal Corps Eng. Labs., Fort Monmouth, N. J.*

The problem of mutual interference between nearby radar systems is becoming increasingly serious. Conventional duplexers are not particularly effective in alleviating the interference problem between systems at the incident power levels in question. An experimental duplexer tube of unique design has been developed and evaluated in an attempt to provide a solution to this interference problem. Preliminary tests to determine performance under interference conditions were most promising.

### 16.5. A New Method for Modulating Electron Beams for Pulse Applications and Linear Amplitude Modulation Systems

G. M. W. Badger, *Eitel-McCullough, Inc., San Bruno, Calif.*

The various methods for modulating electron-beam tube devices are discussed. These include linear amplifier service beam voltage modulation and beam current modulation. The advantages and disadvantages of the methods are presented.

One of the ways in which a beam-type electron device can be beam-current modulated is by means of the modulating anode. The anode of a Pierce type gun is electrically isolated from the other elements of the tube so that a modulating potential can be applied. The electrical characteristics of this new modulating element are discussed, together with some of the outstanding construction and fabricating aspects.

The circuits for pulse and amplitude modulation application are covered. The results of considerable experience with various high power klystrons are presented in terms of overall linearity, efficiency phase shift and other experimental performance data.

The modulating anode principle is being applied in other useful areas such as pulse and rectifier tubes. These tubes are analyzed and performance data are presented.

### 16.6. Behavior of a Backward-Wave Oscillator with External Feedback

F. L. Vernon, Jr., *Hughes Aircraft Co., Culver City, Calif.*

An investigation of the effects of deliberately introduced external feedback on the operating characteristics of a backward-wave tube is made. Essentially, two methods are available for predicting these effects. These are discussed and applied to several cases. It is shown that within certain limits a desired tuning curve or a power curve can be synthesized by using the proper frequency characteristic in the feedback line. Using the same methods, it is shown that the effects on the frequency sensitivity with respect to voltage, current, and other external changes can be predicted in terms of external circuit characteristics. This is experimentally and theoretically examined for some practical cases.

## SESSION 17*

**TUES. 2:30–5:00 P.M.**

**WALDORF-ASTORIA STARLIGHT ROOF**

**General Communications Systems**

*Chairman:* R. D. Campbell, *American Telephone and Telegraph Co., New York, N. Y.*

### 17.1. SSB Modulation for Scatter Propagation

I. H. Gerks and R. P. Decker, *Collins Radio Co., Cedar Rapids, Iowa*

There are many instances in scatter propagation in which, because of great path length, limitations on transmitter power, or restrictions on antenna size it is impractical to achieve sufficient signal-to-noise ratio at the receiver for effective use of frequency modulation. An example is the case of air-ground communication. In this case single-sideband modulation is the optimum method for providing at least one voice channel. Certain problems arise, however, when SSB modulation is used on vhf or uhf scatter circuits. Among these are amplifier

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* Sponsored by the Professional Group on Communications Systems. To be published in Part 8 of the 1957 IRE Convention Record

linearity, carrier frequency stability, effective agc action, and diversity combining. This paper considers practical solutions to these problems and describes results secured over a 1070-mile ionospheric scatter circuit at 38.6 mc.

### 17.2. FM Scatter System Measurements

R. W. Britton and H. D. Hern, *Collins Radio Co., Cedar Rapids, Iowa*

The design of tropospheric scatter communication systems is characterized by the considerations given to the effects of random fading and high propagation loss. To insure equipment performance that is compatible with these stringent design considerations, various comprehensive system measurements must be performed. These measurements are discussed and specific techniques presented with emphasis on diversity combining tests and intermodulation distortion using a random noise loading method. Measured results on one hundred systems is presented in graphical form and compared to the required design values of a typical twenty-four-voice channel system.

### 17.3. Theory of Feedback Around the Limiter

E. J. Baghdady, *Dept. of Elec. Eng. and Research Lab. of Electronics, Massachusetts Institute of Technology, Cambridge, Mass.*

A new theory for the effect of feedback around the limiter upon interference rejection in fm receivers considers interference resulting from superposition of two carriers differing in amplitude and frequency, and distinguishes between ***wide-band*** and ***narrow-band*** feedback. The theory shows that wide-band feedback cannot reduce the magnitude of the frequency spikes caused by the interference regardless of the feedback angle. Inverse narrow-band feedback aggravates the disturbance and raises the limiting threshold. Positive narrow-band feedback, which causes self-oscillation in the absence of input signal, results in pronounced capture improvement and automatic noise squelch, and lowers the limiting threshold.

### 17.4. General Systems Approaches to Telecommunication Optimization Problems

R. E. Kalaba and M. L. Juncosa, *The RAND Corp., Santa Monica, Calif.*

In the past, for telecommunication networks, it has been difficult to treat analytically system problems of optimal design and utilization. The large number of variables involved in describing the various system capacities, demands of the users, and economic factors have forced investigators to rely on various sub-optimization techniques.

By contrast, two recently developed mathematical techniques, ***linear programming*** and ***dynamic programming***, enable one to treat problems—involving perhaps thousands of variables—at the system level in a straightforward and computationally feasible manner. Two telecommunication network extension problems are treated in some detail using these methods. Other applications are indicated.

### 17.5. The Future Air Force Communication System

C. K. Chappuis, *Rome Air Development Center, Griffiss Air Force Base, N. Y.*

The expansion of air communication requirements has proceeded with fantastic rapidity in the span of the last half century. Most of this expansion has occurred during and since World War II. The end of this expansion in numbers of channels and particularly in capabilities is not in sight. It must increase with higher aircraft and missile speeds, and increased ranges and heights. Also, the advent of space satellites will add to the communication problems The adaptation of the most modern civilian communication system techniques and engineering to Air Force purposes is mandatory as a basic premise. Immediate expansion and improvements to meet the Air Force requirements must follow rapidly.

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## SESSION 18*

**TUES. 2:30–5:00 P.M.**

**WALDORF-ASTORIA ASTOR GALLERY**

**Medical Electronics**

*Chairman:* R. L. Butenhoff, *Atomic Energy Commission, Washington, D. C.*

### 18.1. A Magnetic Tape Recording System for Teaching Electrocardiography

G. N. Webb and W. R. Milnor, *Johns Hopkins Hospital, Baltimore, Md.*

A system for recording ecg and other physiological data in the frequency range of 0.10 to 100 cps on magnetic tape is described. Tests of various grades of tape-handling assemblies, modulation, and demodulating equipment are shown and discussed with respect to the parameters of the ecg desirable for teaching purposes.

Two channels of information are recorded on quarter-inch tape using frequency modulated phantastrons. Playback amplifiers and demodulators are transistorized. Complete circuits of modulators, demodulators, and regulated power supplies are shown. Noise, amplitude, dynamic range, and linearity measurements are illustrated.

### 18.2. Ultrasonic Dosimetry for Medical Use

V. T. Tomberg, *Biophysical Research Lab., New York, N. Y.*

Ultrasonics are used for medical treatment in physical therapy The margin between harmful doses and not successful doses is very narrow Dosimetry is difficult to apply because of inadequate electronic construction and insufficient knowledge of the overlapping biological effects of heat, mechanical and chemical nature. The physical and technical problems are discussed.

### 18.3. An Electrical Circuit Analogy for Isotope Kinetics

R. L. Schoenfeld and M. Berman, *Sloan-Kettering Institute, New York, N. Y.*

An electrical circuit analogy is developed to treat the kinetics of isotopic tracers in steady-state physiological systems. The circuit consists of $n$ capacitors connected by non-reciprocal conductances. This structure is analogous to the distribution and steady-state rate of flow of the nonlabeled material. Network functions analogous to driving point and transfer impedances for the electrical circuit are defined from measurements of the kinetics of the traces. Methods carried over from electrical network synthesis permit the formulation of possible models and the mapping of equivalent systems.

### 18.4. Electronic Control of the Artificial Respiration Cycle

L. H. Montgomery, *Vanderbilt University Medical School, Nashville, Tenn.*

A method of using biological potentials, which are amplified, filtered, and integrated to automatically control the flow of air to and from the iron lung or other type of respirator in accordance with the needs of the patient, is described. Preliminary operational results are discussed.

### 18.5. The Incorporation of Picture Storage with the Technique of X-Ray Fluoroscopy

W. Gombash, Jr., *Hughes Aircraft Co., Culver City, Calif.*

This paper deals with the application of a half-tone storage tube for the retention of X-ray fluoroscopic images as received from an X-ray phosphor screen. Using this technique it is possible for the viewing radiologist to obtain a brighter more defined image, while at the same time cutting down on the total dosage of X-ray to the patient. This is accomplished by picking up a short exposure, relatively high intensity X-ray fluoroscopic image with a television camera tube, feeding the video signal through a tv chain and displaying and storing the resultant image on a half-tone direct viewing storage tube for subsequent study.

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## SESSION 19*

**TUES. 2:30–5:00 P.M.**

**WALDORF-ASTORIA JADE ROOM**

**New Operational Techniques Concerning Video Test Signals**

* Sponsored by the Professional Group on Medical Electronics To be published in Part 4 of the 1957 IRE Convention Record.

* Sponsored by the Professional Group on Broadcast Transmission Systems. To be published in Part 7 of the 1957 IRE Convention Record.

*Chairman:* R. N. HARMON, *Westinghouse Broadcasting Co., Inc., New York, N. Y.*

*Panel Members:* J. R. POPKIN-CLURMAN AND F. DAVIDOFF, *Telechrome Manufacturing Corp., Amityville, N. Y.*
J. W. WENTWORTH, *Radio Corp. of America, Camden, N. J.*
R. M. MORRIS, *American Broadcasting Co., New York, N. Y.*
W. B. WHALLEY, *Columbia Broadcasting System, New York, N. Y.*
H. C. GRONBERG, *National Broadcasting Co., New York, N. Y.*
J. THORPE, *American Tel. and Tel. Co., New York, N. Y.*
E. W. CHAPIN, *Federal Communications Commission, Laurel, Md.*
A. ST. MARIE, *Canadian Broadcasting Corp., Montreal, Que., Can.*

This panel discussion deals with the subject of video test signals during the vertical blanking interval. This is a new operational technique which permits the transmission of certain video test signals on a continuous basis during normal program transmission. A standardization of test signals for transmission during the vertical interval has not as yet been made by the in dustry or authorized by the FCC except on an experimental basis. An effort has been made to obtain speakers of authority on this subject. The speakers include representatives from two equipment manufacturers the three major television networks, The Telephone Long Lines Department, and the FCC. A question and discussion period will follow.

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# SESSION 20*

**TUES. 2:30–5:00 P.M.**

## WALDORF-ASTORIA SERT ROOM

## High Fidelity and Home Measurements

*Chairman:* F. H. SLAYMAKER, *Stromberg-Carlson Co., Rochester, N. Y.*

### 20.1. Intermodulation Distortion: Its Measurement and Evaluation

A. PETERSON, *General Radio Co., Cambridge, Mass.*

The unlimited variety of methods of intermodulation distortion measurement are conveniently classified into three major groups according to the number of components in the test signal. At least seven subgroups are sufficiently distinct to merit individual discussion. Since these test methods do not, in general, yield the same information, one or more of them should be selected for each application according to the nature and expected uses of the device under test.

Some recent psychoacoustic investigations are used for judging the significance of the results of intermodulation measurements in terms of audibility of distortion in high-quality reproduction.

### 20.2. Testing High-Fidelity Amplifiers in the Home

W. W. DEAN, *General Electric Co., Syracuse, N. Y.*

Explicit methods of measurement will be described in order to prevent falling into the pitfalls of false or misleading data. Procedures will be outlined for the measurement of frequency response, power output, equalization, preamplifier overload, sufficiency of gain on the phonograph channels, and residual hum and noise.

### 20.3. Disk and Magnetic Tape Phonograph Systems

W. H. ERIKSON, *Radio Corp. of America, Camden, N. J.*

Measurement techniques are described with the objective of performance improvement. Major emphasis is on the thorough evaluation of pickup cartridge performance and living room acoustical measurements on loudspeakers. Pickup measurements include optimum vertical force, low and high-frequency tracking capabilities, IM distortion, and frequency response. The most useful test records and tapes are identified and described. The difficulties of living room loudspeaker measurements are outlined and a measurement technique which minimizes these difficulties is described in detail. Methods of improving loudspeaker performance by equalization are discussed and illustrated. Many performance curves of pickups and loudspeakers are shown.

### 20.4. Improved Low-Frequency Loudspeaker Performance

S. ZUERKER, *General Electric Co., Syracuse, N. Y.*

Present-day home audio systems are usually adequate with respect to frequency response, but improvement of undistorted power output at low frequencies is desirable. The loudspeaker is the weakest element in the system in this respect, although improvements also should be made on power amplifier stability. Analyzing the distorting elements in the loudspeaker and the power requirements, it can be shown that a loudspeaker can be designed to adequately handle all low-frequency power conditions without a significant drop in efficiency, and mounted within an enclosure of smaller than usual dimensions. The gain in maximum power output will be shown in comparison with popular loudspeakers now on the high-fidelity market by means of power response curves.

### 20.5. A Low-Pressure Phonograph Cartridge

W. E. GLENN, *General Electric Co., Schenectady, N. Y.*

A barium titanate phonograph cartridge is described which is designed to meet tracking requirements below 2 grams tracking pressure. The cartridge has a dynamic mass of only 1/10 of a milligram and is designed so that it has no resonances between 25 cycles and 20,000 cycles. The output after compensation for the RIAA record characteristics is about 40 millivolts. Hum and noise are particularly low with this unit. The vertical and lateral compliance, dynamic mass, protection from damage. simplicity of construction, and other factors are considered in the design Construction and performance characteristics are given

### 20.6. A High-Fidelity Phonograph Reproducer

B. B. BAUER AND L. GUNTER, *Shure Brothers, Inc., Evanston, Ill.*

A new phonograph reproducer is described combining together certain features believed to be essential for high-fidelity reproduction. They are: 1) Practical elimination of friction by use of jewel bearings on all pivots; 2) low arm mass combined with freedom from resonances; 3) damping without interference with the freedom of motion of the arm; 4) tracking error correction; and 5) safe needle lift mechanism to avoid damage to the needle and the record. The cartridge is of moving field type with needle tip mass approaching 1 milligram and is magnetically balanced against hum pickup. The reproducer is capable of tracking conventional records with needle load less than two grams.

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# SESSION 21*

**TUES. 2:30–5:00 P.M.**

## NEW YORK COLISEUM MORSE HALL

## Antennas II—Broadband Antennas

*Chairman:* A. ALFORD, *Andrew Alford Consulting Engineers, Boston, Mass.*

### 21.1. High Frequency Steerable Beam Antenna System

E. HUDOCK, *Collins Radio Co., Cedar Rapids, Iowa*

A high-frequency antenna system whose characteristics are those desirable for reliable long-path propagation in the high-frequency range and whose beam is steerable in the azimuth plane is described. High gain at low angles of radiation with patterns containing a minimum of secondary lobes are the chief features of this antenna. The steerable system discussed here is of the Wullenweber type, but of interest is the nature by which two separate and independent arrays are combined in one supporting structure and the wide range of frequencies that each array covers. The lower frequency array is vertically polarized, while that of the higher frequency array is horizontal. Each array has a gain comparable to that of

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a class A rhombic and maintains a satisfactory impedance match (vswr 2:1 and under) over a bandwidth of 2:1.

### 21.2. Constant Beamwidth Broadband Antennas

C. F. PARKER AND R. J. ANDERSON, *Melpar, Inc., Falls Church, Va.*

In a normal horn radiator or aperture type antenna the beamwidth of the main lobe is inversely proportional to the aperture in wavelengths. As frequency is varied the aperture /wavelength ratio varies in direct proportion, providing a one-to-one correspondence between frequency and beamwidth. A new type horn radiator called the "pinwall" antenna has been developed which provides an essentially constant beamwidth over a 4-to-1 band instead of the usual 4-to-1 variation in beamwidth that would result from this frequency change. The basic principle behind the pinwall antenna lies in the replacement of the walls of a horn with pins properly spaced to allow the higher frequencies to radiate from a small aperture while constraining the lower frequencies until larger apertures are reached. End-fire radiators were also investigated and dimensions have been obtained for polyrod antennas which produce a constant beamwidth over a 2-to-1 band.

### 21.3. Broadband Traveling-Wave Antennas

D. K. REYNOLDS, *University of Seattle, Seattle, Wash.*

The large class of traveling-wave antennas includes many examples which are superficially quite different, such as the helical antenna, the dielectric antenna, the corrugated surface, and the long Yagi antenna. The essential similarity of these antennas is discussed, and it is shown that the pattern bandwidth is inherently large, and separable from the impedance bandwidth. Broadband methods of excitation are discussed which allow for operation over large frequency ratios. An example is presented of the broadband excitation of a 15 wavelength Yagi antenna.

### 21.4. Evaluating the Impedance Broadbanding Potential of Antennas

A. VASSILIADIS AND R. L. TANNER, *Stanford Research Institute, Palo Alto, Calif.*

One of the principal concerns of the aircraft antenna designer is to determine whether an antenna having some known terminal impedance characteristics can be compensated to meet a specified vswr-bandwidth relation. Alternatively the problem may be to select from among possible antenna configurations that one which possesses the greatest impedance broadbanding potential.

In this paper the problem is examined by means of modern circuit theory. Antenna impedances are approximated by rational algebraic fractions and the impedances represented by these functions are treated analytically to determine the vswr-bandwidth relationship which may be obtained by optimum compensation. The functions which can be considered by the techniques developed are sufficiently complex to adequately represent the basic impedance characteristics obtained with typical airborne flush mounted antennas over a frequency range which causes the dimensions of the antenna to vary between a small fraction of a wavelength and approximately a wavelength. Examples are given.

### 21.5(A). Frequency Independent Antennas

V. H. RUMSEY, *University of Illinois, Urbana, Ill.*

Any antenna whose shape is specified entirely by angles obviously is independent of the wavelenth. The infinite biconical antenna is a familiar example There is an infinite variety of such frequency independent shapes but they all extend to infinity. The practical application of this idea therefore depends on how rapidly, if at all, the performance of a finite antenna of this kind converges to the infinite case as the frequency is raised. It appears that the input impedance always converges but that the pattern usually does not converge. There are some striking examples where the pattern does converge, thus giving a practical antenna whose pattern as well as impedance is independent of frequency above a certain frequency. The accompanying paper by DuHamel and Isbell is an example.

### 21.5(B). Broadband Logarithmically Periodic Antenna Structures

R. H. DUHAMEL, *Collins Radio Co., Cedar Rapids, Iowa,* AND D. E. ISBELL, *University of Illinois, Urbana, Ill.*

Antenna structures for which the input impedance and radiation pattern vary periodically with the logarithm of the frequency are described. For a particular class of these structures the variation of the electrical characteristics over a period is negligible, the result being an antenna for which the impedance and pattern are essentially independent of frequency over bandwidths greater than ten to one. The antennas are linearly polarized and bidirectional beams with equal principal plane beamwidths are obtained The beamwidth may be controlled to a considerable extent by the geometry of the structure.

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## SESSION 22*

**TUES. 2:30–5:00 P.M.**

**NEW YORK COLISEUM MARCONI HALL**

**Information Theory—Applications**

*Chairman:* M. J. DI TORO, *Polytechnic Research and Development Co., Inc., Brooklyn, N. Y.*

### 22.1. An Inductive Inference Machine

R. J. SOLOMONOFF, *Technical Research Group, New York, N. Y.*

A study has been made of a category of machines that will perform inductive inferences. A simplified model will be described that only uses a few of a more complete set of heuristic devices. Such a machine is able to learn arithmetic operations from a small set of examples.

By using all of the heuristic devices, it is expected that a machine will be able to learn to perform complex tasks for which it was not specifically designed. Proving theorems, playing good chess, and answering questions in English appear to be within ultimate machine capabilities.

### 22.2. Multicase Binary Codes for Nonuniform Character Distributions

F. P. BROOKS, JR., *International Business Machines Corp., Poughkeepsie, N. Y.*

For economical transmission, variable-length coding systems are theoretically best for nonuniform character distributions, but difficulties of decoding and checking limit their practical usefulness. A useful class of variable length codes are the multicase codes, in which the characters (messages) to be represented are divided into two or more cases, with shift characters signaling changes in the case of succeeding characters. Analysis of the properties of multicase coding systems yields a method of generating and selecting economical variable-length codes that can be readily encoded, decoded, and checked.

### 22.3. Binary Transmission Through Noise and Fading

M. MASONSON, *Federal Telecommunication Labs., Nutley, N. J.*

This paper deals with the detection of binary modulated carriers, *e.g.* "on-off" radioteletype, fsk, etc., in the presence of noise and fading. Diversification in transmission and reception and "combining" methods are dealt with. System performance is presented in terms of probabilities of error in reception. System capabilities are presented in terms of their (channel) capacities. Various detection (decision) criteria are considered, *e.g.*, minimization of error probability.

### 22.4. An Estimate of the Degradation in Signal Detection Resulting from the Addition of the Video Voltages from Two Radar Receivers

H. L. McCORD, *Hughes Aircraft Co., Culver City, Calif.*

The detection of a pulse radar signal in the video of a radar receiver is treated as a problem in statistics for three cases: 1) a reference case of one radar, 2) a case where the videos of two radars are added linearly, and 3) the case where only the highest video voltage between the two radars at each instant of time is presented to the detection process. The signal-to-noise power ratio required in the IF of the chan-

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nel containing the signal is computed for a probability of detection of 0.50 and a false alarm probability of $1.00 \times 10^{-8}$. This signal-to noise power ratio for case 2) and case 3) is compared to that of case 1) the reference case, and estimates of the equivalent degradation in transmitted power result. The estimated degradation for the linearly added case is 2.2 db, while that for the peak selected case 3). is 0.20 db within the limits of the assumptions used in the analysis.

### 22.5. The X Carrier: A Telephone Carrier System Employing Bandwidth Compression

J. W. HALINA, *Lynch Carrier Systems, Inc., San Francisco, Calif.*

The paper deals with a carrier system for telephone application which employs a method of speech bandwidth compression described as "coherent modulation." Compression is achieved by modulating the speech spectrum with a function of itself.

Both theoretical results and experimental data are reported for prototypes designed for 3-to-1 and 5-to-1 bandwidth compression ratios. The quality of speech transmitted through the system is surprisingly high. Conventional single tone tests for bandwidth and distortion do not detect the presence of a bandwidth compressor in the system and are thus relatively irrelevant as a measure of performance The system exhibits an improvement threshold which is apparently related to a restricted autocorrelation process employed in coherent modulation.

The interesting characteristics of the equipment are economy of parts, simplicity and utilization of conventional circuitry. Development to date has not reached the point of determining the maximum compression ratios realizable.

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## SESSION 23*

**TUES. 2:30–5:00 P.M.**

**NEW YORK COLISEUM FARADAY HALL**

**Televisual Systems Devices**

*Chairman:* S. T. SMITH, *Naval Research Lab., Washington, D. C.*

### 23.1. The Development of 110° Television Picture Tubes Having an Ion Trap Electron Gun

L. E. SWEDLUND AND L. C. WIMPEE, *General Electric Co., Syracuse, N. Y.*

The factors leading to the development of 110° deflection angle, television picture tubes are reviewed and in particular the problems of designing an electron gun with an ion trap in the smaller, shorter neck of these tubes, are described. In spite of the smaller space available for the electron gun a design was devised which provides full focused beam current and adequate voltage insulation. Although, because of the restricted space. there appeared to be no simple way to provide for an ion trap the problem was solved by means of a novel off-set grid, ion trap structure which centers the focused beam on the screen without requiring that the electron gun be tilted.

The solution of these problems is illustrated by their application to commercial picture tubes.

### 23.2. Image Tube Utilizing Bombardment Induced Conductivity

R. W. DECKER AND R. J. SCHNEEBERGER, *Westinghouse Research Labs., Pittsburgh, Pa.*

A television pickup device utilizing a layer of material exhibiting electron bombardment-induced conductivity as the amplifying and storing element is described The device operates in the simplest type of television camera systems The performance of present experimental tubes indicates ultimate sensitivity comparable to the Image Orthicon It is expected that sensitivity can be increased to the theoretical limits imposed by the efficiency of the photocathode, with resolution adequate for television broadcasting.

### 23.3. New Developments in the Panel Light Amplifier

B. KAZAN, *Radio Corp. of America Labs., Princeton, N. J.*

Using a new electronic system of operation and improved photoconductive materials, an experimental large-area, panel-type, solid-state light amplifier has been developed with light gains of as much as 1000 times. When directly excited by X rays instead of light, such panels produce output pictures 100 times brighter than the conventional fluoroscope screen. The slow decay of the amplifier makes the screen potentially useful for large-screen, bright radar displays, and for the storage of transient images. The nonlinear component materials increase the contrast of the output image, the particular input-output characteristic of the amplifier being determined by the selected mode of operation and the operating frequency.

### 23.4. An Electrostatic Character-Writing Tube

K. SCHLESINGER, B. MAGGOS, AND A. F. HOGG, *Motorola, Inc., Chicago, Ill.*

A 19-inch charactron tube has been developed which uses electrostatic fields throughout for character selection and distribution, as well as for image formation and intensification. Two pairs of conjugate electrostatic yokes are used for letter-selection, and a fifth unit for character-positioning at wide angles (70 degrees).

A long decelerating electrostatic lens is used for image formation, permitting instantaneous change of size between small and capital letters.

A new barrier-mask accelerator has been developed to increase beam velocity twelve-to-one over the last inch of its path ahead of the screen.

### 23.5. A High-Speed, Low Voltage Light Modulator

A. C. KOELSCH, *IBM Research Center, Poughkeepsie, N. Y.*

Barium titanate when held above the Curie temperature exhibits double refraction, upon the application of an electric field, that disappears when the field is removed This Kerr effect is significant in barium titanate for applied fields well below those required for common Kerr cell materials, such as nitrobenzene. This paper will discuss double refraction and the Kerr effect. and in particular, the desirability of operating barium titanate, as an electro-optic light modulator, above the Curie temperature. Experimental results show that light pulses can be attained whose duration are in the low musec region. The effect of mechanical crystal vibrations will be discussed, as well as the light transmission properties of the system.

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## SESSION 24*

**TUES. 8:00–10:30 P.M.**

**WALDORF-ASTORIA STARLIGHT ROOF**

**Applications of Electronics to Air Traffic Control**

*Chairman:* W. R. G. BAKER, *Pres., Radio Electronics Television Manufacturers Association*

*Moderator:* J. L. ANAST, *Systems Planning Advisor to E. P. Curtis, Ass't to the President of of the United States for Aviation Facilities Planning*

*Panel Members:* S. ALEXANDER, *National Bureau of Standards, Washington, D. C.*
C. WHEELER, *Airborne Instruments Lab., Inc., Mineola, N. Y.*
L. PERPER, *Tucson, Ariz.*
V. WEIHE, *Air Transport Association, Washington, D. C.*
S. BERKOWITZ, *The Franklin Institute, Philadelphia, Pa.*

A discussion of plans, research, and development on national aviation air traffic control facilities and systems and the use of electronics in the program.

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* Sponsored by the Professional Group on Aeronautical and Navigational Electronics and Military Electronics. To be published in Part 8 of the 1957 IRE CONVENTION RECORD.

# SESSION 25*

**TUES. 8:00–10:30 P.M.**

**NEW YORK COLISEUM FARADAY HALL**

## Microminiaturization—The Ultimate Technique

*Chairman:* C. BRUNETTI, *General Mills, Inc., Minneapolis, Minn.*

The cry "make it smaller—make it lighter" has been ringing in the radio-electronics engineer's ears since before World War I and it seems that the more he accomplishes along this line, the louder and more insistent the cry becomes. Much, indeed very much was accomplished in the years between 1918 and 1941 but World War II multiplied the demand for "smaller and lighter" by several orders of magnitude. Actually, in far too many cases the pressure for decreased size and weight was too great and reliability was neglected or even lost sight of. Today, therefore, the modifier " . . . but make it more reliable," has been added to the original cry.

To the modern electronics engineer, this is not at all inconsistent, but is reasonably essential if progress in his chosen field is to continue at its accelerating rate presently observable. He was able very creditably to miniaturize his equipment prior to World War II, to subminiaturize during and after the Korean "police action," but today this is not enough; he must microminiaturize if he's to stay in the running. Some very substantial steps in microminiaturization have taken place and many more are in the offing, sometimes not even recognized as such, just as so many revolutionary developments arrive piecemeal and are not appreciated until someone takes time to review the field and point out the effort that is taking place and, surprisingly perhaps, is intelligently coordinated.

This session, therefore, is to spotlight some of the "microminiaturization" that is now with us or under way and to pose a challenge for engineers in the electronics field to do even better in the future. Can he go beyond "microminiaturization" or is this truly "The Ultimate Technique"? Only time will tell, but from one viewpoint, if an item can be reduced substantially in size and/or weight, then it's still in the miniature or subminiature stage and "Microminiaturization—The Ultimate Technique" is yet to arrive.

### 25.1. Philosophy of Microminiaturization Technique

C. BRUNETTI, *General Mills, Inc., Minneapolis, Minn.*

The concept of microminiaturization will be introduced and presented in relation to miniature and subminiature techniques. The increasing need for extension of the art of electronic packaging into the smaller regions will be covered. Guide lines, necessary for development and fabrication of useful and practical microminiature electronics, will be described in general terms.

Some new developments in sciences outside the electronic field, which might encourage new ideas or approaches to microminiaturization, will be presented.

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### 25.2. The Challenge of the Environment

E. F. CARTER, *Stanford Research Institute, Menlo Park, Calif.*

This talk will show the urgent and basic needs for microminiaturization in order to keep up with the rapid progress in electronic applications, particularly in the mobile field. Much imaginative publicity has been given to the high temperature requirements of equipment for missiles and supersonic aircraft and these will be covered authoritatively but there are other fields such as the infantryman's helmet transceiver that are perhaps equally important where the space requirements cannot even be approached without microminiaturization.

### 25.3. Material Constituents and Components

H. A. STONE, JR., *Bell Telephone Labs., Murray Hill, N. J.*

This talk will cover as much of the specific field currently available in the way of new materials and techniques. Available components will be shown and discussed and an indication of what further progress may be expected within the next year or so. Needs for additional materials and new types of components not yet available or on the horizon will be stressed as a challenge for the electronics industry to come forward with a practical solution.

### 25.4. Today's Applications and Equipments

End product uses of microminiature components will be illustrated both by actual equipment and by slides and/or discussion. Speakers will discuss specific fields and items both to show what is now being done and to offer a provocative challenge to others to do even better in the future.

#### Military

BRIG. GEN. E. R. PETZING (Ret), *Moore School of Elec. Eng., University of Pennsylvania, Philadelphia, Pa.*

#### Missiles

J. R. MOORE, *North American Aviation Corp., Downey, Calif.*

The guided missile has provided a primary impetus to microminiaturization. This talk will explain how the necessity of replacing human pilots, navigators, and bombardiers with electromechanical counterparts to make guided missiles has forced a concentration on reliability and reduction in size and weight which can only be achieved by microminiaturization. Because missiles used in military action are completely expended, the reduced cost in airframe weight and propulsion equipment afforded by a minimal size and weight of electronic brains and reflexes has an important effect in achieving "more defense per dollar." Added to this is the fact that microminiaturization permits semi-automatic or automatic production techniques which further reduce the cost of missiles by increasing reliability, reducing the cost of electronic elements, and permitting national standardization for mobilization purposes

#### Methods

*W. W. Hamilton, Elgin National Watch Co., Elgin, Ill.*

How does one go about "microminiaturization?" Probably the most significant progress in the broad field has been accomplished by the watch industry. This talk covers the application of highly precise watch design and manufacturing methods to frontier electronic technology.

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# SESSION 26*

**WED. 10:00 A.M.–12:30 P.M.**

**WALDORF-ASTORIA STARLIGHT ROOF**

## Electronic Computers I—Digital Computers

*Chairman:* J. P. NASH, *University of Illinois, Urbana, Ill.*

### 26.1. An RCA High-Performance Tape Transport System

S. BAYBICK AND R. E. MONTIJO, *Radio Corp. of America, Camden, N. J.*

A high-performance, multichannel digital tape transport has been developed to meet the needs of the data processing industry in general. This is a tubeless equipment which provides very fast start and stop times at very high repetition rates through the use of semiconductor and magnetic components.

This paper describes the electronics and mechanism in detail including the methods employed in obtaining start-stop rates to 120 per second, start and stop times of less than 2 milliseconds, and a start-stop spacing of less than 0.2 inches. The transport handles various widths of tape from $\frac{1}{2}$ to $1\frac{1}{8}$ inches and magnetic heads which provide up to 18 recording tracks.

The weight-type reel servo system, the high-current transistor solenoid driver, and the tape control logic are also described. Several unique features of this equipment have been employed in this developmental equipment to achieve performance and enhance its reproducibility.

### 26.2. A Magnetic Pulse-Current Regulator

J. D. LAWRENCE, JR. AND T. H. BONN, *Remington Rand Univac, Philadelphia, Pa.*

This paper describes a magnetic current regulator using a square hysteresis loop core with two windings. A direct current flowing in one winding holds the core in a saturated condition and provides a current reference for regulation. The core presents a low impedance to a current pulse passing through the other winding until the pulse mmf exceeds the dc mmf. When this happens, the core moves into

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the unsaturated region of its hysteresis loop and presents a high impedance to the current pulse, thereby preventing further increase in this current pulse.

Design considerations are presented. The precision of regulation is directly proportional to the number of turns on the pulse winding and inversely proportional to the magnetizing mmf required by the core in its unsaturated region. Hence, precision of regulation is limited by the amount of air core inductance that can be tolerated and the number of turns that it is physically possible to place on a core. The use of a thin-wall metal bobbin substantially improves the regulation.

### 26.3. Diodeless Magnetic Core Logical Circuits

L. A. Russell, *IBM Research Center, Poughkeepsie, N. Y.*

Magnetic cores having rectangular hysteresis loops have been shown to be useful as a key element in logical circuits for digital computers. However, most of these circuits use diodes for coupling, requiring windings of large numbers of turns, or large cores. A special class of magnetic core logical circuits will be presented in which cores are used for coupling. By avoiding the use of diodes, these circuits offer major advantages in economy, reliability, and compactness in medium speed applications. The basic technique that eliminates the need for diodes requires that some of the cores have a switching threshold. Specific examples of possible circuits and their operating characteristics will be shown.

### 26.4. Digital Computer Designs Circuit for Longest Mean Time to Failure

J. Alman, P. L. Phipps, and D. L. Wilson, *Remington Rand Univac, Div. of Sperry Rand Corp., St. Paul, Minn.*

This paper describes a system of circuit design utilizing a digital computer. The digital computer is programmed in such a manner so that it can compute the circuit that would give the longest mean time for failure of the circuit. This is accomplished by programming into the computer the characteristics of life test of components. The computer computes the circuit many times looking for the combination of circuit components that would still meet the output requirements and give the longest mean life to failure. The main computer output then becomes just the circuit which would give the longest mean time to failure possible with the existing component characteristics which were programmed into the computer.

### 26.5. Considerations in the Design of Character Recognition Systems

E. C. Greanias and Y. M. Hill, *International Business Machines Corp., Endicott, N. Y.*

The basic factors pertaining to the Character Sensing problem are discussed. These are range of style, range of quality, and number of characters and symbols to be recognized. Possible definitions of character quality and units for measuring it are described. The information handling steps of the recognition process are outlined as:

1) Conversion of marks on paper to electrical signals.
2) Discrimination between signal and noise.
3) Reduction of filtered data.
4) Identification of characters and symbols based on reduced data.
5) Validity checks.

The use of digital computers in the simulation of data reduction and recognition are described. A method for generating realistic test specimens on a digital computer is outlined.

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## SESSION 27*

**WED. 10:00 A.M.–12:30 P.M.**

**WALDORF-ASTORIA ASTOR GALLERY**

**Magnetic Recording**

*Chairman:* M. Camras, *Armour Research Foundation, Chicago, Ill.*

### 27.1. An Approach to Quantitative Methods for Evaluation of Magnetic Recording System Performance

C. B. Stanley, *Ampex Corp., Redwood City, Calif.*

The selection of an information storage device appropriate to the form and precision of the data analog requires evaluation of the distortions introduced by the system in terms of the critical data parameters.

The effects of nonuniform tape motion, dynamic range, and frequency response upon direct-, fm-, and PWM-recorded data are discussed. Methods are presented for exhibiting system performance specifications in forms suitable for directly selecting the optimum type of translation appropriate to the accuracy and bandwidth requirements of the data.

### 27.2 The Application of Wow and Flutter Compensation Techniques to FM Magnetic Recording Systems

R. L. Peshel, *Ampex Corp., Redwood City, Calif.*

The sources of error present when data signals are recorded and reproduced on magnetic tape by an fm system are identified, and the effect of each type of error on the reproduced intelligence is analyzed.

Several of the methods employed for the compensation of errors arising from unstable tape motion are compared on the basis of effectiveness, complexity, and cost. The problems associated with the design of a compensation system are reviewed.

The relative merits of two commonly used systems are investigated in detail.

### 27.3. Design of Instrumentation Magnetic Tape Transport Mechanisms

K. Schoebel, *Ampex Corp., Redwood City, Calif.*

The various existing instrumentation tape recorders for use in the field of data reduction are described. The basic design of a new precision magnetic tape transport mechanism is presented in detail from the development of a prototype to the full production. It is demonstrated how mechanical design can reduce wow and flutter to an extremely low value and in many cases eliminate the need of electronic compensation. A complete description of the tape transport mechanism is presented—size, capability, and performance. Modifications of the standard tape transport, such as a 120 inches/sec search-drive, a multispeed drive system, and others are described.

### 27.4. The Reverbetron

P. C. Goldmark and J. M. Hollywood, *CBS Labs., New York, N. Y.*

A magnetic tape reverbetron is described. This device for producing synthetic reverberation has been developed into models suitable for studio or home use. A brief historical outline of the course of development is given.

The subjective effects of reverberation time, percentage of reverberation, and echo repetition rate are discussed. The choice of range for these parameters is given. Design considerations are discussed. A block diagram and schematic are shown, and pictures of studio and home models. A demonstration of the reverbetron unit is given.

Comparison is made with a typical echo chamber by showing comparative frequency response and oscillograms of transient response.

### 27.5. A Multichannel Transducer for Magnetic Recording

H. A. Johnson, *Shure Brothers, Inc., Evanston, Ill.*

This paper describes an unique method of construction for a magnetic transducer element for magnetic recording and reproducing equipment and a stacking means whereby a plurality of elements can be utilized for multichannel applications. Where two or more elements are juxtaposed, such as in stereophonic recording and similar applications, crosstalk considerations become rather significant. In order to arrive at satisfactory crosstalk figures it is necessary to consider two sets of conditions: 1) when only one channel is being recorded at a particular time, and 2) when several channels are being recorded simultaneously. The results obtained from the crosstalk measurements, along with the frequency response characteristic and operating parameters for the elements, are presented.

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* Sponsored by the Professional Group on Audio. To be published in Part 7 of the 1957 IRE Convention Record.

# SESSION 28*

**WED. 10:00 A.M.–12:30 P.M.**

**WALDORF-ASTORIA JADE ROOM**

## Nuclear Instrumentation

*Chairman:* M. A. Schultz, *Westinghouse Commercial Atomic Power, Pittsburgh, Pa.*

### 28.1. Track Recognition System for Scanning Nuclear Emulsions

S. Becker and J. Franceschini, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

An experimental scanner using television techniques has been built to detect and count neutron tracks in the photographic emulsion used in monitoring personnel radiation exposure. The principal problem is the reading of very shor' tracks in a background of artifacts and random grains.

Artifacts or grain clusters with low aspect ratio are rejected by a form of cancellation similar to that used to eliminate radar ground clutter. The duration and continuity of the canceller output signal is then screened to pass signals characteristic of straight tracks with close grain spacing.

Statistics for the automatic scanner are given and compared with those for existing manual techniques. The maximum practical scanning rate is estimated.

### 28.2. Multichannel Analyzer for Time-of-Flight Experiments

H. L. Garner and R. E. Miller, *University of Michigan, Ann Arbor, Mich.* and S. H. McMillan and R. R. Graham, *Strand Engineering Co., Ann Arbor, Mich.*

The work described in this paper was carried out as one portion of a program by the Knolls Atomic Power Laboratory, Yale University, and Columbia University for the Atomic Energy Commission. This program required the development of instrumentation for the collection of data on the velocity distribution of neutrons. This instrumentation was to provide a considerably higher resolution and capacity than previously obtainable.

A cathode-ray tube monitor permits visual examination of the accumulated count of selected groups of channels while the experiment is in progress.

The output unit operates a standard commercial card punch which permits analysis by general purpose computers.

### 28.3. 0.1-Microsecond, 2000-Channel, Electrostatic Storage System for Time-of-Flight Experiments

J. Hahn, *Dept. of Physics, Columbia University, New York, N. Y.*

An instrument has been developed for measuring neutron time-of-flight which has 2000 discrete time channels, each having a width of 0.1 microsecond. A temporary electrostatic storage is used in conjunction with a magnetic drum memory. Neutrons initiated by a cyclotron burst, which occurs at approximately a 60-cps rate, are detected after traversing a fixed flight path, and their times of arrival are stored in the electrostatic memory. Before the next burst occurs the stored information is read into the drum, and the storage tube is cleared. This paper describes the electrostatic storage system and circuitry. The major system blocks which are discussed are: the 10-mc pulsed oscillator and clock pulse generator; the "clock-stopper" or staticizer; the fast carry flip-flop deflection circuit drivers; the current adder stairstep deflection generators, and the storage tube.

### 28.4. New Double-Line Linear Amplifier

G. G. Kelley, *Oak Ridge National Lab., Oak Ridge, Tenn.*

A new amplifier has been developed for scintillation spectrometry; it has overload characteristics suitable for use with very large detectors. This amplifier, called the A-8, recovers from a 4000-times overload signal in about 8 $\mu$sec. There is no positive base line excursion after the main pulse. It uses the double delay line differentiation principle, has a gain of 13,500 in the main amplifier, and an extra factor of 1, 3, or 10 in the preamplifier. The gain control range is by factors of 2 over a total range of 64. Rise time is about 0.18 $\mu$sec. The output stage is capable of driving a 1000-ohm load to greater than 100 volts in both directions. The power requirement is about the same as for an A-1 amplifier.

### 28.5. Radioisotope Thermoelectric Generator

J. L. Briggs, *Rome Air Development Center, Griffiss Air Force Base, N. Y.*

Recent advances in the field of solid-state physics have greatly enhanced the possibility of developing a practical thermoelectric generator for power supply electronic equipments. The possibility of utilizing an efficient thermopile driven by decay energy derived from isotopes is suggested. The course of development of this device at RADC will encompass three problem areas: 1) geometry and packaging of radioisotope thermal source; 2) development of efficient thermopile, and 3) consideration of circuits best suited to this type of power supply.

* Sponsored by the Professional Group on Nuclear Science. To be published in Part 9 of the 1957 IRE Convention Record.

# SESSION 29*

**WED. 10:00 A.M.–12:30 P.M.**

**WALDORF-ASTORIA SERT ROOM**

## Circuit Theory I—Symposium on Modern Methods in Network Theory

*Chairman:* S. Darlington, *Bell Telephone Labs., Murray Hill, N. J.*

The first half of the Symposium consists of three invited papers, each concerned with one of three branches of network theory in which much work is currently being done. These papers by recognized experts in the field will attempt to survey the present state of the art in each area and lead into some of the problems under investigation at present.

The second half will be a panel discussion in which the three previous speakers will be joined by B. J. Dasher (Georgia Inst. Tech.), M. S. Corrington (RCA, Camden), E. S. Kuh (Univ. of California), A. D. Fialkow (Brooklyn Polytech. Inst.), and H. J. Carlin (Brooklyn Polytech. Inst.). The panel will discuss some of the current problems in network theory, the progress being made in them, and the need for future studies.

### 29.1. Synthesis Techniques and Active Networks

J. G. Linvill, *Stanford University, Stanford, Calif.*

Synthesis theory for passive networks, as an aid in conception, neatly separates the possible from the impossible; as a practical technique of network design the theory is useful since paper designs lead to good laboratory designs. With respect to active networks the same two parallel aspects, theoretical implications and practical techniques, emerge but assume different importance.

The realm of theoretical possibilities is expanded when an active element in the form of an ideal amplifier, a negative resistance, or a negative impedance converter is introduced. With the introduction of active elements one can eliminate one kind of reactance without loss of generality, can compensate for incidental lossiness in reactors, and can incorporate power amplification into the network.

Practical considerations introduce a new dimension into active network design. This new dimension arises through the fact that characteristics of active elements are more inclined to drift with time and environment than passive elements. The designer must evaluate sensitivity of performance to changes in active elements and design structures which are maximally tolerant of variations in elements. Some preliminary guides can be given but many interesting problems are yet to be solved.

### 29.2. The Frequency-Time Representation of Signals Using Natural Components

W. H. Huggins, *The Johns Hopkins University, Baltimore, Md.*

A wide class of signal representations are characterized by a resolution into components $k(t; \lambda)$ involving a single parameter $\lambda$, according to the form

$$u(t) = \sum_{\lambda} k(t; \lambda) U(\lambda)$$

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where $U(\lambda)$ is the signal spectrum as a function of $\lambda$. For representation in the "frequency domain," $k(t; \lambda) = \exp(\lambda t)$ for $t > 0$, all components have the *same epoch* at $t = 0$, and each component has a *different waveform* that depends upon the "frequency" parameter $\lambda$. On the other hand, for "time-domain" representation used in sampled-data systems, $k(t; \lambda) = \Delta(t - \lambda)$, where $\Delta$ is some interpolating function such as the cardinal function, all components have the *same waveform* $\Delta(\tau)$ and each component has a *different epoch* that depends upon the sampling instant $\lambda$.

Since a "frequency" representation always implies at least one epoch, and a "time" representation always implies at least one component waveform, "frequency" and "time" are inseparable attributes of a signal representation. A more general, and often simpler and more meaningful, representation involving multiple frequencies and multiple epochs may be written as

$$u(t) = \sum_{j,k} A_{j,k} f_j(t - t_k)$$

where $f_j(\tau) = \exp(\lambda_j \tau)$ are damped natural exponential components that vanish for $\tau < 0$. Because these components decay to zero with the passage of time they permit a concise and physically meaningful definition of a time-varying spectrum. The problem of approximating a given signal with an expression of this sort is discussed and the design of a signal analyzer capable of measuring the parameters $\{A_{j,k}\}$ is described.

### 29.3. Signal Flow Graph and How to Avoid Them

S. J. Mason, *Massachusetts Institute of Technology, Cambridge, Mass.*

Signal flow graph topology and network topology are both useful in linear electronic circuit analysis. Flow graphs lend themselves most easily to circuits principally composed of unilateral elements whereas classical network-topological methods are applicable to circuits containing only reciprocal elements. By a very simple modification the classical topological rules for reciprocal branch network analysis can be adapted to circuits containing nonreciprocal elements, thereby permitting electronic circuit models to be analyzed by inspection and without flow graphs.

## SESSION 30*

**WED. 10:00 A.M.–12:00 NOON**

**WALDORF-ASTORIA GRAND BALLROOM**

### Engineering Management Technique

*Chairman:* H. N. Skifter, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

### 30.1. The Art of Selection of Engineering Management Talent

C. W. Randle, *Booz, Allen, and Hamilton, Chicago, Ill.*

The growth and increasing importance of engineering, in combination with the shortage of engineers and the general shortage of management talent, have increased the desirability of defining engineering management selection standards and measuring engineers against these standards. The characteristics of the engineer vs the characteristics of the manager will be discussed and their differences, as well as their common characteristics, will be brought out.

### 30.2. The Art of Human Relations: Manipulation or Motivation?

A. Levenstein, *Research Institute of America, New York, N. Y.*

The shortage of personnel calls for action, even where there is no prospect of increasing the number of people available. The answer may very well be *not* more people, but better people at work. Under present circumstances, human relations in industry become critical, more important than ever. Unfortunately, this relatively new approach to urgent problems is being seriously weakened by those who believe that manipulative devices can be effective. They can't. Better human performance depends on 1) developing new management attitudes, 2) working out means of communicating those attitudes, and 3) initiating programs of motivation that genuinely identify the areas of mutual interest for all involved in the production process.

### 30.3. The Art of Delegation of Authority

W. J. E. Crissy, *Personnel Development, Inc., New York, N. Y.*

It is axiomatic in management theory and practice that: 1) responsibility inheres in the individual; 2) responsibility cannot be shared; 3) responsibility cannot be delegated, and 4) authority must be commensurate with responsibility if objectives are to be reached. In your business or mine, this means that the head of the company is fundamentally responsible, for better or worse, in all matters, yet obviously the head of a business cannot do everything himself. If he could there would be no need for other people.

An examination of these axioms from the psychological standpoint leads us to many ramifications which *in toto* reflect themselves in the "climate" that exists within the business and the reputation it has in the eyes of its various publics. Specifically, let us examine the dynamics of delegation from the standpoint first of the delegator, and second, the delegate. This separation you will recognize as arbitrary since every executive at whatever level, except the top, holds these two roles simultaneously.

The job assignment of every executive calls for planning, seeing, doing, and measuring activities. At the top echelons of management, the demands are mostly for planning and measuring. In the middle echelons, the demands are mostly for seeing and measuring and at the first level of supervision for seeing, doing, and measuring. Delegation is tied most intimately to seeing or supervising activities though, of course, antecedent planning activity must take place.

Now let us examine why proper delegation is difficult to achieve. First, not all executives have the temperamental make-up needed for delegation. For example, executive insecurity begets a desire to know exactly what every delegate is doing. Then too, executives sometimes view themselves as indispensable—"the whole show." This prevents them from allowing those working for them to be recognized. Second, the intellectual make-up of the individual may make delegation difficult. He may not honestly see the need for extending authority down the line. He may see himself as the "expert knower." Third, lack of requisite communications skill may prevent the executive from delegating. He knows how to do the job himself but cannot communicate what to do and how to do it to others.

Similarly, viewing this from the delegate's standpoint, some people are loath to assume responsibility and find it more secure to clear decisions up the line. Also delegates may not fully understand what is expected and yet are loath to reveal their ignorance. Finally, it is often difficult to straighten out communications on complex assignments and the delegator has one view but the delegate has a discrepant one.

The management ideal should be that everyone at every level have commensurate authority to carry out his responsibilities.

## SESSION 31*

**WED. 10:00 A.M.–12:30 P.M.**

**NEW YORK COLISEUM MORSE HALL**

### Transistor Applications

*Chairman:* L. R. Fink, *General Electric Co., Schenectady, N. Y.*

### 31.1 Circuit Considerations for High-Frequency Amplifiers Using Drift Transistors

J. W. Englund and A. L. Kestenbaum, *Radio Corp. of America, Somerville, N. J.*

This paper describes design considerations for high-frequency circuits using new drift transistors in which the useful frequency response has been increased by an order of magnitude over conventional junction transistors without changing the configuration of the alloy structure. Circuits are presented showing the use of these transistors in a two-stage unneutralized 455-kc amplifier, a neutralized single-stage IF amplifier, a two-stage IF amplifier operating at 10.7 mc for fm receivers, and a three-band portable radio receiver covering the broadcast band (540 to 1630 kc) and short-wave bands (4.75 to 11 mc and 10 to 23 mc). In the portable receiver, the drift transistors are used as an rf amplifier a mixer, and a separate oscillator. The electrical characteristics of the drift transistors are defined and performance curves illustrating their performance at high frequencies are presented.

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* Sponsored by the Professional Group on Broadcast and Television Receivers. To be published in Part 3 of the 1957 IRE Convention Record.

### 31.2. Design Considerations in the First Stage of Transistor Receivers

L. A. FREEDMAN, *RCA Labs., Princeton, N. J.*

This paper includes a discussion of noise performance of transistor rf stages utilizing capacitive antennas and of transistor mixer stages utilizing loop antennas. Examples of the noise performance to be expected with each type of antenna are included. Comparisons are drawn between transistor stages and corresponding tube stages. Consideration is given to design compromises between image rejection and insertion loss for an rf stage employing both tuned input and interstage transformers. The procedure for transformer design for optimum insertion loss-image rejection performance is outlined.

### 31.3. A Six-Transistor Portable Receiver Employing a Complementary Symmetry Audio Output Stage

D. D. HOLMES, *RCA Labs., Princeton, N. J.*

A complementary symmetry audio output stage offers advantages of simplicity and of economy of components over a conventional transformer-coupled arrangement. When applied to a broadcast receiver, the elimination of audio transformers can be instrumental in providing a striking subjective improvement in receiver audio quality. This paper discusses the unique problems which arise with respect to the integration of a complementary symmetry circuit in a complete receiver design. A six-transistor receiver is described featuring novel audio, volume control, and agc circuitry which makes practicable the exploitation of the inherent advantages of complementary symmetry.

### 31.4. Transistor Receiver Circuits

A. PROUDFIT, K. M. ST. JOHN, C. R. WILHELMSEN AND R. J. FARBER, *Hazeltine Corp., Great Neck, N. Y.*

A considerable portion of the cost of a transistor broadcast receiver is represented by the transistors themselves. This paper describes the use of multiple unit grown junction germanium transistors in a manner that effects a significant cost saving in the complete receiver.

The bulk of the cost reduction is due to a lower price for the transistors used. Departures from conventional circuits are necessary to take advantage of the new transistors. A number of these new circuits are described along with a typical receiver. Performance data for the receiver indicate that there has been no compromise in performance with these new techniques.

### 31.5. Transistor Circuit Problems in TV Receiver Design

E. M. CREAMER, JR., L. H. DEZUBE, J. P. MCCALLISTER, *Philco Corp., Philadelphia, Pa.*

The paper discusses the technical problems that arise in designing transistor circuits to perform several television receiver functions. The economic factors are treated in terms of the minimum number of transistors and components required per set function. The technique employed is to consider the basic requirements of the sound and picture transducers and to specify circuits and transistors which will meet these needs. Vacuum tube experience is adapted where useful, although the special properties of the transistor often permit advantageous designs not previously possible. Examples are given of circuits which so far have been investigated. These circuits are generally suitable for use in combination with other circuits employing vacuum tubes.

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## SESSION 32*

**WED. 10:00 A.M.–12:30 P.M.**

**NEW YORK COLISEUM MARCONI HALL**

**Microwave Antennas**

*Chairman:* D. D. KING, *Electrical Communications, Inc., Baltimore, Md.*

### 32.1. A Versatile Multiport Biconical Antenna

R. C. HONEY AND E. M. T. JONES, *Stanford Research Institute, Menlo Park, Calif.*

This paper describes a versatile, multiport, biconical antenna that can be used as a wide-band direction finder or multiplexer. When the antenna is used for reception, incident plane waves excite in the feeding coaxial line both the TEM mode and the orthogonal $TE_{11}$ mode whose azimuthal orientation depends on the direction of arrival of the signal. Directional information can be obtained by measuring the sum of these two modes at four fixed detectors arranged at 90° intervals around the coaxial line. Alternatively the linearly polarized $TE_{11}$ mode can be resolved, with appropriate circuitry, into right- and left-hand circularly polarized $TE_{11}$ modes. Then directional information can be obtained by measuring the phase difference between either of the circularly polarized modes and the TEM mode. This latter form of direction finder is well suited to multiplexing applications because each of its three ports is well isolated from the others. Furthermore, this antenna has the unique property that a signal fed into any one of these three ports will cause the antenna to radiate omnidirectionally in azimuth. A waveguide version of this antenna has been built and tested and found to operate well over the frequency range of 8.2 to 12.4 kmc.

### 32.2. Recent Annular Slot Array Experiments

K. C. KELLY, *Hughes Aircraft Co., Culver City, Calif.*

Experimental results on several X-band designs for annular slot planar arrays are presented. Three types were considered. The first generates a pencil beam perpendicular to the plane of the array and was undertaken to explore the possibility of producing beams of this type without resort to structures consisting of a multiplicity of rectangular waveguides. The second type produced beacon patterns of various shapes in elevation. These beacon antennas were used independently or to excite a slow wave surface antenna. The third type of design produced a pencil beam in or near the plane of the array. Scanning in azimuth was possible by a small motion of the feed point in the radial transmission line which excites the slots.

Slot coupling data pertinent to obtaining specified aperture distributions are presented.

### 32.3. Radiation from Modulated Surface-Wave Structures—I

F. J. ZUCKER, *Air Force Cambridge Research Center, Bedford, Mass.*, AND A. S. THOMAS, *A. S. Thomas, Inc., Boston, Mass.*

A "modulated" surface wave is one whose phase and/or amplitude varies periodically as the wave propagates along the surface. Structures that support such a wave are characterized by an effective index of refraction that varies in the direction of propagation; for example, Simon's "Cigar" antenna (which inspired this work), or a Yagi with variable element spacing or a variable thickness dielectric slab.

Pattern calculations based on harmonic analysis of the spatial frequencies contained in phase-and/or-amplitude-modulated waves show that surface wave antennas can be built 1) with very high end fire gain or 2) with the beam tilted at an arbitrary angle with the surface (including broadside). Preliminary experimental evidence will be presented.

### 32.4. Radiation from Modulated Surface-Wave Structures—II

R. L. PEASE, *Hughes Aircraft Co., Culver City, Calif.*

Approximate expressions are derived for the electromagnetic fields in a surface wave which propagates along a typical modulated structure—a rectangular slab of uniform thickness, mounted on a perfectly conducting surface, whose dielectric constant and/or permeability vary slowly in a sinusoidal manner along the direction of propagation. Radiation patterns are computed for such a structure of finite length set in an infinite ground plane. In contrast to the endfire pattern obtained from a homogeneous slab, it is possible to obtain a tilted beam from the modulated structure by proper control of the amplitude, wavelength, and phase of the structure modulation.

### 32.5. The "Sandwich Wire" Antenna: A New Microwave Line Source Radiator

W. ROTMAN AND N. V. KARAS, *Air Force Cambridge Research Center, Bedford, Mass.*

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* Sponsored by the Professional Groups on Microwave Theory and Techniques and Antennas and Propagation. To be published in Part 1 of the 1957 IRE CONVENTION RECORD.

A new microwave line source configuration is described. Basically, it consists of three continuous co-planar conductors; the outer conductors which are parallel to one another are at ground potential while the center conductor which is excited is bent in a sinusoidal manner with a period (for broadside radiation) equal to the wavelength. The polarization is transverse to the length of the array.

Modifications of the basic design include a printed circuit (metal on dielectric) two-dimensional array which is fed by a corporate structure feed, a helical structure which has an omnidirectional pattern and a circularly polarized linear array. Advantages of this construction are mechanical simplicity, good impedance characteristics, and good control over the aperture illumination.

### 32.6. Recent Developments in the Study of Printed Antennas

J. A. McDonough and R. G. Malech, *Airborne Instruments Laboratory, Inc., Mineola, N. Y.*

Some interesting developments have occurred recently in the field of printed antennas. It is now possible to accurately print end-fire arrays, whereas reference 1 discusses only broadside arrays. A procedure has been developed which permits accurate scaling of lower frequency Yagi arrays to high frequency (microwave) printed Yagis. The exact expressions for the gain function of a rhombic over a semi-infinite perfectly conducting ground plane have been developed. Using these equations, a rhombic was designed and printed. The performance of this rhombic closely correlated prescribed specifications. Two such rhombics have been conductively arrayed with successful resu'ts. Some investigation has been done with printing parasitic rhombics to attain illumination control. Slightly narrower beamwidth and lower side lobes have been recorded. An elliptical configuration has been printed and tested. The performance of the ellipse compares favorably with the performance of the rhombic.

Printed "ladder type" end-fire arrays have been fabricated at *L*, *S*, *C*, and *X* band. These antennas are of the surface wave type and have been made to operate successfully up to 16 wavelengths long at *S* band. Illumination is controlled by modulating the element lengths or the distance between elements.

Broadside arrays have been designed and fabricated which employ capacitive coupling techniques. Short, wide dipoles are closely spaced to provide a phase reversal at the capacitive gap. Illumination probe measurements have been made and the data obtained have been correlated with the radiation pattern.

## SESSION 33*

**WED. 10:00 A.M.–12:30 P.M.**

**NEW YORK COLISEUM FARADAY HALL**

### Electron Tubes—General

*Chairman:* G. M. Rose, Jr., *Radio Corp. of America, Harrison, N. J.*

* Sponsored by the Professional Group on Electron Devices. To be published in Part 3 of the 1957 IRE Convention Record.

### 33.1. Practical Design Theory for Minimization of Vibration Noise in Grid-Controlled Vacuum Tubes

G. Gross, *Raytheon Manufacturing Co., Newton, Mass.*

Theory shows vibration noise minimization is achievable through electrical and/or mechanical design. It is shown, however, that determination of requisite conditions by operations on the plate current equation is invalidated in the island effect region, commonly the region of receiving tube design. This results from a substitution into the plate current equation of geometric functions defining the electrical parameter $u$ which are themselves inapplicable in this region.

The practical design theory presented here is based upon empirical determination of the $u$ function and consequent application of its variance with vibration to either: 1) reduce its vibration effect to zero. or 2) use it to cancel other vibration effects.

Successful application of the theory is illustrated in the design of the CK 6533, a subminiature triode with extremely low vibration output. Similar application to pentodes is indicated.

### 33.2. Electrolytic Tank Measurements of Mesh Grid Characteristics

H. Hsu, *General Electric Co., Syracuse, N. Y.*, and C. E. Horton, *General Electric Co., Owensboro, Ky.*

Mesh grids are finding increased use in such devices as storage tubes, camera tubes, and receiving tubes; but unfortunately most of the available design equations apply only to parallel wire grids with circular wires. A convenient electrolytic tank is described in which the electrostatic fields associated with mesh grids were duplicated. This arrangement permits many of the dimensions of the configurations to be continuously variable. Grid wires of both circular and rectangular cross section were studied in parallel wire grids and in mesh grids with both square and rectangular openings. The method also permits measurements of triangular and hexagonal configurations.

### 33.3. Rare-Earth Oxide Cathodes

L. J. Cronin and J. H. Apelbaum, *Raytheon Manufacturing Co., Waltham, Mass.*

In the field of physical electronics, there is need for electron emitting materials which will compete or improve on existing designs utilizing thorium dioxide. With this in mind, investigations were conducted on the following rare-earth materials—gadolinium oxide, neodymium oxide, samarium oxide lanthanum oxide, didymium oxide, and combinations of these and other oxides. Measurements on these materials for thermionic emission data were taken in simple diodes with the emitting material held in a molybdenum mesh. Processing of the materials and of the vacuum diodes are described. Values are given of the dc and pulse emission data. These data are reported as a function of life and in comparison with a standard 100 per cent thoria cathode. The results indicate this class of materials offers definite possibilities as useful cathode materials.

### 33.4. Temperature Distribution in Anode Structure for Pulse Input

R. N. Ghose, *The Ramo-Wooldridge Corp., Los Angeles, Calif.*

Heat flow equation for the anode structure of a water cooled high power transmitting tube has been solved in cylindrical coordinates subject to boundary conditions which require a constant temperature at the inner surface of the anode and a pulse type of anode current. The temperature distribution in the entire region of the anode structure is expressed, in this paper, both as a function of time and the geometry of the anode. Comparison of temperature distribution at the anode structure is also made for the pulse and sinousoidal types of anode currents. The analysis presented in the paper can be generalized for any anode current distribution.

### 33.5. A Positive Grid Voltage-Space Current Division Test for Power Vacuum Tubes

J. A. Jolly, *Eitel McCullough Inc., San Bruno, Calif.*

A new positive grid voltage-space current division test for power vacuum tubes is presented. The existing test techniques for power vacuum tubes and their limitations in the positive grid voltage-high space current region are indicated. The new test is covered by discussion and data presentation that indicate the ability to measure tube-to-tube electrical variations in the positive grid voltage region. Related mechanical tube variations and dynamic circuit effects for one tube type are illustrated.

A particular application of the positive grid voltage test as related to the evaluation of oxide cathodes is included.

### 33.6. Electron Tubes for Critical Environments

W. H. Kohl, *Stanford Research Institute, Menlo Park, Calif., and Stanford University, Stanford, Calif.*

The rapid advances in the technology of weapons systems, and also in specialized applications for civilian use, are putting ever increasing demands on the reliability of electron tubes operating in critical environments. During the past year, Stanford Research Institute has been under contract with Wright Air Development Center to survey the present state of the art of tube-making for applications where envelope temperatures up to 500°C, high vibration and shock levels, and high neutron radiation flux densities are encountered. This paper will summarize our findings on the basis of a survey of the literature and of numerous interviews with people in research and industry.

# SESSION 34*

**WED. 2:30–5:00 P.M.**

**WALDORF-ASTORIA STARLIGHT ROOF**

## Symposium—Long Range Telemetry and Remote Control

*Chairman:* M. V. KIEBERT, JR., *The Miami Shipbuilding Corp., Miami, Fla.*

### 34.1. Electronic Control and Instrumentation of Extra-Atmosphere Space Craft

M. V. KIEBERT, JR., *The Miami Shipbuilding Corp., Miami, Fla.*

The advent of space flight, initially in unmanned craft and ultimately in manned craft, poses new problems in telemetry and radio control techniques which must be considered at this time. A "holding pattern" of perhaps 60,000 miles or more, a fuel and oxidizer consumption of tons per minute, combined with a reentry problem necessitating expenditure or dissipation of large amounts of energy, will require early adoption of a "common" system of macroscopic capabilities that will dwarf the air navigation and control problems that now confront the aviation industry in the transition from piston to jet aircraft. The presently visualized problems, including propagation phenomena, environmental considerations, flight path, and/or trajectory control will be outlined and compared with the author's previously published paper "Electronic Control and Instrumentation of Aircraft," which was presented at the joint R.Ae.S.I.A.S. meeting in London in 1947.

### 34.2. Progress in Telemetering of Data from High Velocity Missiles in Upper Atmosphere

R. J. BURKE, *Lockheed Corp., Palo Alto, Calif.*

In order to collect data concerning the reentry of bodies into the earth's atmosphere, Lockhead Missile Systems Division has been carrying on an extensive program for the Air Force using a multistage rocket. This vehicle is designed to leave the atmosphere powered by its first stage and then to turn over and reenter at very high speeds attained by the thrust of its second and third stages. A large amount of information has already been obtained by telemetering from these vehicles and although much of it is still classified, some indication can be given of the problems encountered, and in particular a description will be given of a peculiar phenomenon observed in the variation of radio frequency field strengths under these flight conditions.

### 34.3. Long Range Telemetering Reception

J. B. WYNN, *RCA Missile Test Center, Melbourne, Fla.*

Reception problems concerned with long range telemetering will be discussed. Consideration will be given to signals arriving at low elevation angles, originating from vehicles at considerable slant ranges, but reasonable altitudes.

This paper will be based on unclassified experience obtained in the operating area included in the Air Force Missile Test Center.

### 34.4. Some Dynamic Aspects of Control at Long Ranges

D. T. SIGLEY, *Firestone Tire and Rubber Co., Los Angeles, Calif.*

The problem of instrumenting space craft from the point of view of the control problem will be discussed. Questions of shading, angular velocities, accelerations, and accuracies of measurement and control will be discussed.

### 34.5. Problem in Measuring the Space Environment of the Earth

A. J. MAX, *Sandia Corp., Albuquerque, N. M.*

This paper will discuss instrumentation and telemetering for space craft to measure physical data, such as magnetic radiation, solar radiation, gamma radiation, sunspot activity, etc. It will include a discussion on environmental problems, such as shock and acceleration encountered by this instrumentation.

### 34.6. A Microwave Telemetry Relay

K. A. HALL, B. J. LAMBERTY, AND L. S. TAYLOR, *Flight Determination Lab., White Sands Proving Ground, N. M.*

A 2200-mcps multichannel telemetry relay system has been developed at White Sands Proving Ground. This equipment is designed to provide a high capacity net interlinking all data-recording sites and the fixed and mobile receiving stations on the proving ground. In addition, simultaneous calibration of all stations is possible. The relay employs the superheterodyne principle to translate all the vhf telemetry signals as a unit to the microwave region for transmission to the recording stations. The system features a large dynamic range. Close channel spacing is utilized for efficient use of bandwidth. Unattended operation of the equipment is possible.

* Sponsored by the Professional Group on Telemetry and Remote Control. To be published in Part 5 of the 1957 IRE CONVENTION RECORD.

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# SESSION 35*

**WED. 2:30–5:00 P.M.**

**WALDORF-ASTORIA ASTOR GALLERY**

## Speech Analysis and Audio Amplifiers

*Chairman:* H. F. OLSON, *Radio Corp. of America Labs., Princeton, N. J.*

### 35.1. A Demonstration of the Representation of Speech by Poles and Zeros

S. H. CHANG AND R. BACH, JR., *Northeastern University, Boston, Mass.*

Aside from the phonemic description of speech the most effective representation of speech is supplied by the dynamic model. The latter concept of excitation functions and system functions is very appealing to communication scientists. Work remains to be done to demonstrate how successful this concept is in solving speech compression problems. This paper demonstrates our progress to date.

### 35.2. A High Efficiency Speech Amplifier

H. SULLIVAN, *David Bogen Co., Inc., New York, N. Y.*, AND J. NELSON, *Flushing, N. Y.*

A 500-watt audio amplifier for speech use operating at an over-all efficiency of 80 per cent is described. The plate efficiency of the output stage is approximately 98 per cent. These high efficiencies are achieved by using infinitely clipped speech so that the output consists of rectangular pulses and the output tubes operate as switches. The method of rating tubes for this service, the efficiencies to be expected, and the precautions to be taken to assure high articulation, are discussed. Word articulation scores using the PB50 word list exceed 90 per cent.

### 35.3. 50 Watt High Quality Transistor Audio Power Amplifier

A. B. BERESKIN, *University of Cincinnati, and The Baldwin Piano Co., Cincinnati, Ohio*

A transistor audio power amplifier capable of delivering power in excess of 50 watts has been developed. The power delivering capacity of this amplifier is down less than 3 db at 20 and 15,000 cycles while the low level frequency response is down 3 db at about 60 kc. The desirable operating characteristics have been achieved by strict Class B operation of the push-pull output stage in a three stage circuit capable of accepting a very large amount of over-all feedback. Provisions are available for matching any load impedance between the limits of 5 and 24 ohms.

### 35.4. Low Noise Transistor Microphone Amplifier

J. J. DAVIDSON, *Radio Corp. of America, Camden, N. J.*

The attraction of transistors for use in low-level amplifiers has long been obvious. Such factors as size, weight, power consumption, and economy have indicated the potential superiority of transistors over vacuum tubes.

* Sponsored by the Professional Group on Audio. To be published in Part 7 of the 1957 IRE CONVENTION RECORD.

Until recently, however the prime requirement of low noise factor could not be met with any consistency. Although transistors under optimum conditions appear inherently inferior to vacuum tubes under optimum conditions the difference is slight, and other factors, such as reliability and stability may swing the balance. Some trends and criteria for low noise design are discussed, including results on some experimental transistors. An experimental microphone amplifier is utilized as the embodiment of the design requirements.

### 35.5. Circuit Considerations for Audio Output Stages Using Power Transistors

R. MINTON, *Radio Corp. of America, Somerville, N. J.*

This paper discusses design considerations for audio-frequency output stages using alloy-junction *p-n-p* power transistors. Optimum transistor operating conditions for practical circuit designs are given.

Both class A single-ended and class B push-pull power output stages are evaluated, and the effects of variations in source and load impedance are discussed. Methods of obtaining the desired stability of dc bias with variation in ambient temperature are described. Experimental data are presented in the form of curves of maximum power output power gain distortion and efficiency of a typical application using the new power transistor.

## SESSION 36*

**WED. 2:30–5:00 P.M.**

**WALDORF-ASTORIA JADE ROOM**

### Transistorizing Nuclear Instrumentation

*Chairman:* R. F. SHEA, *General Electric Co., Schenectady, N. Y.*

### 36.1. Transistorizing Nuclear Instruments—A Status Report

R. F. SHEA, *General Electric Co., Schenectady, N. Y.*

The present status of the use of transistors in nuclear instrumentation will be reviewed. Possible areas of application which are still untapped will be listed, and a discussion of the research necessary to foster transistorization to nucleonic instrumentation will be given.

### 36.2. Noise in Transistor Nucleonic Pulse Amplifiers

A. R. JONES, *Atomic Energy of Canada, Ltd., Chalk River, Ontario*

An experimental study of the high-frequency noise in surface barrier transistors has been made. It was found that above about 10-kc shot noise predominated. This noise followed the pattern predicted by Van der Zeil in its dependence upon standing currents and frequency. The application of these results to the design of a nucleonic pulse amplifier is discussed. In particular, a design for a preamplifier employing transistors is examined to see how the noise may be minimized without sacrifice of the required bandwidth. Also the choice of time constants employed in the shaping of the pulses in order to obtain the best energy resolution is considered.

### 36.3. Recent Advances in Transistorizing Reactor Controls

K. H. KLEIN, *Daystrom, Inc., Archbald, Pa.*

This paper will describe a transistorized reactor control system including in all-transistor scaler and some other transistorized developments that are to be carried out in an effort to reduce size and weight in reactor controls. Other instruments that don't lend themselves to full transistorization will also be described.

### 36.4. Transistorized Time-of-Flight Analyzer with Ferrite Core Memory

E. J. WADE AND D. S. DAVIDSON, *General Electric Co., Schenectady, N. Y.*

The neutron time of flight analyzer is one of the basic tools used by the nuclear physicist and various designs have been built which take advantage of continuing improvements in the electronic and computer fields.

The development of the random access ferrite core memory made it practical to greatly increase the number of storage channels while at the same time reducing the number of vacuum tubes.

Improvements in transistors have made it feasible to use them with a great saving in space and power requirements especially in computer-type circuits where the requirements are primarily switching which minimizes the necessity for close control of transistor characteristics and reduces the dissipation requirements.

Accordingly, a completely transistorized 256-channel analyzer has been designed and built which is housed with power supplies in a cabinet only 19 inches wide, $8\frac{3}{4}$ inches high, and 13 inches deep.

### 36.5. Transistorized Counting-Rate Meter and High-Voltage Supply

A. PEARLMAN, *Universal Atomics Corp., Westbury, N. Y.*

In order to provide a scintillation-counter-type survey meter suitable for medical and airborne ore-locating use, a light-weight transistorized system has been developed. The unit described weighs approximately 22 lbs. complete with scintillation crystal, photomultiplier tubes, and lead shield.

The counting-rate meter amplifier, discriminator, and high-voltage supply are transistorized and operate off self-contained batteries. However, an auxiliary ac power supply is incorporated into the instrument for operation from standard 115-volt 60-cycle power.

## SESSION 37*

**WED. 2:30–5:00 P.M.**

**WALDORF-ASTORIA SERT ROOM**

### Symposium on Applications of Computers in Biology and Medicine

*Chairman:* P. H. MITCHELL, *Air Research and Development Command, Baltimore, Md.*

*Panel Members:* R. PEPINSKY, *Dept. of Physics, Pennsylvania State University, University Park, Pa.*
M. EDIN, *National Heart Institute, National Institutes of Health, Bethesda, Md.*
B. CHANCE, *Johnson Research Foundation, University of Pennsylvania, Philadelphia, Pa.*
J. W. MAUCHLY, *Remington Rand Univac, Philadelphia, Pa.*
R. S. LEDLEY, *Operations Research Group, George Washington University, Washington, D. C.*

Just as other standard electronic techniques have been applied to the solution of complex biological and medical problems computer techniques are developing rapidly in these fields. Some of the present and future applications to be discussed include:

1) The visualization of complex organic molecular structure.
2) Analog and digital computer representations of the dynamics of metabolism.
3) Medical data processing for correlative, diagnostic, and searching purposes.
4) The probable and the possible in computer applications.
5) Monte Carlo methods in cell multiplication processes.

## SESSION 38**

**WED. 2:30–5:00 P.M.**

**NEW YORK COLISEUM MORSE HALL**

### Color Television

* Sponsored by the Professional Group on Nuclear Science. To be published in Part 9 of the 1957 IRE CONVENTION RECORD.

* Sponsored by the Professional Groups on Medical Electronics and Electronic Computers. To be published in Part 4 of the 1957 IRE CONVENTION RECORD.

** Sponsored by the Professional Group on Broadcast and Television Receivers To be published in Part 3 of the 1957 IRE CONVENTION RECORD.

*Chairman:* W. O. Swinyard, *Hazeltine Research, Inc., Chicago, Ill.*

### 38.1. Developments in Color TV in Europe

C. J. Hirsch, *Hazeltine Corp., Little Neck, N. Y.*

An up-to-date review of proposals for color television system standards in Europe will be presented by the speaker, Chairman of the CCIR Television Preparatory Committee, U. S. Department of State, who has an intimate first-hand knowledge of this technical situation.

### 38.2. Brightness Enhancement Techniques for the Single-Gun Chromatron

R. Dressler, P. Neuwirth and J. Rosenberg, *Chromatic Television Laboratories, Inc., New York, N. Y.*

Single-gun color cathode-ray tubes must display color video information in a sequential manner. United States color signal standards utilize a subcarrier of 3.58 megacycles. At this frequency color gating for the single-gun display is accomplished with sine waveforms which contribute to a loss in light output.

This paper describes two techniques, "single-key-out double blue" and "chroma modulated 10.7 megacycles," which allow recovery of the light loss incurred by these gating waveforms. In order to describe the enhancement processes, basic light outputs from Lawrence tubes without gating will be described along with the simple tripler decoding technique which contributes a 66 per cent light loss. These enhancement techniques allow 50–60 foot lamberts light output from presently available tubes.

In addition, circuits and system errors will be discussed

### 38.3. Accuracy of Color Reproduction in the "Apple" System

J. B. Chatten and R. K. Gardner, *Philco Corp., Philadelphia, Pa.*

In the "Apple" system the mechanism of color selection is quite different from that of a color receiver system using a three-gun display tube. In the Apple system this mechanism is a sequential sampling process wherein the crt drive waveform, the transfer characteristic of the electron gun, the shape of the scanning aperture, and the stripe structure are all factors in defining the reproduced color.

An intensive study has been made of the relationship between the crt drive signal and the reproduced color wherein all the above mentioned factors have been considered. From this it is possible to make a comparison of color reproduction (both chromaticity and luminance) between an Apple receiver and an assumed "standard monitor," as implied by the FCC color television broadcast standards.

### 38.4. Recent Improvements in the Apple Beam Indexing Color Picture Tube

H. R. Colgate, C. P. Comeau, D. P. Kelley, P. D. Payne, and S. W. Moulton, *Philco Corp., Philadelphia, Pa.*

This paper covers some recent improvements in the Apple tube. Significant improvements in brightness and color saturation resulting from the use of unequal color line widths are described. Contrast improvements obtained by better black guard line materials are discussed. A description of changes in screen and index geometry which have eliminated certain receiver functions is included in the paper. Long-term life test results are presented.

### 38.5. An Advanced Color Television Receiver Using a Beam Indexing Picture Tube

R. A. Bloomsburgh, A. Hopengarten, R. C. Moore, and H. H. Wilson, *Philco Corp., Philadelphia, Pa.*

This paper describes a new receiver using a beam indexing display and includes circuit construction, and performance data. The receiver is designed to utilize the picture tube described in the preceding companion paper. Important design criteria will be discussed together with circuit examples. Complete information on receiver alignment is included.

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## SESSION 39*

**WED. 2:30–5:00 P.M.**

**NEW YORK COLISEUM MARCONI HALL**

### Microwaves I—Components

*Chairman:* A. G. Clavier, *Federal Telecommunication Labs., Nutley, N. J.*

### 39.1. A Broadband Fixed Coaxial Power Divider

J. Reed and G. J. Wheeler, *Raytheon Manufacturing Co., Wayland, Mass.*

The design of a coaxial line power divider of any fixed power division ratio and maximally-flat performance is described. The performance of an actual model is tested over a 30 per cent band by the Deschamps method as well as other schemes and a comparison of these schemes described. The design includes a stub support so the device will stand high power. A refinement to the existing theory is discussed.

### 39.2. Broadband Waveguide-to-Coaxial Transitions

G. J. Wheeler, *Raytheon Manufacturing Co., Wayland, Mass.*

This paper describes three transitions from waveguide to coaxial line, each of which covers a frequency range of more than 30 per cent. These include:

1) A transition from WR 159 waveguide to 50 ohm coaxial line with a vswr $\leqq$1.14 from 4400 to 7000 mc. This is a "normal" transition in that the coaxial line feeds into the center of the broad face of the waveguide.
2) A colinear end-on transition from RG 48/U waveguide to standard 50 ohm $\frac{7}{8}$ inch coax with vswr $\leqq$1.10 from 2600 to 3700 mc.
3) A colinear end-on transition from 2.840 $\times$0.875 inches id waveguide to standard 50 ohm $\frac{7}{8}$ inch coax with vswr $\leqq$1.13 from 2600 to 3800 mc.

### 39.3. Transmission Properties of Hybrid Rings and Related Annuli

H. T. Budenbom, *Stavid Engineering, Inc., Plainfield, N. J.*

Advances made in the transmission analysis of conventional hybrid rings since their invention are summarized. Some new properties of hybrid rings and related annuli are noted. Some new hybrid rings nonconventional in number of taps and/or perimeter, are described and analyzed. A method of using hybrid rings in tandem to attain improved conjugacy bandwidth and depth is given. Experimental data are included.

### 39.4. Development of Circularly Polarized Microwave Cavity Filters

C. E. Nelson and W. L. Whirry, *Hughes Aircraft Co., Culver City, Calif.*

The use of circularly polarized microwave cavities leads to a simplification in the art of microwave filter design. Only one cavity (wtih two coupling apertures) is required to produce a four-port waveguide filter that is a reflectionless diplexer (*i.e.*, bandpass and/or band elimination). A rectangular waveguide with one coupling aperture can excite a circularly polarized cavity mode by $H_z$ and $H_z$ or $E_y$ and $H_z$ coupling. The latter type of coupling can be accomplished through apertures in the side walls of a circular cylinder cavity, freeing the end plates for tuning and temperature compensation. For commercial use of these filters the $TE_{112}$ cavity mode was chosen as the optimum degenerate mode on the basis of minimum cavity wall losses and maximum separation of unwanted cavity modes. Experimental test results at $X$-band frequencies are included for several of these cavities.

Several novel filters are introduced which use 3-db waveguide couplers and two apertures to excite a circularly polarized cavity mode. These filters are also reflectionless and have the same general properties as the other circularly polarized cavity filters.

### 39.5. Design of Improved Microwave Low-Pass Filters Using Strip-Line Techniques

Lt. R. A. Van Patten, *Rome Air Development Center, Griffiss Air Force Base, N. Y.*

* Sponsored by the Professional Group on Microwave Theory and Techniques. To be published in Part 1 of the 1957 IRE Convention Record.

Low-pass filters find application as components in narrow band transmitting, receiving and test equipment to control spurious responses, harmonic output, and interference. Strip-line low-pass filters are easily designed and give very satisfactory performance at a fraction of the cost of other types in the frequency range from 500 to 7000 mc. This report gives a design procedure for strip-line low-pass filters including two improvements over conventional designs.

It is shown that by cutting away a small portion of the dielectric material surrounding the fine conductor in the high impedance or inductive sections of the filter the stop band width can be effectively extended by as much as 100 per cent. A formula for the characteristic impedance of the resulting two dielectric, high impedance transmission line structure is derived. A method not involving stubs for realizing $m$-derived sections is given. The resulting $m$-derived sections have the advantage of smaller size and do not become antiresonant at approximately twice the cutoff frequency as do the quarter wavelength open circuited stubs commonly used. An approximate method for taking fringing capacitance into account is given. The problem of spurious responses caused by waveguide mode propagation is discussed and two solutions are offered.

### 39.6. Broad-Band Frequency Stabilization of a Reflex Klystron by Means of An External High "Q" Cavity

M. Magid, *Hughes Aircraft Co., Culver City, Calif.*

The cavity frequency stabilization system operating over narrow frequency bands described previously has been extended for 40 per cent bandwidth operation in rectangular waveguide. The development of a variable coupling type high "$Q$" cylindrical stabilization cavity and a compact phase shifter-mode suppressor required for the system is discussed. These components operate over the entire bandwidth of $1\times\frac{1}{2}$-inch waveguide. The system has been applied to two klystrons differing electrically. An improvement in the frequency stability of these klystrons by factors between 4 and 40 has been attained, with a small reduction in power output. The frequency stabilization system can be applied to most klystrons operating into $1\times\frac{1}{2}$-inch waveguide.

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## SESSION 40*

**WED. 2:30–5:00 P.M.**

**NEW YORK COLISEUM FARADAY HALL**

### Production Techniques

*Chairman:* J. J. Farrell, *General Electric Co., Syracuse, N. Y.*

### 40.1. The 14E Automatic Component "Wire-Wrap" Machine

H. F. Wilson, *Gardner-Denver Co., Grand Haven, Mich.*

Acceptance of solderless connections made by hand-held "Wire-Wrap" tools in the electronics industry has paved the way for a more automatic means of fabricating circuits. This method fastens pig-tail components by forming and wrapping the leads simultaneously, eliminating further operations such as lead straightening, cutting, and soldering.

By eliminating heat from the process and gripping components by the leads only during handling and wrapping, even the most fragile components are protected from damage during the process.

### 40.2. An Approach to Airborne Digital Computer Equipment Construction

P. E. Boron and E. N. King, *Hughes Aircraft Co., Culver City, Calif.*

This paper is a discussion of one method of building airborne digital equipment, making use of the unitized etched wiring plug-in philosophy and utilizing an all-etched wiring harness to make the large number of connections between plug-in units. Points of emphasis are miniaturization, reliability, small weight, accessibility, and manufacturability of the equipment.

### 40.3. An Automatic Dip Soldering Machine

V. O'Gorman, *United Shoe Machinery Corp., Beverly, Mass.*

An automatic high-production dip-soldering machine for printed circuit boards is described. The machine fluxes, preheats, dipsolders, cleans the residual flux from the boards, and unloads automatically at a rate of 1000 per hour.

The physical principles of the soldering process are discussed. The influence of these principles on the basic design of the machine and the adjustments required to effect the proper balance of the several process variables are described, as well as the experience obtained from its use in production.

### 40.4. Encapsulation of Electronic Circuits

R. Calicchia, *Rome Air Development Center, Griffiss Air Force Base, N. Y.*

RADC is actively engaged in a program aimed at developing experimental design data for engineers confronted with problems of selecting proper encapsulents for electronic equipment. This paper will indicate the quantitative effects of the encapsulating media upon the electrical characteristics of the embedment. Of major interest, is the work initiated on electrical performance of resistors, capacitors, inductors, and simple circuits at frequencies up to 250 mc. It was necessary in this work to investigate the electrical and mechanical properties of various resins in order to select the most suitable encapsulents for specific applications.

### 40.5. Development of Interconnecting Wiring

D. J. Keller, *Sperry Gyroscope Co., Great Neck, N. Y.*

Old style bulky interconnection harnesses are now being superseded by streamlined miniature versions which compact an almost unbelievable footage of wire into a given space. The ramifications of the problem, particularly in the interconnection of miniature components, has led to a categorized approach, which includes coverage of electrical noise and shielding philosophy, as well as physical and mechanical problems which must be anticipated. In order to understand and fully appreciate the depth of this problem, a brief history of early wiring is presented with graphic illustrations showing the relative improvements in the manufacture of wire, cable, and cabling techniques. Actual models are displayed showing the simple to complex relations which have evolved. As an aid to the systems engineer, a harness preparation sequence is shown, step by step, with an analysis of the problems which are introduced as the interconnections become more complex. Parallel items such as ground points, special connectors, adapters, markers, color coding, current carrying capacity, harness tapes, etc. are discussed with visual examples of each category. A prediction for the future is postulated.

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## SESSION 41*

**THURS. 10:00 A.M.–12:30 P.M.**

**WALDORF-ASTORIA STARLIGHT ROOF**

### Electronic Computers II—Symposium on Computers in Simulation, Data Reduction, and Control

*Chairman:* R. D. Elbourn, *National Bureau of Standards, Washington, D. C.*

A rapidly expanding class of computer applications that includes the simulation of complex systems in real time, on-line reduction of experimental data, and automatic process control is introducing new problems that were not met in the mathematical and business applications.

Problems of speed and control affect the design of the computer itself. For sufficient speed it may have to be wholly or partly analog or, if digital, it may have to be a special purpose rather than general purpose machine. The control may have to permit interruption for special tasks and then return automatically to a former task.

Communication between the computer and the rest of the system may involve problems of conversion between analog and digital data, of multiplexing many data channels into one computer channel, of smoothing or interpolating sampled data, or of preparing displays suitable for human beings.

The first two speakers will deal generally with these problems, then three speakers will describe their solutions in three specific applications:

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* Sponsored by the Professional Group on Electronic Computers. To be published in Part 4 of the 1957 IRE Convention Record.

1) A large combined analog-digital simulator.
2) The digital control of machine tools.
3) The reduction of wind-tunnel data.

After these talks there will be a round table discussion of present and future solutions of these problems.

# SESSION 42*

## THURS. 10:00 A.M.–12:30 P.M. WALDORF-ASTORIA ASTOR GALLERY

## Circuit Theory II—Transistor and Amplifier Circuit Design

*Chairman:* J. BRAINERD, *University of Pennsylvania, Philadelphia, Pa.*

### 42.1. Some Useful Techniques for Overcoming Frequency Limitations of Distributed Amplifiers

P. H. ROGERS, *University of Arizona, Tucson, Ariz.*, AND B. F. BARTON AND L. A. BEATTIE, *University of Michigan, Ann Arbor, Mich.*

A major limitation on the high-frequency capability of distributed amplifiers results from the presence of grid and plate lead inductances. These inductances approach series resonance with the input and output tube capacities of the tube, respectively, as the operating frequency range is increased. Negative mutual cancellation becomes increasingly difficult as the series inductance of the line sections is reduced in the interests of bandwidth. It is shown that the frequency capability of conventional circuits may be surpassed through the use of dummy line sections between tubes. It is shown that a constant $-K$ line gives nearly the maximum obtainable bandwidth. An amplifier with $4\times150$ A tubes and a cutoff of 490 mc is described.

### 42.2. Regeneration Effects in Double Tuned Band-Pass Amplifiers

P. BURA, *Canadian General Electric Co., Ltd., Toronto, Canada*

The amplitude and delay response of maximally flat band-pass amplifiers with double-tuned interstage circuits are distorted by the presence of unwanted feedback through the plate-to-grid capacitance.

It is shown that the amount of distortion introduced depends on a regeneration coefficient $\alpha$. When $\alpha$ is sufficiently large oscillations will result. The critical value of $\alpha$ will be calculated for the double-tuned interstage circuits and compared with that for the single-tuned circuits.

The effect on the delay response is calculated and methods of flattening the delay and amplitude response are discussed

### 42.3. A New Junction-Transistor High-Frequency Equivalent Circuit

R. D. MIDDLEBROOK, *California Institute of Technology, Pasadena, Calif.*

A small-signal equivalent circuit for a junction transistor is presented which is applicable to alloy or grown types of *p-n-p* or *n-p-n* transistors, and which is valid from dc up to twice alpha cutoff. The equivalent circuit is in the form of four short-circuit admittances, each of which can be represented by a simple network of lumped elements constant with frequency. The derivation is based on physical principles and takes into account base widening and collector barrier capacitance. Equations for the equivalent circuit element values are given either in terms of physical parameters or in terms of six practical measurements. The four-admittance representation is given both for common emitter and common base connections, and a relation between the common emitter and the common base cutoff frequencies is derived and experimentally verified. Measurements of the real and imaginary parts of the four admittances as functions of frequency for several transistors show excellent agreement with the values predicted by the equivalent circuit.

### 42.4. Circuit Applications of Semiconductor Junction Capacitance

F. H. DILL, JR. AND L. DEPIAN, *Carnegie Institute of Technology, Pittsburgh, Pa.*

This paper describes various applications of the voltage-dependent capacity of semiconductor junction diodes. The most promising applications seem to be those utilizing a voltage controlled sweeping oscillator. Some of these are frequency modulation, frequency control, search receivers, panoramic adapters, and spectrum analyzers in general. In these applications there are advantages to using the junction capacitance rather than reactance tubes or ferroelectric capacitances.

### 42.5. Pulse Circuit Applications of a New Semiconductor Device

R. A. STASIOR, *General Electric Co., Syracuse, N. Y.*

A new process high-frequency npn silicon transistor designed for military use offers characteristics well suited to pulse circuits. It permits high reliability, low cost direct coupled transistor logic circuitry operating at 500 kc. Other applications to be discussed will include multivibrators and pulse amplifiers.

### 42.6. Design of Junction Transistor Multivibrators by Driving-Point Impedance Methods

J. J. SURAN, *General Electric Co., Syracuse, N. Y.*

An analytical method for the design of junction transistor multivibrators by the use of two-terminal driving point techniques is presented. The synthesis method is based upon the point of view that the circuit design rather than the transistor properties should determine the operating characteristics of the multivibrator. Such factors as dc stability, parameter tolerances, frequency response characteristics, maximum switching speed, and trigger requirements are discussed.

It is shown that the two terminal design method may be used to
1) Completely determine by analytical means the dc design of the circuit for a given set of specifications,
2) Calculate the value of the coupling capacitors required to meet certain speed requirements,
3) Determine the mode of operation of the multivibrator, *e.g.*, astable, monostable or bistable,
4) Predict the maximum repetition rate at which the circuits may be driven or free-run,
5) Predict the minimum charge requirement for triggering the monostable or bistable circuits.

The design method reveals the differences which are important between silicon and germanium transistor multivibrators. In addition such factors as circuit efficiency as a function of temperature stability and parameter tolerances are discussed. Experimental verification of the theoretical results is presented.

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# SESSION 43*

## THURS. 10:00 A.M.–12:30 P.M.

## WALDORF-ASTORIA JADE ROOM

## Component Parts I

*Chairman:* L. KAHN, *Aerovox Corp., New Bedford, Mass.*

### 43.1. Ceramic Filter Capacitors for VHF and UHF

H. M. SCHLICKE, *Allen-Bradley Co., Milwaukee, Wis.*

"Feed Through" capacitors (FT) (fed through metallic shield) and "Stand-Off" (SO) capacitors (used without shield) have one electrode on ground while the other provides an undisturbed passage for dc or low frequency, but forms the short circuit terminal for very high frequency. They form an essential circuit element in grid and tank circuits and the dc supply points of amplifiers and oscillators, preventing undesired feedback, oscillations, interference, and radiation; or expressed positive terms, they are used for decoupling purposes.

At vhf and uhf, standard ceramic FT and SO, (nonceramic ones even more so) fail markedly to perform satisfactorily because of resonance (FT and SO) and because of inductive coupling effects (SO)

Generalizing from the superiority of discoidal (up to 3000 $\mu\mu$f only) over tubular FT's, the concept of minimizing or even eliminating

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phase shift to deresonate capacitors in the frequency range of interest is investigated systematically.

For the case of dielectrics alone, saw-tooth corrugations (unthreaded or threaded), untapered corrugations (impractical), and quasi-corrugations are evolved resulting in capacitors of arbitrarily high capacitance values and approaching the performance of ideal capacitors.

When ferrites are added, extremely simple new structures are obtained performing much better even than ideal capacitors, and if constructed as SO are also free of inductive coupling.

Finally a miniaturized, extremely effective low-pass filter for military equipment is described. It will replace bulky, unsatisfactory designs used so far.

Thus, a whole family of novel circuit elements has been developed having no presently available counterpart equivalent in performance or size. These elements will provide a definite improvement and simplification of commercial and military equipment working at vhf and uhf frequencies.

### 43.2. Manufacture and Measurement of Close Tolerance Temperature Compensating Ceramic Capacitors

N. Rudnick, *Glenco Corp., Metuchen, N. J.*

Temperature compensating ceramics from P100 to N750 are now blends of two parent compositions. Close tolerances derive from corrective additions of the appropriate parent to a presampled batch. A new glass coating has a reduced effect on the temperature coefficient and extends the temperature range. Coefficients are measured by an unique 1-mc bridge.

### 43.3. New Subminiature Metallized Paper Capacitors

P. P. Grad, *Aerovox Corp., New Bedford, Mass.*

A newly developed subminiature metallized paper capacitor is described. Its outstanding characteristics are a 50 per cent volume reduction from the smallest presently available units, a nearly flat temperature coefficient of capacitance, and very high insulation resistance even at elevated temperatures A new impregnating technique used for an unconventional impregnant for metallized paper capacitors is described.

### 43.4. The Use of Pulse Packages in Line-Type Pulsers

A. Luna, *Filtron Co., Inc., Flushing, N. Y.*

To be discussed is the natural development and the role of the pulse package in Radar transmitters today, and how in development work it materially reduces the time and effort required of the transmitter engineer—also how in production it can reduce the size and cost of the pulser and result in improved performance. The reasons for its even greater desirability in very narrow pulse work will be brought out.

To be discussed also is the role of the trigger pulse package, which can be of material help to engineers both in development work and in the product.

### 43.5. The Silver Oxide—Cadmium Alkaline Secondary Battery

P. L. Howard, *Yardney Electric Corp., New York, N. Y.*

A battery has been developed which substitutes cadmium for the zinc as negative active material in the silver-zinc secondary battery. By this change in negative the ability to accept CP charge and withstand over charging has been greatly improved.

This couple is more suitable for medium and low rate applications than for high rate use. The discharge voltage is approximately 1.2 volts per cell with 30 watt hours per pound and 1.5 watt hours per cubic inch. The cycle performance of this battery is rather uniform for several hundred cycles. Two to three thousand shallow cycles (50 per cent capacity) have been obtained from these batteries. The general characteristics of this system will be discussed.

## SESSION 44*

**THURS. 10:00 A.M.–12:30 P.M.**

**WALDORF-ASTORIA SERT ROOM**

**Industrial Electronics**

*Chairman:* E. W. Leaver, *Electronic Associates, Ltd., Willowdale, Ontario, Canada*

### 44.1. The Canadian Mail Handling System and the Problem of Coding

M. M. Levy and A. Barszczewski, *Post Office Dept., Ottawa, Ontario, Canada*

The Canadian Post Office has under development an Electronic Automatic Mail Sorting System. The efficient operation of the system is made possible by use of a proper code and coding methods.

The proper names of towns, villages, and streets are converted into a special code suitable for electronic handling.

The process of coding is performed mentally by the operators according to a set of simple rules. The addresses of letters are read by the operators in special reading stations and mentally coded, the code is then marked on the envelopes by a special keyboard. Afterwards the remaining operations are performed completely automatically.

The prototype of the system is briefly described. The paper is dealing mainly with the coding problem and human factors involved in it. A clear account is given of various codes and coding aspects. Description is given of minimization of errors. The realization of a working system did depend to a large extent on suitable coding and therefore its importance is stressed.

Although this paper is restricted to the specific application only, nevertheless the methods used have large potential possibilities in other fields where realization of electronic sorting and filing of data is essential.

### 44.2. Stabilized Magnetic Amplifier Circuits

H. W. Patton, *Airpax Products Co., Baltimore, Md.*

This paper describes dc to dc magnetic amplifier circuits which have effective input resistances of more than one million ohms and which can be made to have accuracies of better than a quarter of a per cent. These circuits are based on the potentiometric circuitry, a recent development in the magnetic amplifier art.

These newer potentiometric amplifier circuits have solved many of the tough design problems previously encountered, and have opened new paths towards designing reliable accurate, control and measurement equipments.

This paper reports recent work in this field, and by way of illustration shows what can be accomplished in terms of a typical design together with its actual performance characteristics.

### 44.3. New Techniques for the Control of Resistance Welding Machines

J. L. Solomon, *Sciaky Bros., Inc., Chicago, Ill.*

This paper describes a new and unique electronic control which has been applied to the control of the many functions required in the resistance welding process, and which has application to any system or machine requiring a series of timed controlled functions.

The control of resistance welding requires a series of accurately measured time intervals which follow in sequence for controlling the application of welding current and force to the material being welded. The timing circuits utilized in the past have definite shortcomings and limitations due to change in circuit components which affects the timing functions. A new concept of control is utilized, that of timing by counting cycles of the line frequency. This paper describes the use of the Dekatron glow transfer tube as applied to a system of control for resistance welding machines.

### 44.4. Pulse-Firing and Recovery-Time Characteristics of the 2D21 Thyratron

J. A. Olmstead and M. Roth, *Radio Corp. of America, Harrison, N. J.*

This paper describes the physical processes which occur within the 2D21 thyratron during pulse firing and recovery of grid control. Both of these characteristics depend upon retention or loss of grid control. This grid control is greatly reduced when positive ions are present within the tube. The positive ions surround the grid, forming a space-charge sheath which tends to neutralize the effect of external potentials applied to the grid.

In order to fire the 2D21 with short grid pulses (50 microseconds or less), the grid must be driven positive to speed the ionization process and the resulting loss of grid control.

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Pulse firing is primarily a function of anode voltage, grid voltage and the grid-pulse duration.

In order to recover control of the 2D21 thyratron after conduction, the grid-terminal voltage must be sufficiently negative to reestablish grid control. Recovery time is primarily a function of anode voltage, anode current prior to extinction, and grid-terminal voltage. The grid bias and grid-circuit resistance combined with the afterglow grid current determine the grid-terminal voltage.

Illustrative data for both of these characteristics are presented.

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# SESSION 45*

## THURS. 10:00 A.M.–12:00 NOON
## WALDORF-ASTORIA
## GRAND BALLROOM

## Reliability Programs

*Chairman:* J. R. Steen, *Sylvania Electric Products, Inc., Woburn, Mass.*

### 45.1. Air Force Ground Electronic Equipment Reliability Improvement Program

J. J. Naresky, *Rome Air Development Center, Griffiss Air Force Base, N. Y.*

As the demands of modern warfare have dictated increases in electronic equipment complexity, reliability has been correspondingly decreased, thus necessitating the institution of a comprehensive program of reliability improvement in the Department of Defense. This paper describes the factors leading to the development of the program at the Rome Air Development Center and the methods of its implementation on radar and other ground electronic equipment developed by RADC.

The RADC Reliability Program—past, present, and future—is discussed. The general areas covered are: 1) Methods of educating design engineers on reliability; 2) development, verification and use of a reliability predication technique; 3) development of automatic monitoring equipment; 4) component and circuit reliability improvement techniques, and 5) insertion of quantitative reliability requirements into equipment specifications.

### 45.2. A Reliability Program

R. E. Kuehn, *International Business Machines Corp., Owego, N. Y.*

The introduction of complex electronic systems into general usage has made the organization of reliability engineering groups a necessity. This group can function best by reporting to the manager of engineering as an independent evaluation and service agency. The tasks of reliability engineering include the development and application of reliability prediction techniques; the operation of a failure reporting and analysis system; the selection, qualification, and application of components; the environmental and life testing of components, units, and systems, and the evaluation of systems on the bench and in the field for reliability, maintainability, accuracy, and operational suitability.

### 45.3. A Reliability Program for R and D Projects

E. F. Dertinger, *American Bosch Arma Corp., Garden City, N. Y.*

The proposed paper will describe a long-range reliability program which has been inaugurated during the initial design stage of developing and producing an inertial guidance system for ballistic missile application. This program represents, to the writer's knowledge, the first all-out attempt to "design-in" complete missile system reliability.

Considerable effort will be extended toward presenting the procedures and techniques developed for "comparative evaluation" of functionally-suitable component parts and "reliability qualification" of: 1) parts selected for design incorporation, 2) assemblies and major components, and 3) complete systems. The philosophy of Arma management as regards reliability and the support given to this program will be described.

### 45.4. The Role of Quality Engineering in Procuring and Producing Reliable Products

R. A. Hulnick, *International Business Machines Corp., Kingston, N. Y.*

To assure production of a reliable product, quality engineering must formulate for the manufacturer a plan which encompasses a design of acceptable reliability, adequate controls over procurement and processing, and techniques for continuous performance evaluation for the purpose of product improvement.

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# SESSION 46*

## THURS. 10:00 A.M.–12:30 P.M.

## NEW YORK COLISEUM
## MORSE HALL

## Symposium—Digital Techniques for Problems in Telemetering and Remote Control

*Chairman:* C. H. Hoeppner, Jr., *Radiation, Inc., Melbourne, Fla.*

### 46.1. A High-Speed Digital Data-Handling System

G. F. Anderson, *Radiation, Inc., Melbourne, Fla.*

This paper describes a high-speed digital data-handling system which provides digitizing, editing, and processing of pwm and fm/fm data This system digitizes a maximum of 22 channels of analog data at a channel-sampling rate of 22,000 data points per second. An over-all accuracy of one part in 1024 is provided, utilizing a 10-bit binary straight code. Simultaneous sampling of the data to an accuracy of 33 microseconds is provided. Editing is provided by the Quick-Look analog recording system. The processing system generates an output code compatible with the IBM 650 computer; however, with slight alterations, the output format can be made compatible with other types of digital computers.

### 46.2. Magnetic Tape Playback and Digital Conversion of Telemetered Flight Data for Entry into Digital Computers

G. C. Dannals, *Radiation, Inc., Melbourne, Fla.*

Paper describes an automatic digital conversion and data processing system designed and built by Radiation, Inc. for the AC Spark Plug Division of General Motors Corp. Decommutated and separated pdm and fm/fm data recovered during magnetic tape playback of telemetry records is electronically sampled and converted to ten-bit binary code and further processed to suitable form for entry into high-speed digital computers. Data editing, intermediate recording, and processing control features are incorporated. Other salient features as regards system accuracy, high-speed handling, and operational flexibility are discussed.

### 46.3. Design Considerations for Super Speed Perforated Tape Digital Recording

J. Bellinger, J. T. MacNeill, and C. F. West, *Soroban Engineering Co., Melbourne, Fla.*

Until recently, techniques for recording digital data on perforated tape have been perfected to meet the requirements of the communication industry. The recorder described in this paper probably represents the first tape perforator designed for instrumentation as well as digital computer output data recording applications.

The design now permits recording of standard 5, 6, 7, or 8-hole code patterns to be reliably performed at controlled rates up to 240 codes per second. The perforator executes a basic

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* Sponsored by the Professional Group on Telemetry and Remote Control. To be published in Part 5 of the 1957 IRE Convention Record.

recording cycle of one code in approximately 4 milliseconds. Since a recording cycle can only be executed on demand associated electronic control circuits will permit recording of random or a synchronous data at variable rates up to 240/second.

### 46.4. A High-Speed Binary-to-Binary-Decimal Translator

C. A. CAMPBELL, *Radiation, Inc., Melbourne, Fla.*

A high-speed translator has been developed which accomplishes binary-to-binary-decimal translation yet requires only about three vacuum tubes per bit.

By suitably arranging the feedback and trigger logic circuits the binary decades accept counts with weights of 1, 2, 4, 6, or 8. With minor circuit changes the counter will operate with any integral weight from 1 through 9.

Arranging these decades in cascade, inputs with weights of 16, 32, 64, etc., are accepted and totaled to decimal readouts, which may be in either binary or decimal presentation.

The present decade requires about five microseconds per binary digit for computation; however, the state of the art can allow an increase of at least tenfold in this speed.

### 46.5. Simplicity for Reliable Low-Cost Operation in a Digital Data-Processing System

J. W. PRAST, *Bell Aircraft Corp., Buffalo, N. Y.*

A simple and straightforward digital data-processing system for telemetering purposes was developed in 1954 and has been in operation since that time for processing special precision information during test flights of guided missiles.

Experience gained in operational use will be reported with emphasis on reliability and cost aspects.

The system features 0.1 per cent accuracy, inherent reliability, and low cost on the advantageous side with limitations in versatility as a disadvantageous feature; it was originally designed for processing fm-type information essentially not changing over a sampling interval of approximately 50 milliseconds. Improvements and additional applications for processing information in analog-voltage form and fully automatic operation will be discussed.

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## SESSION 47*

**THURS. 10:00 A.M.–12:30 P.M.**

**NEW YORK COLISEUM MARCONI HALL**

### Millimicrosecond Instrumentation —Special Topics

*Chairman:* K. J. GERMESHAUSEN, *Edgerton, Germeshausen, and Grier, Inc., Boston, Mass.*

### 47.1. Millimicrosecond Photography with an Electronic Camera

R. C. MANINGER AND R. W. BUNTENBACH, *Precision Technology, Inc., Livermore, Calif.*

Photography of detonation propagation, and growth and decay of electrical discharges requires ultrahigh speed cameras. Such a camera and some unique problems associated with its development are described in this paper. The camera consists of a specially designed image converter tube, associated pulse circuitry, and oscilloscope recording camera. The tube with its associated circuitry acts both as a shutter and as the means for moving images across the face of the stationary recording film. The camera can take a multiple framed sequence with arbitrary time spacing between frames and with exposure times during each frame as short as twenty millimicroseconds.

### 47.2. An Image Converter for High-Speed Photography

R. G. STOUDENHEIMER AND J. C. MOOR, *Radio Corp. of America, Lancaster, Pa.*

A developmental image converter tube having electrostatic focus, a shutter grid, and electrostatic deflection is intended for multiple frame photography of high-speed events with exposures as short as ten millimicroseconds. The image converter and its operating characteristics are described.

### 47.3. A Millimicrosecond Pulse Generator Using Secondary Emission Tubes

J. A. NARUD, *Harvard University, Cambridge, Mass.*

This paper describes a regenerative pulse circuit using a single secondary emission tube that is able to generate pulses having a rise time of 6 mμs and a width continuously variable from 25 mμs to 12 μs. First, a theoretical discussion of the circuit is given in which expressions for pulse width and resolving time are derived. The effect of the magnitude and shape of the applied trigger signal on the response of the circuit is also considered. At the end of the paper various practical realizations of the circuit are presented. Among others these include a millimicrosecond pulse generator and a fast pulse-height discriminator.

### 47.4. A Fractional Microsecond Light Source of High Intensity

R. L. FORGACS, *Ford Motor Co., Dearborn, Mich.*

Flash photography of rapidly moving objects requires a triggered light source with high intensity and short duration. This is particularly important for Schlieren and interferometer photography of shock waves where only a small fraction of the generated light can be used. Analysis of the various methods for discharging a capacitor through a spark gap light source indicates the superiority of a scheme in which a fast rising capacitor voltage produces gap breakdown. The major disadvantage of this method is that variation in the firing voltage results in fluctuations in light intensity and firing delay. In the developed circuit an auxiliary triggered gap with a unique electrode arrangement is used both to illuminate the light gap and to initiate the capacitor charging. This illumination causes the firing voltage on successive discharges to be constant within ±2 per cent. The light pulse duration is 0.18 and 0.53 μs for one and 10 joules energy respectively.

### 47.5. The Electrograph

R. A. BRODING, J. C. WESTERVELT, AND J. D. SCHROEDER, *Tulsa, Okla.*

A general purpose oscillograph is described wherein a completely dry processing, including sensitizing, exposure, development, and fixing of the record, is accomplished within the instrument as a continuous process. Trace vector speeds in the order of 600 inches per second are possible. The Electrograph combines the features of the conventional photographic oscillograph and the direct-writing oscillograph into one package. Such features as the use of high sensitive pencil type galvanometers, daylight exposure of the supply and take-up rolls, photographic timing lines, rectilinear recording, ability for traces to cross one another, viewing of the recording, permanent fixing and the completely processed record available on completion of the recording are possible for the first time.

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## SESSION 48*

**THURS. 10:00 A.M.–12:30 P.M.**

**NEW YORK COLISEUM FARADAY HALL**

### Microwaves II—Switches

*Chairman:* A. G. FOX, *Bell Telephone Labs., Red Bank, N. J.*

### 48.1. Precision High-Speed Microwave Switch

W. E. FROMM, S. H. KLUG, AND K. S. PACKARD, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

A three-port, single-pole, double-throw switch designed to transmit microwave power from either of two ports to the third or vice versa. The difference in transmission (or the difference in attenuation) of either of the possible paths of transmission has a required maximum variation of 0.003 db over any 40-megacycle band in the frequency range over which the instrument is designed to operate. The corresponding phase differential has a variation of less than 0.05 millimicroseconds over a 40-megacycle band. The cross-talk between the two paths is down by more than 70 db. The maximum vswr seen at either of the input ports is 1.06.

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As presently designed, the instrument covers a 9 per cent frequency range in the *C*-band region. 10 db of attenuation exists through either path of transmission. The switching frequency is synchronous at 33 cycles per second. Seventy per cent of each cycle is useful time during which the transmission characteristics of one or the other of the two paths are constant and the vswr's are within their aforementioned limits. The remainder of the time is used as switching time.

In its present form the unit has waveguide input and output ports while the switching action is accomplished in strip line circuitry by means of a motor driven rotating attenuator.

### 48.2. Fast Acting Microwave Switch

H. H. WEICHARDT, *General Ceramics, Keasbey, N. J.*

This paper describes a novel mechanical microwave switch.

In its present embodiment the switch is primarily applicable to low-power circuitry. The switching time is of the order of one microsecond, and the stop-power attenuation is better than 60 db at all frequencies of the useful microwave spectrum. The intrinsic insertion loss in the pass-power condition is negligibly small, and the total insertion loss can be kept negligibly small over a very wide frequency band by the use of proper input and output transitions. The short switching time was obtained by using an elastic inner conductor in a slab or strip transmission line, which upon a small longitudinal displacement of one end experiences a much larger lateral displacement of the center of the section and thereby contacts the outer conductors causing a very effective short. The driving force is obtained from a greatly simplified relay structure.

Due to the very small motion of all moving parts and the absence of frictional joints, the switch has shown exceptional good long-life properties

### 48.3. High Speed Ferrite Microwave Switches

G. S. UEBELE, *Hughes Aircraft Co., Culver City, Calif.*

A microwave circulator utilizing ferrite elements can be converted into a fast acting microwave switch by placing the ferrite in a pulsed magnetic field. The operation of the circulator may be based on the Faraday rotation phenomenon or on the nonreciprocal phase shift observed in a ferrite loaded rectangular waveguide. Problems in the design of the microwave structure as well as the electronic driving circuitry are discussed. Various techniques for the reduction of eddy currents in the waveguide are presented and the performance characteristics of various switches given.

### 48.4. An L-Band Ferrite Coaxial Line Modulator

B. VAFIADES AND B. J. DUNCAN, *Sperry Gyroscope Co., Div. of Sperry Rand Corp., Great Neck, N. Y.*

A new *L*-band modulator is described which utilizes low saturation magnetization ferrites to achieve a high percentage modulation over an eight per cent bandwidth. The device is designed in $\frac{7}{8}$ inch coaxial line, and modulation is affected by alternately biasing the ferrite into—and away from—gyromagnetic resonance using a longitudinally applied magnetic field. A thin walled tube of ferrite aluminate is located concentrically on the coaxial line center conductor and its saturation magnetization is such that operation is from zero field to resonance. Two concentrically wound coils are used to independently achieve ferrite biasing at the midpoint of the resonance curve and to achieve amplitude modulation. Design theory leading to the development of the modulator is presented. Information is likewise included on the pertinent physical and electrical characteristics of the modulator.

### 48.5. Ferrite Microwave Detector

D. JAFFE, J. C. CACHERIS, AND N. KARAYIANIS, *Diamond Ordnance Fuze Labs., Washington, D. C.*

By inclusion of certain second-order terms in Polder's permeability tensor a nonlinearity results which predicts detection of an amplitude modulated microwave signal. The amplitude of the demodulated signal is a function of applied dc magnetic field and increases up to fields corresponding to ferromagnetic resonance where the variation is of the normal dispersion form. The detected low frequency field magnetostricts a ferrite rod the vibration of which is observed by means of a polarized $BaTiO_3$ ceramic rod bonded to the ferrite. The length of the composite rod is adjusted for mechanical resonance. Experimental results have verified the predictions of square law detection, dependence of the phase of the low frequency field on the sign of magnetization, and the monotonic increase of demodulation with increasing bias field.

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## SESSION 49*

**THURS. 2:30–5:00 P.M.**

**WALDORF-ASTORIA STARLIGHT ROOF**

### Electronic Computers III—Mainly Analog

*Chairman:* J. D. NOE, *Stanford Research Institute, Menlo Park, Calif.*

### 49.1. Computation with Pulse Analogs

N. RUBENFELD, *W. L. Maxson Corp., New York, N. Y.*

A special purpose semidigital computer has been built to handle a computation involving two analog frequencies and an analog voltage using novel techniques. The equation to be solved is

$$fx = \frac{E}{f_a} \times f_b.$$

The computation is done in two basic steps; the first step is computing a repetition rate $f_r = E/f_a$ and the second step is multiplying two frequency analogs, $f_r \times f_b$. The entire computer is built using basic blocks, such as, flip-flops, gates, blocking-oscillators, and is completely transistorized. Because of the flexibility of the basic concepts, the dynamic range of the computer is virtually unlimited.

### 49.2. A Cyclic Digital-to-Analog Decoder

G. H. MYERS, *Rome Air Development Center, Griffiss Air Force Base, N. Y.*

This paper describes a counting decoder which converts a binary number into a duration-modulated pulse many times in a computing cycle without sacrificing the economy of an ordinary, basic decoder. The device, which is part of the TRADIC computer, uses only transistors and achieves its features by using the "cyclic" nature of binary numbers. That is, if a binary number is placed in a register and a fixed amount is continually subtracted from it, the original number will reappear after a constant time interval. Decoding the same number many times has the effect of giving the decoder a considerably greater bandwidth, which is an important consideration in control or servomechanism problems. The decoder also has certain other features which make it desirable in control applications.

### 49.3. An Automatic Analog Computer Method for Solving Polynomials and Finding Root Loci

L. LEVINE AND H. F. MEISSINGER, *Hughes Aircraft Co., Culver City, Calif.*

Various analog computer techniques are available which provide the means for rapid solution of polynomials with an accuracy suitable for engineering purposes. Methods heretofore described in the computer literature, however, have the shortcoming of requiring a step-by-step procedure for finding the roots.

In this paper a new analog technique is discussed by which the roots are determined automatically. This is an application of the "method of steepest descent." A suitably chosen function $W$ is minimized in the computer by a continuous adjustment of the coordinates of a point in the complex $z$ plane until a stable equilibrium is reached. This corresponds to a root of the polynomial. In a polynomial of $n$th degree $n$ such equilibrium points can be found. The coordinate adjustment follows the gradient of the function $W$ and therefore leads to the minimum at the fastest possible rate.

The computer may be operated in two modes: 1) searching for individual roots from points arbitrarily chosen in the complex $z$ plane, 2) tracking the roots while coefficients of the polynomial are being varied. The latter mode is ideally suited for plotting root loci in the complex frequency plane.

This paper contains a discussion of mathematical and practical aspects of the steepest descent method and gives several alternative computer circuits. Illustrative examples of steepest descent paths and root loci obtained on the computer are included.

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### 49.4. Magnetically Controlled Counters

E. A. SANDS, *Armonk, N. Y.*

A magnetically controlled counter will be described in which the count determining circuit is a pair of magnetic cores. A simple theoretical analysis will be made using the principles of equivalent core impedance. Some practical circuits will be shown, and deviations from predicted behavior will be discussed. Methods of designing units to produce reliable counts from scale 2 to scale 16 will be indicated. Means of presetting will be pointed out, and practical limitations on the use of the device will be explained.

### 49.5. Systematic Tracing of Discrepancies in Analog Computers

M. PLOTKIN, *Naval Air Development Center, Johnsville, Pa.*, AND E. GROSSWALD, *University of Pennsylvania, Philadelphia, Pa.*

Large analog computers are for the most part used in control, or closed loop, problems. In problems of this sort an error in one location, whether due to incorrect plugging or to a faulty component, causes errors throughout the machine. Should the errors result in observed discrepancies, by comparison of the machine output with a reference solution either obtained independently or produced earlier by the same computer, it is sometimes difficult in closed loop problems to find the cause. This paper proposes a method for locating systematically the source of such discrepancies.

## SESSION 50*

THURS. 2:30–5:00 P.M.

WALDORF-ASTORIA
ASTOR GALLERY

### Circuit Theory III—Network Design Techniques

*Chairman:* L. A. ZADEH, *Institute for Advanced Study, Princeton, N. J.*

### 50.1. Pulse-Forming Networks Approximating Equal-Ripple Flat-Top Step Response

A. D. PERRY, *Bell Telephone Labs., Inc., Murray Hill, N. J.*

A simple iterative procedure of analog computation yields a set of two-terminal networks whose step responses approximate flat-top pulses in an approximately equal-ripple manner with $\pm 0.1$ per cent ripple. Digital computations and experimental studies verify the designs.

The approximation is carried out directly in terms of element values for a conventional negative-mutual PFN. The procedure is flexible enough to allow such additional constraints as equality of capacitors and/or simple nonlinear loads. Designs have also been carried out in which secondary and tertiary couplings were prescribed by the configurations of the coils. Specified, finite $Q$'s are easily accommodated.

The singularity patterns of the networks have an intriguing regularity which has not yet been adequately exploited.

### 50.2. Interstage Network Design with Practical Constraints

B. F. BARTON, *University of Michigan, Ann Arbor, Mich.*

A common method of interstage design, based on prototype circuits with all zeroes of transfer at infinity, leads to low-pass and band-pass circuits containing inductances without shunt capacitance. These coils and therefore the theoretical response become increasingly difficult of realization as the interstage impedance level is raised in the interests of gain. The present paper demonstrates design procedures leading to circuits in which all coils have shunt capacitance. The resulting networks may have either maximally flat or Tchebycheff pass band shapes. The networks have zeroes of response at real frequencies which may be useful in filter applications. Measured characteristics illustrating the advantages of these circuits are presented. Design curves are included.

### 50.3. Synthesis of Lumped Parameter Precision Delay Line

E. S. KUH, *University of California, Berkeley, Calif.*

In the design of a delay line one can break down the problem into two parts: first, to provide the required time delay and bandwidth with a network as simple as possible, and secondly, to have a good time response. This paper presents such a design technique for precision time domain applications. The network obtained is a tandem connection of a low-pass ladder which provides the shape of the time response and all pass bridge structures which give the desired time delay.

The approximation to the ideal delay function is based on the potential analog method in both cases. A method of estimating the time domain error from the frequency domain error is used to determine the maximum permissible phase deviation from linearity of the all pass network.

### 50.4. On the Approximation Problem in Filter Design

A. PAPOULIS, *Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

The problem of determining the network function of a filter of a certain order with an amplitude characteristic having the smallest possible "cutoff" interval for a given variation in the stop-band and pass-band, is considered. It is shown that the optimum characteristic has the equal ripple property in the stop band and the pass band, and that it is unique. From the uniqueness a simple relationship between the zeros and poles of the corresponding network function is established, and a method results for the determination of this function. The reduction in the cutoff interval, gained by limiting the stop band, is evaluated.

### 50.5. Recent Advances in the Synthesis of Comb Filters

W. D. WHITE AND A. E. RUVIN, *Airborne Instruments Laboratory, Inc., Mineola, N. Y.*

Recent developments in radar and allied fields have placed increasing emphasis on the demand for comb filters having carefully controlled characteristics. This paper reviews the synthesis procedures used in the design of such filters including an experimental verification of certain selected examples. The techniques shown are applicable not only to MTI filters having controlled velocity response, but also to video integrators for signal-to-noise enhancement. Emphasis is placed on an analogy between comb filters and conventional low-pass, and high-pass filters. This analogy permits one to draw on the wealth of information in the conventional filter design art.

### 50.6. Explicit Formulas for Tchebycheff and Butterworth Ladder Networks

L. WEINBERG, *Hughes Research Labs., Culver City, Calif.*

Green found the closed-form formulas for the element values in a resistance-terminated ladder network that has a maximally-flat (Butterworth) or equal-ripple (Tchebycheff) characteristic. These formulas apply only when all the zeros of the reflection coefficient $\rho$ are chosen to lie in one half-plane. For $D=1$ and for $n$ odd, a matched symmetrical network is obtained by the use of Green's formulas. For the case of $n$ odd and any value of input-to-output resistance ratio, new formulas have been found for the element values when the zeros of $\rho$ are chosen to alternate in the left and right half-planes. The networks obtained are related to the symmetrical ones given by Green's formulas for $n$ odd and $D=1$.

## SESSION 51*

THURS. 2:30–5:00 P.M.

WALDORF-ASTORIA
JADE ROOM

### Component Parts II

*Chairman:* A. W. ROGERS, *Signal Corps Eng. Labs., Fort Monmouth, N. J.*

### 51.1. Thermistors for the Gradual Application of Heater Voltage in Thermionic Tubes

J. J. GANO AND G. F. SANDY, *Massachusetts Institute of Technology, Lincoln Lab., Lexington, Mass.*

Thermistors, which are thermally-sensitive resistors having large negative temperature coefficients of resistance, can be aptly used for

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* Sponsored by the Professional Group on Component Parts. To be published in Part 6 of the 1957 IRE CONVENTION RECORD.

the gradual application of heater voltage in thermionic tubes, thus diminishing thermal transients and reducing mechanical failures. Commercially available units in washer form and properly mounted in stacks can accommodate heater loads up to 1200 watts. One stack can be used to limit the current at any instant to less than 120 per cent of normal for a wide range of loads. The method of determining the rating of a thermistor and the number required in series connection is also presented.

### 51.2. New Levels of Performance for General Purpose Resistors in Army Applications

R. A. OSCHE, *Signal Corps Engineering Labs., Fort Monmouth, N. J.*

A critical life evaluation, based on 7,000,000 resistor-hours, will be reviewed, together with the electrical, mechanical, and environmental attributes and the performance characteristics and deficiencies of general purpose (carbon composition) fixed resistors. It will be shown how new levels of performance for general purpose resistors, via Signal Corps development effort, have been obtained without sacrificing the size and mass producibility features of the composition types. The new performance level evaluation is based on 15,000,000 resistor-hours.

### 51.3. Measurement and Effects of Error Rate in Precision Potentiometers

S. B. RASMUSSEN, *Helipot Corp., Newport Beach, Calif.*

Linearity tolerance, the usual designation for potentiometer accuracy, is inadequate for predicting accuracy of rate systems or servo differentiators. A new figure of merit, called error rate tolerance, is proposed. Mathematically, error rate is a derivative of linearity error.

Several means for measuring error rate have been investigated, including a servo differentiator, which has proved superior in laboratory tests.

Inherent damping in a servomechanism tends to attenuate errors due to error rate. In a position servo, high error rates may lead to intermittent instability.

The figure for error rate is useful in estimating the frequency content of the linearity error signal, and in predicting variations in the loop gain of servomechanisms.

### 51.4. Theory, Measurement and Reduction of Precision Potentiometer Linearity Errors

F. FRITCHLE, *Helipot Corp., Newport Beach, Calif.*

Conventional engineering has in the past reduced potentiometer errors to a great degree. However, future reduction of residual errors will be accomplished only by analytical study of these errors.

There are four possible sources of linearity error in a potentiometer:

1) resistance wire nonlinearity;
2) machine winding errors;
3) mechanical or potentiometer geometry errors; and
4) electrical loading error from internal leakage.

Each of these errors is discussed and equations are given, the study of which indicates the parameters that are most critical.

Photos and typical recordings are shown of three newly developed machines for measuring these errors.

Methods of reducing each of these errors are discussed.

### 51.5. Vibration and Shock Resistant Relay Designs

A. P. BOYLAN, *Signal Corps Eng. Labs., Fort Monmouth, N. J.,* AND J. L. PFEFFER, *Struthers-Dunn, Inc., Pitman, N. J.*

Subminiature general purpose and sensitive relays combining the desirable features of ruggedness, compactness, contact multiplicity, high contact current capacity, vibration, shock, and centrifugal acceleration resistance, heretofore unavailable in commercially procureable relays, have been developed under the recently completed Signal Corps program with Struthers-Dunn, Inc. These structures were designed specifically to fill an urgent need of missile and airborne equipment manufacturers, for sensitive and general purpose relays.

This paper will discuss the development goals of this program, the approach employed by the contractor, the problems encountered and the eventual solutions, the design parameters established as the best means of satisfying the severe environmental requirements, the performance thresholds of the developed structures, and the contributions which this program has made in advancing the state of the relay art.

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## SESSION 52*

**THURS. 2:30–5:00 P.M.**

**WALDORF-ASTORIA SERT ROOM**

### Analysis and Techniques for Improved Reliability

*Chairman:* G. B. CARTER, *International Business Machines Corp., Kingston, N. Y.*

### 52.1. Guided Missile Reliability vs Complexity

S. W. LICHTMAN, *Hughes Aircraft Co., Culver City, Calif.*

Reliability by the Poisson law of random chance failure is related to the exponential product of number of component parts in series and average failure rate per component part. Attempts have been made to express missile reliability in terms of total number of parts and in terms of key functional parts, corresponding to highest failure rates. The first, strictly complexity approach, leads to uncertainty regarding how exhaustive to make the parts count and to ambiguity concerning the lumping of similar parts into manageable categories. The latter approach yields results compatible with the first method but avoids these ambiguities. Since key functional parts are related to excellence of engineering and to production accomplishments, missile reliability is intrinsically more identified with product maturity than with total parts count.

### 52.2. Narrow Limit Gage Sampling Procedure

H. G. HARDING AND S. PRICE, *International Business Machines Corp., Kingston, N. Y.*

Manufacturers report that narrow limit (NL) gage sampling procedures are gaining in importance as the preferred method for quality-sampling of machined piece-parts. This advance in popularity of the NL technique is attributed to the increased sensitivity of measurement, and the saving in time, as compared with the normal sampling procedure specified in MIL STD 105A. In the "Proposal for a Narrow Limit Gage Sampling Procedure," the NL Technique is applied, for the first time, to the requirements of Incoming Inspection. By means of typical operating characteristic (OC) curves there is demonstrated that the NL technique, as applied to incoming inspection. effects the following results: 1) reduces incoming-inspection time by 50 per cent, 2) guarantees a quality assurance that is equal to MIL STD 105A, 3) rejects are disclosed at a more rapid rate than is true of normal sampling procedures, and 4) provides a concrete means for establishing a vendor-rating. In the presentation of the "Proposal for a Narrow Limit Gage Sampling Procedure," the applied technique is particularly slanted to incoming-inspection processes covering large lots of electrical and electronic components.

### 52.3. Increased Reliability through DC Overpotential Testing of Electronic Components

V. WOUK, *Beta Electric Corp., New York, N. Y.*

The use of high voltage dc for checking the insulation characteristics of electronic components, particularly those used in high voltage circuits, is shown to have important advantages over ac testing.

By means of proper interpretation of resisttance characteristics, component failures can often be anticipated. The technique of testing with high voltage dc, which is of growing importance in the electric power field, is offered for consideration in the electronic field.

Specific applications to capacitors are recommended as a worthwhile project for the radio-electronic industry to consider. Some typical dc test equipment is described.

### 52.4. Reliability Prediction Technique for Use in Design of Complex Systems

H. E. BLANTON, *Hycon Eastern, Inc., Cambridge, Mass.*

A relatively simple reliability-prediction technique has been developed which can be instituted during the early stages of the design of complex systems. Through the use of *reliability diagrams* (obtained by modifying engineering block diagrams) and basic rules from

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probability theory, ***reliability formulas*** for proposed designs are derived. Effects of secondary failures and variations in requirements for successful performance are included. By evaluating the reliability formulas using the best available component-reliability estimates, alternative designs can be compared and the need for redundancy or component improvement is established. An example based on an airborne telemetering system illustrates the technique.

### 52.5. Analysis and Engineering Study of the Environmental Vibrations and Shock Characteristics of a New Military Airborne Gimbaled Equipment

D. EHRENPREIS, *David Ehrenpreis, Consulting Engineers, New York, N. Y.*

The outline of the analysis, engineering study, and theoretical investigation of the steady-state vibrations and sudden impulse shock dynamic characteristics of a new military airborne system are presented. Certain important natural frequencies and responses due to dynamic inputs are determined based upon this analysis of the complex framework.

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## SESSION 53*

**THURS. 2:30–5:00 P.M.**

**NEW YORK COLISEUM MORSE HALL**

### Symposium—Low Level Multiplexing for Telemetering and Remote Control

*Chairman:* C. H. DOERSAM, JR., *Sperry Gyroscope Co. Inc., Great Neck, N. Y.*

### 53.1. Low Level Signal Multiplexing

D. W. HILL AND A. S. WESTNEAT, JR., *Applied Science Corp. of Princeton, Princeton, N. J.*

This paper discusses system criteria for multiplexing a group of balanced milivolt data sources, with particular emphasis on commutator performance and on amplifier characteristics. Certain practical vacuum tube circuit applications are described, while development trends, involving greater uses of transistors, are discussed.

### 53.2. A Low-Level Electronics Switch

E. DORSETT, *Radiation, Inc., Melbourne, Fla.*

Fundamental limitations in certain electronic switching methods render them completely unsatisfactory for applications in the microvolt region. The development of a suitable switching method employing solid state components involved investigation of numerous alternative methods. Environmental requirements, miniaturization, stability, linearity, and efficiency were prime considerations in the development, which resulted in a switch with a wide variety of possible applications. The switch developed was evaluated in the Laboratory and under operational considerations as a component in an airborne telemetry system. Data assembled during these evaluations are presented, along with photographic slides and a display of the equipment.

### 53.3. A Unique Wide-Band Transistorized Pulse Amplifier

W. T. EDDINS, *Radiation, Inc., Melbourne, Fla.*

A transistorized amplifier for feeding analog voltage samples into a high-speed analog-to-digital converter is described. Due to the nature of the signal, a dc amplifier is found necessary, but good transient response, stability of gain, and low dc drift must be maintained. To provide good isolation, a high input impedance must be obtained, but output impedance must be kept low.

A unique method of applying negative feedback is used to obtain stability of gain, low output impedance, and wide bandwidth, and yet maintain exceptionally high input impedance. DC drift is minimized through the use of balanced input circuits, a thermistor, and a novel method of maintaining constant temperature about transistors.

The impact of transistors upon the field of dc amplifiers is discussed.

### 53.4. Completely Transistorized Strain Gage Oscillator

W. H. FOSTER, *Electronic Eng. Co. of California, Los Angeles, Calif.*

Associated with flight testing of aircraft and missiles is the problem of telemetering and/or recording stress information. Such data are usually gathered by a resistive type strain bridge transducer that converts stress variations to voltage level changes. The output of such a bridge is not directly compatible with fm telemetering or magnetic tape recording equipment. Accordingly, it is necessary to employ a converter for changing amplitude varying data to frequency varying data, thus allowing fm telemetering or magnetic tape recording of the data.

Described herein is a completely transistorized strain gage oscillator for accomplishing this conversion. This unit has been designed and tested in the laboratory, and the first units are now under construction. Its primary use is in conjunction with a resistive type strain bridge. The frequency of oscillation of the strain gage oscillator is caused to deviate as a linear function of bridge strain by injection of the bridge output signal into the local feedback loop at a point where it is in quadrature phase with the oscillator signal. The transistorized strain gage oscillator should prove to be a very practical working unit which occupies a volume of only 3.5 cubic inches and which operates over a broad ambient temperature range with long life expectation. The prototype has undergone extensive temperature stability testing throughout the temperature range from −55°C to +100°C. The specifications of the transistorized strain gage oscillator are comparable to those of high quality vacuum tube predecessors, but a distinct improvement over vacuum tube models is realized in the reduction of power consumption by a factor of 95 per cent.

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## SESSION 54*

**THURS. 2:30–5:00 P.M.**

**NEW YORK COLISEUM MARCONI HALL**

### Instrumentation II

*Chairman:* W. H. FENN, *Hughes Aircraft Co., Culver City, Calif.*

### 54.1. Low Level Transistorized DC Amplifier With Improved Stability

A. WARNICK AND C. N. SAVAGE, *Ford Motor Co., Dearborn, Mich.*

An analysis of the critical factors in the design of differential-transistor amplifier with dc stability and noise factor suitable for use with microvolt signals is presented. A successful design using two differential stages and a complementary output stage is described. The problems encountered in the selection of appropriate components are discussed. An analysis and detailed performance data are presented for the amplifier without feedback, and with multiloop feedback. The procedure for measuring the data presented is outlined.

### 54.2. A Unique Standard-Frequency Multiplier

J. K. CLAPP AND F. D. LEWIS, *General Radio Co., Cambridge, Mass.*

A 10,000:1, four-stage multiplier utilizing quartz-controlled triode oscillators and an external cavity-controlled klystron oscillator as highly selective narrow-band filters is described.

Results of experimental tests of performance of phase-locked quartz-crystal and cavity klystron oscillators are given. The remaining phase jitter and noise are far below the levels found in usual multiplier circuits.

Phase comparison tests of two such multipliers operating from a single 100-kc frequency standard are described, and results obtained are given in photographs of the cro display.

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A novel method of separating and observing the effects due to hum frequencies is briefly described. Results are shown in photographs.

### 54.3. Measurement of the Complex Permeability of Magnetic Materials Over the Frequency Range of 50 to 500 Megacycles

I. Bady and R. J. Franklin, *Signal Corps Eng. Labs., Fort Monmouth, N. J.*

Four methods for measuring complex permeability at very high frequencies are described. They are based on the use of a high-frequency $Q$ meter, cavity resonator, very-high frequency bridge, and slotted line. A feature common to all methods is that the sample holder is coaxial, and the sample is in the shape of a cylindrical ring. Test procedures and formulas are given. Sources of error and relative advantages of the methods are discussed. Measurements of several samples by the different methods show excellent agreement in almost all cases and the discrepancies which do occur are accounted for by sources of error discussed in the paper.

### 54.4. An Automatic Impedance Plotter Based on a Hybrid-Like Network with A Very Wide Frequency Range

C. B. Watts, Jr. and A. Alford, *Alford Manufacturing Co., Boston, Mass.*

In this paper is described an instrument which measures impedance and presents it in the form of a trace on a Smith Chart. Impedance plotters of this type work over wide frequency ranges, for example, 50–250 mc or 180–900 mc.

This automatic impedance plotter obtains its rf information from a hybrid-like network, which compares an unknown impedance with a standard impedance. A reference signal of constant amplitude and a smaller signal from the hybrid-like network are fed into a polar displayer. The smaller signal is essentially proportional to the reflection coefficient of the unknown impedance with respect to the standard load. The polar displayer is a network which converts the relative amplitude and the phase of the smaller rf signal into two audio frequency square waves which can be applied to the two axes of an oscilloscope or a two-axis recorder to obtain a direct polar display of the rf vector.

The automatic impedance plotter has been found to be very useful in the development of rf loads, switches, and of other devices which are intended to introduce small reflections. Reflections as small as those equivalent to 1.01 swr can be observed. All measurements can be readily referred to the characteristic impedance of a standard rigid line.

### 54.5. Automatic Indication of Receiver Noise Figure

A. J. Hendler and F. G. Haneman, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

A general technique which provides an automatic and continuous indication of receiver noise figure is described. A device utilizing this technique employs modulated gaseous discharge noise sources and post detection synchronous integration allowing a continuous and direct-reading indication of noise figure.

The advantages of the automatic noise figure indication over conventional manual noise figure measurements are discussed as well as the relative accuracy of the two measurements. Techniques for the determination of the optimum design parameters for crystal mixers, traveling wave tubes, and rf amplifiers are also outlined.

Using the same techniques an instrument was developed which provides a continuous indication of radar receiver noise figure during normal radar operation. A pulse modulated noise signal, synchronized with the radar trigger, is injected into the receiver at the end of alternate radar sweeps. The signal is attenuated by a directional coupler; the added noise therefore causes negligible system deterioration.

### 54.6. High Precision Sawtooth Generator

L. J. Torn, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

In the field of radar data processing applications frequently arise for ultraprecise single-scale timing devices. Typically, such requirements occur in the time demodulation process involved in the conversion of radar video time data into the equivalent range data.

Since early in World War II, sawtooth generators have been used as the time base for this demodulation process. These devices have previously been limited to accuracies in the order of one part in one thousand or less. Further: they require frequent recalibration and trimming, at best a tedious and difficult procedure.

Recent military requirements for higher accuracies, coupled with extreme reliability and stability, have led to a development program for a high-precision sawtooth generator. The new device was required to be accurate to better than one part in four thousand over long periods of time, without use of trimming controls of any type. In addition it was required that accuracy be maintained under field environmental conditions.

The theory and circuit development required for this new time-base generator are presented in this paper. As will be seen from the discussion, the nature of the requirements is such that the final development was forced to overcome the difficulties inherent in both high-stability dc amplifiers and high-frequency feedback amplifiers.

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## SESSION 55*

**THURS. 2:30–5:00 P.M.**

**NEW YORK COLISEUM FARADAY HALL**

* Sponsored by the Professional Group on Microwave Theory and Techniques. To be published in Part 1 of the 1957 IRE Convention Record.

## Microwaves III—General

Chairman: E. Weber, *Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

### 55.1. The Optimum Spacing of Bead Supports in Coaxial Line at Microwave Frequencies

D. Dettinger, *Wheeler Labs., Great Neck, N. Y.*

In coaxial transmission lines, the inner conductor is commonly supported by dielectric beads spaced at intervals along identical sections of line. At microwave frequencies, where the bead spacing is large compared with the wavelength, there result sharp peaks of reflection corresponding to reinforcement of the reflections of individual beads. Assuming that the reflections of the beads themselves have been reduced to a uniform minimum by careful design and production control, it is advantageous to disperse the beads in each section so as to reduce the cumulative reflection peaks. A dispersal has been found which is believed to yield the minimum value of peak reflection.

### 55.2. Multiple-Line Directional Couplers

J. P. Shelton, Jr., *Melpar, Inc., Falls Church, Va.*

The problem of many directionally coupled parallel waveguides is considered from the aspects of continuous and discrete uniform coupling. The case of a finite number of continuously coupled lines is solved by a normal mode transformation. This case is extended to an infinite number of lines, and a Bessel function solution is obtained. A comparison of solutions for finite and infinite sets of coupled lines indicates that the Bessel function solutions are sufficiently accurate for most practical systems. Discrete coupling is considered, and it is shown, by means of scattering matrices and normal modes, that perfect match and directivity for a multiple-line directional coupler are not theoretically possible, except in the limit of very small coupling per aperture. Experimental verification is given for a flexible arrangement of five Transvarcoupled waveguides.

### 55.3. Effects and Measurement of Harmonics in High Power Waveguide Systems

M. P. Forrer and K. Tomiyasu, *General Electric Microwave Lab., Palo Alto, Calif.*

With increasing power levels of radar systems, the adverse effects due to harmonics of the carrier frequency have become significant. Such harmonic energies can propagate through the waveguide system in many modes. Because of the ambiguous modal distribution of the propagating harmonic energies in a waveguide, the determination of the relative power levels of the propagating modes is an extremely complex problem.

For an arbitrary superposition of propagating modes in a lossless, straight, rectangular waveguide, a measuring technique has been developed to determine the power of each mode. The procedure employs electric probes, sliding across the waveguide walls at various cross sections. The mathematical formalism linking the measured probe data with the power of each mode is described and a brief description of experimental results is given.

### 55.4. Microwave Dielectric Properties of Solids for Applications at Temperatures to 3000°F

D. M. Bowie, *Melpar, Inc., Falls Church, Va.*

Measured values of dielectric constant and loss tangent of a number of plastic and ceramic materials at numerous temperatures ranging from room temperature to as high as 3000° F are reported. Tests were made at frequencies of 200 mc and 3000 mc. Refinements in the existing shorted-transmission-line measuring technique have been made in order to provide nearly uniform accuracy over a wide range of temperature.

Materials have been selected for investigation on the basis of their potential utility in high-temperature microwave applications. An extensive investigation is reported on pure materials and mixtures in the form of foams, laminates, and filled bodies, including Teflon, silicone resins, epoxy resins, glasses, and ceramics. The temperature range in every case is extended to 3000°F or to the limit of the material.

Results show generally an increase in both dielectric constant and loss with temperature, except for highly expansive materials, which show a decrease in dielectric constant with temperature. For plastics, temperature limits are indicated above which irreversible changes in dielectric properties occur. Curves of dielectric properties vs time at fixed elevated temperatures are shown for these materials.

### 55.5. Traveling-Wave Cavity for Ferrite Tensor Permeability Measurements

L. A. Ault, E. G. Spencer, and R. C. LeCraw, *The Diamond Ordnance Fuze Labs., Washington, D. C.*

A waveguide annular ring is described for use in measuring the components of the tensor permeability of ferrites. Unidirectional wave propagation within the ring, together with the resonant properties of the ring, constitute a traveling-wave cavity in a bridge circuit. The ferrite sample, which may be a small rod or thin slab, is placed in the waveguide ring at a position away from the side walls at which the rf fields are circularly polarized. As the applied dc field is varied, the loss and dispersion components of the permeability change, resulting in an unbalanced bridge. The components of permeability are measured by a calibrated phase shifter and a calibrated attenuator in the ring, which are used to rebalance the bridge.

The merits of this new method are evaluated and compared with other cavity perturbation methods for measuring tensor permeability components.

### 55.6. The Principle of a Non-Gyromagnetic Ferrite Phase-Shifter

S. Wenglin, *New York University, New York, N. Y.*

If a linearly polarized plane electromagnetic wave propagates through an infinite ferromagnetic medium, magnetized parallel to the magnetic component of the wave, the value of the permeability in the propagation constant is the incremental permeability. This incremental permeability, and thus the phase delay, may be varied by varying the magnetic bias. Magnetic properties may be controlled by utilizing a medium consisting of ferrite spheres embedded in a dielectric. Restrictions on the size and distribution of spheres are those usual for the application of the Clausius-Mosotti formula. A more general form of the Clausius-Mosotti formula is derived which relates the $B$, $H$ values in the mixture to the $B$, $H$ values in the ferrite spheres.

# IRE News and Radio Notes

## Eastham Wins Armstrong Medal

Melville Eastham (A'13–M'13–F'25) was recently awarded the Armstrong Medal by the Radio Club of America "in recognition of his outstanding contributions to the art of precision measurements in the radio and electronic field."

M. Eastham

Mr. Eastham, for fifty years a design engineer, was the founder of the General Radio Company, Cambridge, Massachusetts, and its president from 1915 to 1944.

Active in IRE committee work, Mr. Eastham was Treasurer from 1928 to 1940, Manager from 1922 to 1927 and Director from 1927 to 1941. He also served as secretary to the Boston IRE Section, and as IRE representative on other organizations. In 1937 he was awarded the IRE Medal of Honor.

Mr. Eastham also holds membership in the American Association for the Advancement of Science, American Institute of Electrical Engineers, Acoustical Society of America, American Physical Society, and American Meteorological Society.

## Chairmen Listed for Seventh Region Meeting at San Diego

The 1957 Seventh IRE Region Conference will be held at San Diego, California, April 24–26. The theme of the conference will be "Electronics in Space."

L. G. Trolese is general chairman of the conference. Assisting him will be Jack Bowers, secretary; Don Root, exhibits; R. T. Silberman, publicity; R. E. Honer, technical program; J. P. Day, finance; L. H. Orpin, entertainment; R. U. F. Hopkins, transportation; Chesney Moe, student program; D. C. Kalbfell, member at large; and San Diego section chairman Robert Kirkman and Al Drayer, secretary.

Over fourteen technical sessions are planned. Session chairmen will be: L. L. Beranek, M.I.T., *Audio-Video;* R. G. Stegen, Canoga Corp., *Microwave;* J. M. Pettit, Stanford Univ., *Wednesday Luncheon;* C. Moe, San Diego State College, *Student Papers;* J. B. Smyth, Smyth Research Associates, *Antennas and Propagation;* L. M. Silva, Beckman Instruments, *Data Handling and Automation;* D. M. Stuart, CAA, Technical Development Center, *Electronic Aids to Air Navigation;* J. E. Keister, General Electric, *Electron Devices;* R. A. Helliwell, Stanford Univ., *International Geophysical Year;* Norman Moore, Litton Industries, *Electron Tubes;* B. S. Billington, Oak Ridge National Laboratory, *Nuclear Activation and Damage;* Martin Klein, Rocketdyne, North American Aviation, *Instrumentation.*

Technical tours and ladies' activities will also be included in the conference schedule.

## IRE Adds New Subsection

The IRE Executive Committee, at its meeting on January 3, approved the formation of the Lehigh Valley Subsection of the Philadelphia IRE Section.

On the following day, the name and status of the Centre County Subsection were changed to the Central Pennsylvania Section.

## Southwestern and Simulation Conferences to Hold Simultaneous Sessions April 10–13

The Ninth Annual Southwestern IRE Conference and Show, sponsored by the Houston IRE Section, and the Second National Simulation Conference, sponsored by the IRE Professional Group on Electronic Computers, will be held at the Shamrock-Hilton Hotel, Houston, Texas, April 10–13. Five technical sessions of both conferences will be held simultaneously.

Exhibits for both conferences will open April 11, and close on the 14th. The registration desk, however, will be ready for visitors from 2 to 10 p.m. on the 10th so that arrangements for technical tours and a ladies' program can be made early.

Instrumentation and geophysics will be discussed during the technical sessions of the Southwestern Conference on April 11; computer devices and physical simulation, during the Simulation Conference sessions on that day. On the following day, technical sessions of the Southwestern Conference will be on communications and medical electronics; the Simulation Conference will discuss function generators and physical simulation. On the 13th, the Southwestern Conference will hold a session on industrial applications of electronics, and simultaneously the Simulation Conference will discuss special applications of simulation techniques.

Four events mark the social calendar of both conferences. Since the National IRE Board of Directors will hold their April meeting in Houston at this time, there will be a luncheon in their honor on the 11th, and on the evening of the following day a reception will also be held. Following the reception there will be a buffet dinner and an aquacade.

A luncheon on the 12th will feature a speech by J. A. Van Allen, professor of physics at Iowa University and U.S.—I. G. Y. Committee member, on the latest developments in the earth satellite program and its part in the International Geophysical Year.

Harvey Wheeler of KPRC-TV is chairman of the bi-conference committee. Committee members are: R. L. Ransome, Section Chairman; W. J. Greer, Vice-Chairman and Special Guests; P. E. Franklin, Business Manager and Housing; L. H. Gollwitzer, Secretary and Registration; F. C. Smith, Jr., Program; K. O. Heintz, Exhibits; J. K. Hallenburg, Tours and Banquet; R. A. Arnett, Technical Chairman; E. F. Neuenschwander, Arrangements; M. A. Arthur, Publicity; and Mrs. K. H. Woehst, Ladies.

### Calendar of Coming Events

National Biophysics Conference, Columbus, Ohio, March 4–6

EJC Second Annual Nuclear Science and Engineering Congress, Convention Hall, Philadelphia, Pa., March 11–14

**IRE National Convention, Waldorf-Astoria and New York Coliseum, New York City, March 18–21**

Colloquium on Radiation Effects, Shriver Hall, Johns Hopkins University, Baltimore, Md., March 27–29

British Radio & Electronic Component Show, Grosvenor House and Park Lane House, London, England, Apr. 8–11

**Industrial Electronics Conference, Ill., Inst. of Tech., Chicago, Ill., April 9–10**

First National Nuclear Instrumentation Conference, Atlanta, Ga., Apr. 10–12

**Ninth Southwestern Regional Conference & Show, Shamrock-Hilton Hotel, Houston, Tex., April 11–13**

**National Simulation Conference, Shamrock-Hilton Hotel, Houston, Tex., April 11–13**

**PGTRC National Telemetering Symposium, Philadelphia, Pa., April 14–16**

**Annual Engineering Convention and Exhibition of the IRE Buenos Aires Section, Buenos Aires, Argentina, April 22–28**

**Symposium on Role of Solid State Devices in Electric Circuits. Engrg. Society Bldg., New York City, April 23–25**

**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**

**Electronic Components Conference, Morrison Hotel, Chicago, Ill., May 1–3**

**Symposium on Microwave Ferrites and Devices & Applications, Western Union Auditorium, New York City, May 9–10**

**National Aero. and Nav. Electronics Conference, Dayton, Ohio, May 13–15**

Fifth Annual Semiconductor Symposium of the Electrochemical Society, Statler Hotel, New York City, May 13–16

**PGPT First Annual Conference on Production Techniques, Willard Hotel, Washington, D. C., June 6–7**

**PGMIL First National Meeting, Sheraton-Park Hotel, Washington, D. C., June 17–19**

International Symposium on Physical Problems of Color Television, Paris, France, July 2–6

**WESCON, Fairmont Hotel and Cow Palace, San Francisco, Calif., Aug. 20–23**

URSI Twelfth General Assembly, Boulder, Colo., Aug. 22–Sept. 5

**Special Technical Conference on Magnetic Amplifiers, Penn Sheraton Hotel, Pittsburgh, Pa., Sept. 4–6**

**Industrial Electronics Symposium, Morrison Hotel, Chicago, Ill., Sept. 24–25**

## Three-Day EJC Conference Revealed New Developments

The three-day 1956 Eastern Joint Computer Conference & Exhibition opened at New York City's Hotel New Yorker and Manhattan Center, December 10, 1956. This sixth annual meeting, jointly sponsored by the IRE, American Institute of Electrical Engineers and Association for Computing Machinery, focused attention on the theme "New Developments in Computers."

A record registration of several thousand electronic engineers and business executives viewed the equipment displays of more than thirty leading manufacturers and research laboratories in the computer field. The keynote address was made by H. T. Engstrom of the National Security Agency. During the conference technical papers discussing various aspects of computer systems, components, circuits, input-output devices and high-speed memories were presented by twenty-nine authorities in the field.

As a permanent record of this exchange of information, each conference registrant received a free copy of the conference proceedings, which included a complete text of all technical papers presented at the conference.

## Johns Hopkins Univ. Holds March Colloquium on Radiation

The Office of Naval Research, Navy Department, and the Glenn L. Martin Company will co-sponsor a colloquium on radiation effects on materials to be held on the campus of Johns Hopkins University at Shriver Hall, Baltimore, Md., March 27–29.

The colloquium is planned for the presentation of unclassified educational review papers on the subject of radiation effects on materials. The program has been designed to appeal to the engineer and the research scientist interested in radiation effects, and the development and application of nuclear reactors for power. Approximately one hour will be allowed for each paper's presentation and ample time will be available for discussion from the floor.

The first day's program will consist of papers on defects in solids and current concepts of radiation effects, experimental approaches to radiation studies, radiation sources and dosimetry, radiation effects on physical and metallurgical properties of metals and alloys, influence of radiation upon corrosion behavior, and surface properties of metals and alloys.

Topics to be discussed on the second day of the colloquium will be effects of radiation on electronic and optical properties of inorganic dielectric materials, effects of radiation on semiconductors, and effects of radiation on organic materials. On the last day of the colloquium, papers will be presented on the effects of radiation on reactor components. A round table, open forum discussion will then follow.

Further information may be obtained from: J. G. Morse, Supervisor, Radiation Effects Unit, Nuclear Division, Mail 711, Glenn L. Martin Co., Baltimore 3, Md., and J. J. Harwood, Head, Metallurgy Branch, Office of Naval Research, Navy Department, Washington 25, D. C.

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| | Vol. AU-4, No. 1 January–February 1956 (27 pages) | .75 | 1.10 | 2.25 |
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### WESCON PAPERS DEADLINE SET FOR MAY 1

Authors wishing to present papers at the 1957 WESCON Convention to be held in San Francisco on August 20–23 should send 100–200-word abstracts, together with complete texts or additional detailed summaries, to the Technical Program Chairman, D. A. Watkins, Stanford Electronics Laboratories, Stanford University, Stanford, California, by May 1, for consideration by the Technical Program Committee. Authors will be notified whether or not their papers have been accepted by June 1.

### USA-URSI MEETS MAY 22–25

The USA National Committee of URSI has scheduled its spring meeting for May 22–25 at Hotel Willard, Washington, D. C. Co-sponsors of the meeting will be the IRE PGAP, PGCT, PGI, PGIT, and PGMTT.

A general session will be held for all participants on the morning of May 23, followed by one or more sessions in each of these fields: Commission 1—Radio Measurements and Standards; Commission 2—Radio and Troposphere; Commission 3—Ionospheric Radio; Commission 4—Radio Noise of Terrestrial Origin; Commission 5—Radio Astronomy; Commission 6—Radio Waves and Circuits; and Commission 7—Radio Electronics.

Authors should submit titles and 100-200-word abstracts of their proffered papers on or before March 15. Papers should be sent to the appropriate person chosen from the following: Commission 1—E. Weber, Polytechnic Institute of Brooklyn, 99 Livingston St., Brooklyn, N. Y.; Commission 2—J. B. Smyth, Smyth Research Associates, 3930 Fourth Ave., San Diego 3, Calif.; Commission 3—L. A. Manning, Dept. of Elec. Eng., Stanford University, Stanford, Calif.; Commission 4—A. W. Sullivan, Engineering and Industrial Experimental Station, Univ. of Florida, Gainesville, Fla.; Commission 5—F. T. Haddock, Observatory, Univ. of Michigan, Ann Arbor, Mich.; Commission 6—E. C. Jordan, Dept. of Elec. Eng., Univ. of Illinois, Urbana, Ill.; Commission 7—W. G. Shepherd, Dept. of Elec. Eng., Phillips Hall, Cornell University, Ithaca, N. Y.; PGAP—H. G. Booker, School of Elec. Eng., Phillips Hall, Cornell University, Ithaca, N. Y.; or PGMTT—F. G. Marble, Boonton Radio Corp., Intervale Road, Boonton, N. J.

### FIRST FALL NEREM HEARS SIXTEEN TECHNICAL PAPERS

The New England Radio-Electronics Meeting, held for the first time in the fall, took place at the Hotel Bradford, Boston, Massachusetts, November 15–16. Nearly fifteen hundred persons attended the meeting, which was sponsored by the Boston and Connecticut Valley IRE Sections.

The following papers were presented at the meeting: *The Information Theory Outlook on Communication Systems of the Future,*

R. M. Fano; *Communications—A Look Ahead,* E. I. Green; *Exploitation of the Tropospheric Propagation Mode—Present and Future,* C. L. Mack, Jr.; *Company Standards Programs in the Electronics Industry,* R. G. Munroe; *The Science of Value Analysis,* L. D. Miles; *The Potentialities of Long Range Radio Aids to Navigation,* J. A. Pierce; and *Altimeters,* F. T. Wimberly.

Also *Electronic Analog Computers,* J. M. Embree; *Automation—Its Impact on Management,* C. W. Boyce; *The Problem of the Creative Engineer,* C. D. Orth, 3rd; *Present and Future Network Synthesis,* E. A. Guillemin; *Applied Circuit Theory,* W. N. Tuttle; *Nonlinear Circuitry,* Ernst Weber; *Perception and Pattern Recognition in Automata,* O. G. Selfridge; *Characteristics of Manual Control Systems,* J. I. Elkind; and *The Role of Men in Automation Systems,* B. F. Green, Jr.

E. F. Carter, Director of the Stanford Research Institute at Palo Alto, Calif., and the dinner speaker, discussed "Idea Acceleration."

## Information Available for April Television Conference

The Eleventh Annual Spring Technical Conference on Television will be held this year at the Engineering Society Building in Cincinnati, Ohio, April 26–27. This conference will be sponsored by the Cincinnati IRE Section in cooperation with the IRE Professional Groups on Broadcast and Television Receivers, and Broadcast Transmission Systems.

A. T. Bolt, Jr. and V. J Scott are co-chairmen of the conference, and J. R. Ebbeler of Avco Manufacturing Corporation is papers chairman.

Information concerning advance registration or hotel reservations can be obtained from C. B. Shaw, Jr., Hangar Three, Lunken Airport, Cincinnati 6, Ohio. Advertising and exhibition privileges can be obtained upon request from J. R. Ragen, 3825 Mantell Ave., Cincinnati 36, Ohio.

## PGEC Sponsors Fellowship

During the past two years, negotiations were completed which presently enable the PGEC to sponsor an annual Fellowship for graduate study in computing. The amount of the stipend is $2,000, plus tuition not to exceed $1,000. Initially this Fellowship is underwritten by the PGEC for two years, although it is hoped that it can be supported on a continuing basis.

This Fellowship is administered through the Office of Scientific Personnel of the National Academy of Sciences, and the selection of the successful recipient is made by that office. The first Fellowship has already been awarded for the 1957–58 academic year. There are no restrictions as to the university or college to be attended, but it is necessary for the Fellow to specialize in computing or a closely allied field.

Applications for this Fellowship should be addressed to: Fellowship Office, Office of Scientific Personnel, National Academy of Sciences, 2101 Constitution Avenue, N.W., Washington 25, D. C.

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| Ultrasonics Engineering | PGUE-1: June 1954 (62 pages) | 1.55 | 2.30 | 4.65 |
| | PGUE-4: August 1956 (75 pages) | 1.50 | 2.25 | 4.50 |
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| | PGVC-6: July 1956 (82 pages) | 1.55 | 2.30 | 4.65 |
| | PGVC-7: December 1956 (53 pages) | 1.20 | 1.80 | 3.60 |

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## PGMTT Holds Meeting May 9-10

The annual meeting of the IRE Professional Group on Microwave Theory and Techniques will take place at Western Union Auditorium, New York City, May 9–10. The program will emphasize reports on advances in microwave ferrite devices and their applications, non-reciprocal propagation in gas media, and microwave solid-state devices. Invited speakers include Benjamin Lax of M.I.T.'s Lincoln Laboratory who will discuss microwave properties and investigations of semiconductors, Louis Goldstein of the University of Illinois whose topic will be "Non-Reciprocal Electromagnetic Wave Propagation in Ionized Gaseous Media," and H. E. D. Scovil of Bell Telephone Labs who will present a new approach to a solid-state microwave amplifier in his talk "Solid State MASER."

There will be two sessions of contributed papers, abstracts for which are due March 1. Authors should send 200-word abstracts to Sam Weisbaum, Technical Program Chairman, Bell Telephone Labs., Murray Hill, N. J.

The last session of the meeting will be devoted to a round-table discussion on design limitations of microwave ferrite devices and cover high power effects, low frequency limits, high frequency limits, anomalous propagation in ferrite loaded waveguides, below saturation behavior of ferrites, "fast" ferrite devices (depending upon relaxation time), bandwidth problems, and material losses. The discussion will be led by the following panelists: H. Seidel, Bell Tel. Labs.; G. S. Heller, M.I.T. Lincoln Lab.; R. C. LeCraw, Diamond Ordnance Fuse Labs.; J. O. Artman, Harvard University; P. H. Vartanian, Sylvania; H. Carlin, Microwave Research Institute; and D. L. Fresch, Trans-Tech, Inc.

The annual award for the best paper appearing in the Transactions of PGMTT will be presented at a dinner scheduled for the evening of May 9.

## Technical Committee Notes

Chairman P. A. Redhead presided at a meeting of the **Electron Tubes** Committee held at IRE Headquarters on January 11. The committee reviewed the Proposed Standard on Testing of Storage Cathode-Ray Tubes. This is one section of the Proposed Standard on Electron Tubes: Methods of Test which the committee is now preparing.

The **Nuclear Techniques** Committee met at the National Bureau of Standards in Washington, D. C. on December 18 with Chairman G. A. Morton presiding. The entire meeting was devoted to the review of the Proposed Standard on Definitions of Terms, which is presently in preparation in the committee.

Chairman M. W. Baldwin presided at a meeting of the **Standards** Committee held at IRE Headquarters on January 10.

The Proposed Standard on Electron Tubes: Noise Definitions was discussed, amended and unanimously approved for publication as an IRE Standard.

Consideration of the revisions in the Standards on Gas-Filled Radiation-Counter Tubes: Definitions of Terms, 1952 was postponed at the request of G. A. Morton, Chairman of the Nuclear Techniques Committee. Dr. Morton wishes to have his committee review the Standard before it is approved by the Standards Committee.

Review of the Proposed Standard on Navigation Aids: Direction Finder Measurements was started, and will continue at the next meeting of the committee.

# Books

## Engineering Mathematics by K. S. Miller

Published (1956) by Rinehart & Co., 232 Madison Ave., N. Y. 16, N. Y. 408 pages+8 index pages+xiii pages. Illus. 9×6. $6.50.

There are many text books on the subject of advanced mathematics for engineers and/or physicists and more are coming out every year. A new addition to this field which should prove quite valuable to a senior undergraduate or first-year graduate curriculum on communications engineering is the subject book listed above.

This book contains many of the topics often included in similar books. These topics include determinants and matricies, special integrals such as the gamma function, beta function, error function, elliptic integrals and so on, linear differential equations including solutions in series and the well-known equations of Legendre, Hermite and Bessel, Fourier series and integrals and the La Place transformations. In addition, there is a chapter on network theory and, even more important, a very well-written chapter on random functions which is a very vital subject to modern communication theory.

Since the book is apparently restricted to use in an electrical engineering curriculum, it does not include any topics which might be considered of greater interest to the physicist such as heat flow problems, vibrating strings and plates, or hanging cables. Although it is doubtful if everything that is included in the book could be covered in a one-year course, still it is felt that there are two important topics that should have been included to make it a more complete and useful reference book to a graduate student of electrical engineering; namely, complex variable theory, especially integration in the complex plane and electromagnetic theory, and fields and waves in transmission lines and waveguides.

It is to be particularly noted that no mathematical topic is introduced merely to exhibit mathematical gymnastics; instead, new mathematical techniques and methods are introduced in conjunction with the solutions to interesting and useful physical problems. It is this approach which makes it of particular interest to this reviewer.

HAROLD STARAS
RCA Laboratories
Princeton, N. J.

## Mechanical Design for Electronics Production by J. M. Carroll

Published (1956) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 334 pages+9 index pages+1 page of bibliography+iii pages. Illus. 9×6 $6.50.

As mentioned in the preface, the volume is intended primarily for the working engineer. The book serves as an introduction, due to its comprehensive coverage, to bring awareness of the many important factors related to product design and production techniques providing much needed reliability and reduction of costs. A listing of the chapter titles may serve to indicate the subjects discussed—Chapter 1, The Importance of Mechanical Design; Chapter 2, Space Planning in Equipment Design; Chapter 3, Chassis Design and Layout; Chapter 4, Fabricating the Chassis; Chapter 5, Manufacturing Small Parts; Chapter 6, Shielding Components and Equipment; Chapter 7, Potting, Embedding and Encapsulating; Chapter 8, Moving Parts; Chapter 9, Electric Motors and Rotating Components; Chapter 10, Wiring and Soldering; Chapter 11, Assembly Methods; Chapter 12, Cabinet Construction; and Chapter 13, Environmental Factors in Equipment Design.

Chapters 1, 2, 8, 10 and 12 contain a wealth of information which should be of interest to each electronics, mechanical and production engineer. Chapter 7 on potting is a good introduction to a subject all too often unfamiliar to the product designer who lacks knowledge of plastic materials and processing techniques. It was a bit disappointing to find that the very important subject of protective coatings was not discussed more thoroughly. Chapter 11 which gives a comparison of three well known inline assembly machines, is quite useful. With the growing interest in reliability and ruggedization, Chapter 13 did not quite give sufficient emphasis to the importance of environmental testing and its influence on mechanical design of equipment. This chapter would have been considerably strengthened had it contained a discussion of component and sub-assembly mounting to withstand environmental stresses. The subject of cooling, very important when miniaturization is considered and with electronic systems becoming more and more complex, deserves more treatment. The bibliography, although good, should be expanded to provide adequate references to all the subjects discussed.

All in all, though, the book serves a useful purpose by supplying sufficient detail to be of some aid to the design engineer.

L. K. LEE
Ramo-Wooldridge Corporation
Los Angeles 45, California

## Électricité, 6th ed., by G. Goudet

Published (1956) by Masson et Cie, 120, Boulevard Saint-Germain, Paris 6, France. 900 pages+4 index pages+1 appendix page+3 pages of bibliography. 517 figures. 9¾×6¾. 4.500 F., paper; 5.100 F., cloth.

This treatise on the theory of electricity is a revised edition of a part of the well-known course of G. Bruhat dealing with general physics. Bruhat's work has long been considered in France as a classic for undergraduate and postgraduate students in French universities, as well as more generally for engineers and research men in quest of a reliable review of up-to-date fundamentals in this field. The author died during World War II, a victim of his sense of duty toward profession and country. Professor Goudet has been entrusted with the difficult task of revising this outstanding book. No better choice could have been made. Dr. Goudet has done the job with a feeling of devoted gratitude toward his former professor, as well as with utmost competence.

The book had been written with the characteristic clarity and logic often associated with the scientific teachings of outstanding French professors. However, it was in need of substantial revision because of the rapid evolution of thought and knowledge after the end of the last World War. An instance of the decisions which Professor Goudet had to make is the choice between the initial parallelism of magnetic and electric point masses as usually utilized before the war in French universities, and the derivation of magnetic properties of matter from electric currents, leaving the concept of magnetic mass as a pure mathematical tool with no physical support as is generally preferred in today's scientific circles. Prof. Goudet has introduced the latter view in the book, and proceeded very methodically from the static phenemena of both electric and magnetic nature, through quasi-stationary problems, to crown the whole electromagnetic theory with Maxwell's equations, which he has explained carefully by means of a number of applications amply demonstrating their practical usefulness as well as their profound content. A chapter follows on the Lorentz' transformation and the relativity theory.

Though the modern ideas about the way matter is constituted are explained and alluded to in the first part of the book, it is in the latter chapters that microscopic phenomena are more specifically considered. The chapter on electrons and the examination of their properties in accordance with classical, relativistic and wave mechanics is an excellent introduction to modern concepts of which the author has explained the general laws in succeeding chapters dealing with various particle constituents of atoms and molecules. The book closes with a clear, though brief review of the basic properties of solid state physics and vacuum tubes.

The Giorgi system of rationalized units has been consistently utilized throughout the book The bibliography is deliberately limited to classical publications, with preponderant references to French works.

This reviewer had the occasion to discuss a few points with the author about his revision, and is fully aware of the thoroughly competent and painstaking work which Prof. Goudet has done to bring up-to-date a treatise of such repute and quality. This book is unreservedly recommended to all radio engineers for its clarity, precision and care in establishing definitions of concepts and demonstrations of properties.

A. G. CLAVIER
Federal Telecommunications Labs.
Nutley, N. J.

## Electronic Analog Computers, 2nd ed., by G. A. Korn and T. M. Korn

Published (1956) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 425 pages+11 index pages+10 pages of bibliography+xiv pages. Illus. 9¼×6¼. $7.50.

This second edition reflects recent developments in the rapidly changing field of analog computers. For an adequate understanding of the material, a knowledge of elementary electronics, elementary feedback theory, and mathematics through elementary differential equations is required.

likely to appreciate the practice of depicting numerical values of significant circuit elements in a schematic drawing of commercial equipment.

H. S. Black
Bell Telephone Laboratories, Inc.
Murray Hill, N. J

**Reference Data for Radio Engineers, 4th ed., ed. by H. P. Westman**

Published (1956) by International Telephone & Telegraph Corp., 67 Broad St., N. Y. 4, N. Y. 1121 pages+29 index pages. Illus. $8\frac{1}{2} \times 6$. $6.00

The widespread acceptance of its earlier editions by engineers in all fields indicates the utility of this compilation of general information. As specialization increases, few engineers can avoid searching for data in their own or in other fields some of which may be but remotely connected to electronic matters. A surprising amount of material has been added to this edition, which is now about six times the size of the first edition.

The material is concisely presented, with enough explanatory data to avoid confusion as to values, terms, and units or the conversions so often encountered with the use of data from diverse sources. Since the earlier editions are so well known we will only call attention to some of the new material introduced. The compilation has been the responsibility of many engineers from among the various divisions of I. T. & T., each contributing in his particular field of specialization, which contribution is then collectively reviewed by a group of engineers having broad experience.

Some of the new or completely revised material now included are: network filter design, transistors and their circuitry, probability and statistics, information theory, magnetic amplifiers, metallic rectifiers, scattering matrixes in waveguides, scatter propagation, television pick-up tubes, nuclear physics, patent practices, and much more. Among the unusual and welcome items is a description of materials and processes for printed circuits, industry standards for many components, design practices for handling shock, vibration and other use hazards, patent coverage of inventions, and even a summary of the U. S. military nomenclature system!

This volume will probably be kept close at hand as a "try first" source of information by many engineers.

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Consultant
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## Recent Books

*Arcs in Inert Atmospheres and Vacuum*, ed. by W. E. Kuhn. Papers presented at the Symposium on Arcs in Inert Atmospheres and Vacuum of the Electrothermics and Metallurgy Division of the Electrochemical Society, April 30–May 1, 1956, San Francisco, Calif. John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $7.50.

Berkeley, E. C. and Wainwright, Lawrence, *Computers: Their Operation and Applications*. Reinhold Publishing Corp., 430 Park Ave., N. Y. 22, N. Y. $8.00.

*Equivalent Radio Tubes Vade-Mecum*, 13th ed., ed. by J. A. Gijsen. P. H. Brans, Ltd., Antwerp, Belgium.

Eustis, William, *A Primer to the Automatic Office*. Automation Management, Inc., Westboro, Mass. $7.50.

Lanczos, Cornelius, *Applied Analysis*. Prentice-Hall, Inc., 70 Fifth Ave., N. Y. 11, N. Y. $9.00.

Lang, D. G., *Marine Radar*. Pitman Publishing Corp., 2 W. 45 St., N. Y. 36, N. Y. $4.75.

Lessing, Lawrence, *Man of High Fidelity: Edwin Howard Armstrong*. J. B. Lippincott Co., East Washington Square, Philadelphia, Pa. $5.00.

Mueller, G. V., *Introduction to Electrical Engineering*, 3rd ed. McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $7.50.

*Proceedings of the CERN Symposium on High Energy Accelerators and Pion Physics* (Geneva, Switzerland, June, 1956). The two volumes are obtainable from CERN Service d'Information, Case postale 25, Geneva 15, Switzerland. 40 Swiss francs per volume.

*Proceedings of the RETMA Symposium on Reliable Applications of Electron Tubes*. Engineering Publishers, GPO Box 1151, N. Y. 1, N. Y. Paperbound. $5.00.

Thourel, L., *Les Antennes*. Dunod, 92, rue Bonaparte, Paris 6, France. 4.800 F.

# Professional Groups†

**Aeronautical & Navigational Electronics**—Joseph General, 6019 Highgate Dr., Baltimore 15, Md.

**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.

**Electronic 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, Radiation, Inc., Melbourne, Fla.

**Ultrasonics Engineering**—J. F. Herrick, Mayo Foundation, Univ. of Minnesota, Rochester, Minn.

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**San Francisco (7)**—J. S. McCullough, 1781 Willow St., San Jose 25, Calif.; E. G. Goddard, Stanford Research Institute, Menlo Park, 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.
**South Bend-Mishawaka (5)**—J. L. Colten, 149 E. Tasher, South Bend, Ind.; P. G. Cox, R. R. 2, Box 441, Osceola, Ind.
**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, General Electric Co., H.M.E.E. Dept., Bldg. 3, Industrial Park, Syracuse, N. Y.; G. M. Glasford, Elec. Eng. Dept., Syracuse University, 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.
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**Toronto (8)**—F. J. Heath, 830 Lansdowne Ave., Toronto 4, Ont., Canada; H. F. Shoemaker, Radio College of Canada, 86 Bathurst St., Toronto 2B, 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.
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**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.
**Wichita (6)**—J. W. Patterson, 300 N. Broadway, Wichita 2, Kan.; H. C. Davis, 1908 S. Parkwood Lane, Wichita 18, Kan.
**Williamsport (4)**—W. H. Bresee (Secretary), 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)**—O. La Plant, 325 N. "J." St., Oxnard, Calif.; W. L. MacDonald, Naval Air Missile Test Center, Code MT-4, Point Mugu, Calif.
**Charleston (3)**—A. Jonas, 105 Lancaster St., N. Charleston, S. C.; F. A. Smith, Route 4, Melrose Box 572, Charleston, S. C.
**East Bay (7)**—H. F. Gray, Jr., 2019 Mira Vista Dr., El Cerrito, Calif.; D. R. Cone, 6017 Chabolyn Terr., Oakland 18, Calif.
**Erie (1)**—J. D. Heibel, 310 W. Grandview, Erie, Pa.; D. H. Smith, 3025 State St., Erie, Pa.
**Gainesville (3)**—W. E. Lear, Dept. of Elec. Eng., Univ. of Fla., Gainesville, Fla. (Chairman)
**Hampton Roads (3)**—R. L. Lindell, 2731 Jason Ave., Norfolk 9, Va.; J. E. Eller, Waterview Apts., Apt. E-3, Portsmouth, Va.
**Lancaster (3)**—W. T. Dyall, 1415 Hillcrest Rd., Lancaster, Pa.; P. W. Kaseman, 405 S. School Lane, Lancaster, Pa.
**Las Cruces-White Sands Proving Grounds (6)**—Officers to be elected.
**Lehigh Valley (3)**—J. J. Karakash, Dept. of Elec. Eng., Lehigh Univ., Bethlehem, Pa.; L. G. McCracken, Dept. of Elec. Eng., Lehigh Univ., Bethlehem, Pa.
**Memphis (3)**—R. N. Clark, Box 227, Memphis State College, Memphis, Tenn. (Chairman)
**Mid-Hudson (2)**—J. C. Logue, IBM Eng. Lab. Annex, Poughkeepsie, N. Y.; R. R. Blessing, IBM South Rd., Poughkeepsie, N. Y.
**Monmouth (2)**—W. M. Sharpless, Box 107, Bell Tel. Labs., Red Bank, N. J.; A. P. King, Bell Tel. Labs., Box 107, Red Bank, N. J.
**New Hampshire (1)**—Officers to be elected.
**Orange Belt (7)**—J. Tampico, 2709 N. Garey Ave., Pomona, Calif.; R. E. Beekman, 113 N. Lillie, Fullerton, Calif.
**Palo Alto (7)**—W. B. Wholey, 25044 La Loma Dr., Los Altos, Calif.; A. M. Peterson, 14846 Manuella Ave., Los Altos, Calif.
**Panama City (3)**—Officers to be elected.
**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)**—H. H. Arnold, 548 S. Westview Dr., Winston-Salem, N. C.; C. A. Norwood, 830 Gales Ave., Winston-Salem, N. C.
**Quebec (8)**—R. E. Collin, 590 Avenue Mon Repos, Ste. Foy, Quebec, Can.; R. M. Vaillancourt, 638 Avenue 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.; F. E. La Fetra, 22700 Erwin St., Woodland Hills, Calif.
**Shreveport (6)**—E. D. Nuttall, Box 1407, Shreveport, La.; R. W. Bains, 827 Martha Lane, Shreveport, La.
**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. La Plante, Philips Laboratories, Inc., S. Broadway, Irvington, N. Y.
**Western North Carolina (3)**—J. G. Carey, 1429 Lilac Rd., Charlotte, N. C.; R. W. Ramsey, Sr., 614 Clement Ave., Charlotte, 4, N. C.

# Annual Conference on Electronics in Industry

ILLINOIS INSTITUTE OF TECHNOLOGY, CHICAGO, ILLINOIS
APRIL 9–10, 1957

**Tuesday, April 9**
**Morning**

Registration.

*Welcoming Address*, E. H. Schulz, Director of Elec. Eng., Armour Res. Found.

*Introduction*, Eugene Mittelmann, consulting engineer.

### SESSION I

*Chairman:* E. H. Jones, Northwestern Univ.

*Basic Instrumentation*, W. A. Wildhack, National Bureau of Standards.

*Communication Problems Between Instruments, Controls and Man*, H. B. Ziebolz, Askania Regulator Co.

*Economic and Technical Aspects of Industrial Electronics*, E. D. Cook, G. E. Co.

**Noon**

Lunch

**Afternoon**

### SESSION II

*Chairman:* E. A. Keller, Panellit Corp.

*Electronics in a Chemical Company*, R. C. McMillen, E. I. duPont de Nemours & Co.

*Principles and Techniques for Direct Reading Digital Transducers*, W. H. Kliever, consultant.

*Process Monitoring by Dielectric Constant*, W. H. Howe, The Foxboro Co.

*Panel Discussion.*

**Thursday, April 10**
**Morning**

### SESSION III

*Chairman:* Eugene Mittelmann, consulting engineer.

*Application of Magnetic Amplifiers in Industrial Instrumentation and Control*, W. A. Geyger, Naval Ordnance Laboratory.

*Automatic Card Programmed Control of Reversing Rolling Mills*, E. H. Browning, Westinghouse Elec. Corp.

*Solid State Devices in Industrial Electronics*, Lloyd DeVore, Stewart-Warner.

**Noon**

Lunch

**Afternoon**

### SESSION IV

*Chairman:* R. C. Johnson, Barber-Coleman Co.

*Some New Aspects of Nuclear Instrumentation in Industrial Electronics*, N. C. Anton, Anton Electronics Labs., Inc.

*Some Applications of Analog Computer Techniques to Control System Design*, Ernest Goggio, Tammen and Dennison, Inc.

*Selection of Reliability Levels in Equipment Design*, Harold Gabarino, Armour Res. Found.

*Panel Discussion.*

# Symposium on the Role of Solid State Phenomena in Electric Circuits

ENGINEERING SOCIETIES BUILDING, NEW YORK CITY
APRIL 23–25, 1957

**Tuesday morning, April 23**

Registration.

*Opening Remarks*, E. Weber, Polytechnic Institute of Brooklyn.

*The Solid State*, P. P. Ewald, Polytechnic Institute of Brooklyn.

*Physical Phenomena of the Solid State*, R. C. Fletcher, Bell Telephone Laboratories.

*Circuit Aspects of Solid State Phenomena*, E. W. Herold, Radio Corporation of America.

**Afternoon**

### BASIC PROCESSES AND TECHNIQUES

*The Versitron—A New Solid State Quantum Mechanical Amplifier*, M. W. P. Strandberg, Massachusetts Institute of Technology.

*The Solid State Maser*, J. O. Artman, Harvard University.

*Ferroelectrics*, W. P. Mason, Bell Telephone Laboratories.

*Solid State Devices Using Indium Antimonide*, T. S. Moss, Royal Aircraft Establishment, England.

*Plastic Semiconductors*, H. Mark, Polytechnic Institute of Brooklyn.

**Wednesday morning, April 24**

### SEMICONDUCTOR PROPERTIES AND TECHNIQUES

*Compound Semiconductors*, G. Fischer and W. B. Pearson, National Research Council, Canada.

*Quasi-Electric and Quasi-Magnetic Fields in Non-Uniform Semiconductors*, H. Kroemer, Radio Corporation of America.

*Transit Time Effects in Depletion Layer Transistors at Microwave Frequencies*, W. Gartner, Signal Corps Engineering Laboratories.

*The Response Time of Diffused Junction Transistors*, E. L. Steele and B. R. Gossick, Motorola, Inc.

*Electronic Commutators by Means of Minority Carrier Deflection in Semiconductors*, G. C. Sziklai and L. R. Hill, Westinghouse Electric Corporation.

**Afternoon**

### MAGNETIC PROPERTIES AND TECHNIQUES

*The Use of Superconductors for Sensitive Electrical Measurements*, I. M. Templeton, National Research Council, Canada.

*Superconductivity and Its Applications to Electric Circuits*, H. O. McMahon, Arthur D. Little, Inc.

*Domain Theory of Switching*, A. Papoulis, Polytechnic Institute of Brooklyn, and T. C. Chen, Burroughs Corp.

*Development of Ferrite Materials for Microwave Applications*, D. L. Fresh, Trans-Tech, Inc.

*Harmonic Generation in Ferrites*, E. Stern, Sperry Gyroscope Company.

**Thursday morning, April 25**

### PHOTO-TECHNIQUES

*Survey of Photoconductive Processes in High Resistivity Materials*, J. Dropkin, Polytechnic Institute of Brooklyn.

*Luminescence Effects Having Circuit Applications*, R. E. Halsted, General Electric Company.

*Photo Conductive Switching Device*, A. Bramley and J. Bramley, Allen B. DuMont Laboratories.

*Infra-Red and Microwave Modulators Using Germanium*, A. F. Gibson and J. W. Granville, Radar Research Establishment, England.

*Optical and Electrical Properties of Cadmium Sulfide and Zinc Sulfide Single Crystals and Applications*, I. Broser, Fritz Haber Institute, Max Planck Gesellschaft, Germany.

*Persistent Internal Polarization and Applications*, H. Kallmann and J. Rennert, New York University.

**Afternoon**

### ROUND TABLE DISCUSSION

*Moderator:* E. W. Herold.

*Question:* To what extent will the challenging needs of the circuit designer be met by new and unexploited developments in the solid state art?

# Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Electronic and Radio Engineer*, incorporating *Wireless Engineer*, London, England

**NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.**



**The Index to the Abstracts and References published in the PROC. IRE from February, 1955 through January, 1956, is published by the PROC. IRE, April, 1956, Part II. It is also published by *Electronic and Radio Engineer*, incorporating *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.**

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.

## ACOUSTICS AND AUDIO FREQUENCIES

**534.2+[538.566:535.4 316**
**First Correction to the Geometric-Optics Scat'ering Cross-Section from Cylinders and Spheres**—(*See* 414.)

**534.2 317**
**Propagation of [sound] Waves in a Medium with Random Inhomogeneities of the Refractive Index**—V. A. Krasil'nikov and A. M. Obukhov. (*Akust. Zh.*, vol. 2, pp. 107–112; April–June, 1956.) A review including 21 references to Russian work.

**534.2 318**
**Focusing of Sound Waves by Inhomogeneous Media**—L. M. Brekhovskikh. (*Akust. Zh.*, vol. 2, pp. 124–132; April–June, 1956.) The cases considered include 1) focusing by reflection of a spherical wave at the boundary of an inhomogeneous medium, and 2) waveguide propagation. The region of the caustic is considered in detail.

**534.2 319**
**Waveguide Propagation of Sound in Inhomogeneous Media**—Yu. L. Gazaryan. (*Akust. Zh.*, vol. 2, pp. 133–136; April–June, 1956.) The field is calculated of a point source in a horizontally stratified medium, the sound velocity in which varies with height in a specified manner. A similar velocity distribution law has been discussed previously by Epstein (*Proc. Nat. Acad. Sci., Wash.*, vol. 16, pp. 627–637; October 15. 1930, 1931 Abstracts, p. 31) for em waves.

**534.2 320**
**Correlation Properties of a Wave in a Medium with Random Inhomogeneities**—L. A. Chernov. (*Akust. Zh.*, vol. 2, pp. 211–216; April–June, 1956.) The coefficients of the longitudinal space autocorrelation of the amplitude and phase are calculated; it is shown that the longitudinal correlation spreads over a longer distance than does the transverse correlation. The coefficients of the time autocorrelation of amplitude and phase fluctuations are also determined. Some experimental results in hydroacoustics are discussed. For related earlier work, see 3233 of 1955.

**534.21 321**
**Experiments on Acoustic Relaxation Processes using Electrical Models**—K. Walther. (*Acustica* vol. 6, pp. 245–251; 1956. In German.)

**534.232 322**
**Excitation of a Quartz Radiator Simultaneously at Several Frequencies**—G. D. Mikhailov, N. V. Tikhonova. and I. M. Yadrova. (*Akust. Zh.*, vol. 2, p. 231; April–June, 1956.) A preliminary note on the excitation of vibrations in a quartz disk at two frequencies not harmonically related. The two exciting oscillators were connected respectively between the common electrode covering one face of the disk and the semicircular electrodes covering the other face.

**534.232-14-8 323**
**Dielectric Heating of Quartz Oscillating in Several Organic Liquids**—S. Parthasarathy and V. Narasimhan. (*Z. Phys.*, vol. 145, pp. 508–510; June 15, 1956. In English.) Measurements were made on crystals driven at 8.7 mc and at 2.84 mc; the results indicate that the dielectric heating is greater in liquids of lower sound-absorption coefficient.

**534.232-14-8 324**
**The Output of a Quartz Transducer Oscillating in its Fundamental and Higher Harmonics**—S. Parthasarathy and V. Narasimhan. (*Z. Phys.*, vol. 145, pp. 511–514; June 15, 1956. In English.) Measurements were made on a crystal oscillating in various liquids; the observed output is much smaller than predicted by theory.

**534.24 325**
**Use of Layers Preventing the Production of Transverse Waves in the Reflection of a Longitudinal Wave at the Boundary of a Solid Body**—M. A. Isakovich. (*Akust. Zh.*, vol. 2, pp. 150–153: April–June, 1956.)

**534.26 326**
**Scattering and Radiation of Waves by Statistically Inhomogeneous and Statistically Oscillating Surfaces**—M. A. Isakovich. (*Akust. Zh.*, vol. 2, pp. 146–149; April–June 1956.) Extension of previous work (2969 of 1953): see also 2541 of 1953 (Eckart).

**534.26 327**
**Scattering of Sound Waves by Sinusoidal and Sawtooth Surfaces**—A. N. Leporski. (*Akust. Zh.*, vol. 2, pp. 177–181; April–June, 1956.) An experimental investigation is reported. The results are compared with Lysanov's theory (see 328 below).

**534.26 328**
**An Approximate Solution of the Problem of Scattering of Sound Waves at an Irregular Surface**—Yu. P. Lysanov. (*Akust. Zh.*, vol. 2, pp. 182–187; April–June, 1956.) The calculation is based on the solution of an approximate integral equation for the normal component of the velocity potential at the scattering surface.

**534.3:534.78 329**
**Some Results of an Analysis of a Singer's Voice**—S. N. Rzhevkin. (*Akust. Zh.*, vol. 2, pp. 205–210; April–June, 1956.) The frequency characteristics of a trained bass voice and an untrained baritone voice are presented graphically and discussed.

**534.6-8 330**
**Differential Method of Measuring the Absorption of Ultrasonic Waves in Liquids**—I. G. Mikhailov and G. N. Feofanov. (*Akust. Zh.*, vol. 2, pp. 194–198; April–June, 1956.)

**534.61 331**
**Absolute Method for Sound Intensity Measurement**—D. R. Pardue and A. L. Hedrich. (*Rev. Sci. Inst.*, vol. 27, pp. 631–632; August, 1956.) The method described is based on measurement of the temperature variations associated with the sound wave, using a thermometer with a very fast response.

**534.7 332**
**The Unity of Speech and Hearing**—H. Mol. (*PTT-Bedrijf*, vol. 7, pp. 69–75; July, 1956.) The interdependence of the two functions in communication by language and the shortcomings of purely objective measurements are stressed.

**534.845 333**
**The Absorption of Sound in Air at Audio Frequencies**—E. J. Evans and E. N. Bazley. (*Acustica*, vol. 6, pp. 238–245; 1956.) Measurements have been made in a large reverberation chamber with very small surface absorption, at frequencies from 1 kc to 12.5 kc, for values of relative humidity from about 5 per cent to 85 per cent and a temperature of 20°C. The values obtained for the absorption due to the air are in close agreement with values given by relaxation theory. A general expression is

derived for the attenuation of sound in air as a function of frequency, humidity, and temperature.

621.395.61.089.6:534.612 334

**The Thermophone, an Aerodynamic Pistonphone**—P. Riéty. (*Acustica*, vol. 6, pp. 251–258; 1956. In French.) Rigorous theory presented previously (657 of 1956) is supplemented by a discussion based on the physical processes, *i.e.*, energy transformations, in the thermophone. The operation of the device is compared with that of the pistonphone. A formula derived for the acoustic pressure inside the thermophone is identical with that obtained previously. The device is of interest for calibrating microphones, though not so accurate as the reciprocity method.

## ANTENNAS AND TRANSMISSION LINES

621.372 335

**The Launching of a Plane Surface Wave**—G. J. Rich. (*Proc. IEE*, Part B, vol. 103, pp. 787–788; November, 1956.) Discussion on 1548 of 1955.

621.372.2 336

**Theory of Helical Line Surrounded by a Cylindrical Semiconducting Envelope**—E. G. Solov'ev and L. V. Belous. (*Radiotekhnika, Moscow*, vol. 11, pp. 31–35; April, 1956.) The propagation of em waves along a helix inside a quartz tube coated with a semiconducting layer is discussed. An equation is derived indicating the dependence of the attenuation on the thickness $d$ of the semiconducting layer, the conductivity $\sigma$, retardation and frequency. Curves show the calculated and observed dependence of the attenuation on $\sigma \cdot d$.

621.372.2:621.372.8 337

**Surface Electromagnetic Waves on a Comb Structure**—L. A. Vainshtein. (*Zh. Tekh. Fiz.*, vol. 26, pp. 385–397; February, 1956.) Theory is developed based on the functional equations used in the rigorous theory of radiation from the open end of a waveguide. An exact characteristic equation of the surface waves is derived, as well as a simplified approximate form; a method is proposed for estimating the accuracy of the latter equation. Curves show the dispersion properties of periodic comb structures

621.372.2+621.372.8]:621.385.029.6 338

**Electron Waves in Retarding Systems**—L. A. Vainshtein. (*Zh. Tekh. Fiz.*, vol. 26, pp. 126–140, 141–148; January, 1956.) The linear theory of electron waves in retarding systems is discussed in relation to the excitation of waveguides. The characteristic wave equation is derived and its solutions are studied. The physical meaning of the quantities appearing in the equation is established. The following particular retarding systems are considered: 1) waveguide filled with a dielectric, 2) transmission line of comb type and 3) helical line. The results obtained are compared with those published in the literature.

621.372.21 339

**Free Oscillations in Simple Distributed Circuits**—A. B. Hillan. (*Wireless Eng.*, vol. 33, pp. 279–290; December, 1956.) Analysis is presented for a representative distributed circuit in the form of a finite length of uniform transmission line, 1) with a pure inductance at one end and an open-circuit at the other, or 2) with a pure capacitance at one end and a short-circuit at the other. Distinction is drawn between the cases where the energy is initially stored in the lumped-circuit element and where it is stored in the line. The method is useful for determining the effect of measuring circuits and cables on line waveforms.

621.372.8 340

**A Low-Attenuation Waveguide Free From Phase and Attenuation Distortion**—H. G. Unger. (*Arch. Elekt. Übertragung*, vol. 10, pp. 253–260; June, 1956.) For the $H_{01}$ mode in a circular metal waveguide with a suitably dimensioned coaxial inner tube of low-loss dielectric, the attenuation/frequency curve exhibits a minimum and the group-velocity /frequency curve exhibits an extreme value at the same frequency, at which propagation is hence free from distortion. For wide-band applications, the diameter and wall thickness of the dielectric tube must be small.

621.372.8 341

**Influence of a Semiconducting Film on the Attenuation of Radio Waves in Waveguides with Circular Cross-Section**—V. V. Malin. (*Radiotekhnika i Elektronika*, vol. 1, pp. 34–37; January, 1956.) The attenuation of $H_{0m}$, $E_{0m}$ and $E_{nm}$ waves in a cylindrical waveguide whose inner surface is coated with a material having a complex permittivity is calculated. The formulas derived are valid for $\epsilon'' \ll \epsilon'$ and $\epsilon'' \gg \epsilon'$ as well as for $\epsilon'' \approx \epsilon'$. Graphs show that the maximum attenuation for all modes except the $H_{0m}$ occurs at $\epsilon'' \simeq \epsilon'$; it is assumed that the film is thick compared with the surface film of the metal.

621.372.8 342

**Scattering by a Semi-infinite Resistive Strip of Dominant-Mode Propagation in an Infinite Rectangular Waveguide**—V. M. Papadopoules. (*Proc. Camb. Phil. Soc.*, vol. 52, pp. 553–563; July, 1956.) A calculation is made, using Laplace transforms, of the scattering produced by a strip arranged parallel to the electric field in the mid-plane of the waveguide. Formulas are derived for the amplitude of the scattered waves, and numerical results are given for various values of the surface resistivity of the strip.

621.372.8:621.318.134:538.221 343

**Relations between the Structure of Ferrites and Conditions for their Resonance in Waveguides. Unidirectional Guides**—Suchet. (See 512.)

621.396.67+[621.372.8:538.63]+621.396.11 +[538.566:535.42/.43 344

**Symposium on Electromagnetic Wave Theory**—(See 415.)

621.396.674.3:621.396.93 345

**The Use of a Horizontal Dipole as a Direction-Finding Antenna**—G. Millington. (*Marconi Rev.*, vol. 19, pp. 97–118; 3rd Quarter 1956.) The problem is considered in relation to the type of wave radiated by the transmitter, neglecting site, and instrumental errors. Conditions are analyzed for both small and large angles of elevation at the receiver: examples show that errors may be very large in the latter case. A comparison is made between the vertical-frame and horizontal-dipole antennas for u.s.w. d.f.; the nature and order of bearing errors are similar for the two types.

621.396.677 346

**Ground Antennas**—J. Grosskopf. (*Nachrichtentech. Z.*, vol. 9, pp. 241–244; June, 1956.) Simple analysis indicates that the performance of ground antennas can be predicted both qualitatively and quantitatively from accepted theory: freedom from noise is no greater than for other types of antenna. Measurements on several different systems support this view.

621.396.677 347

**Investigations of Ground Antennas (Long-Wave Directional Receiving Installations)**—W. Kronjäger and K. Vogt. (*Nachrichtentech. Z.*, vol. 9, pp. 245–249; June, 1956.) Symmetrical antennas were used in the measurements reported, on account of their relative insensitivity to local noise. The results indicate that these antennas have similar receiving properties to a frame antenna with vertical axis of rotation. At a communication station with sufficient ground space for a crossed ground antenna, such an antenna is on account of its simplicity, preferable to an equivalent crossed-frame system.

621.396.677 348

**An Experimental Design Study of some S- and X-Band Helical Antenna Systems**—G. C. Jones. (*Proc. IEE*, Part B, vol. 103, pp. 764–771; November, 1956.) The general characteristics of helical antennas are summarized, and measurements of radiation patterns are reported for single and multiple types. Methods such as end loading and tapering are discussed for increasing beam width and bandwidth. Satisfactory techniques have been developed for sealing the helices inside foamed insulating materials to improve their rigidity. Wide-band antennas giving linearly polarized radiation can be produced using rectangular helices.

621.396.677:523.16 349

**Radio Astronomy and the Jodrell Bank Telescope**—Lovell. (See 421.)

621.396.677.3 350

**Note on the Fourier Coefficients for Tchebycheff Patterns**—H. E. Salzer. (*Proc. IEE*, Part C, vol. 103, pp. 286–288; September, 1956.) A formula derived for the feeding coefficients for optimum beam patterns for antennas is equivalent to a set of formulas given by Duhamel (2225 of 1953) but is more convenient for computation.

621.396.677.3 351

**The Determination of the True Side-Lobe Level of Long Broadside Arrays from Radiation-Pattern Measurements made in the Fresnel Region**—R. H. T. Bates and J. Elliott (*Proc. IEE*, Part C, vol. 103, pp. 307–312 September, 1956.)

621.396.677.3 352

**Influence of a Parasitic Antenna in a Rectangular Array**—G. Boudouris. (*J. Brit. IRE* vol. 16, pp. 571–584; October, 1956.) "A rectangular antenna array has a parasitic element placed at its center. All the antennas making up the array are symmetrical half-wave dipoles parallel to one another and not displaced in the sense of their axes, and the system radiates in free space. The antennas are assumed to be thin and in the form of wires. Radiation diagrams and the gain of the network are considered from the double standpoint of the influence of the parasitic element and of the geometric configuration of the rectangle. The formulas produced are developed in the form of graphs. Some comments relative to the case of earthed dipoles are given." See also 554 of 1950.

621.396.677.3:621.372.6 353

**N-Terminal Networks**—Bloch. (See 383.)

621.396.677.31 354

**Optimal Linear Arrays of Antennas, Radiating Perpendicular to the Axis**—V. L. Pokrovski. (*C.R. Acad. Sci. U.R.S.S.*, vol. 109 pp. 769–770; August 1, 1956. In Russian.) The problem considered previously by Dolph (2487 of 1946) is solved for arbitrary radiator spacing. The calculation is carried out for arrays with symmetry about a mid-point, but the calculated optimum current distribution applies to all linear arrays

621.396.677.71 355

**Radiation Patterns of Circumferential Slots on Moderately Large Conducting Cylinders**—J. R. Wait and J. Kates. (*Proc. IEE* Part C, vol. 103, pp. 289–296; September, 1956.) "Computed patterns are presented for thin half-wave circumferential slots on circular conducting cylinders of infinite length. The cases considered are for single and diametrically opposed slots on cylinders whose circumferences vary from 3 to 21 wavelengths."

621.396.677.8 356

**Wide-Angle Scanning Performance of Mirror Antennas**—J. F. Ramsay and J. A. C. Jackson. (*Marconi Rev.*, vol. 19, pp. 119–140; 3rd Quarter, 1956.) Design details are discussed of a coma-corrected zoned paraboloidal mirror and a spherical mirror antenna, the latter based on the work of Ashmead and Pippard (1342 of 1947). Results of performance tests on both types using offset feed agree closely. A comparison of mirror and lens antennas shows that mirrors are preferable for line scanning, a lens being more advantageous where a large volume of scan is required. Scanning charts show possible "pin-cushion" and "barrel" distortion; a generalized, approximate chart has been developed for indicating the scanning performance of both mirrors and lenses.

621.396.677.83 357

**The Insertion Equivalent of Passive Reflectors [in microwave links]**—C. Micheletta. (*Alta Frequenza*, vol. 25, pp. 275–304; June /August, 1956.) An expression is derived which relates the field strength received at an antenna via a deviating reflector to that received at the same antenna without the reflector, over a path of the same total length. By applying the formula to reflectors of various shapes the validity of the usual approximations is assessed. Numerical examples are given for paths with one or two reflectors. The minimum distance to which the formula is applicable is found by comparison with the method of Jakes (1243 of 1953).

621.396.677.85 358

**A Study of the Field Distribution at an Axial Focus of a Square Microwave Lens**—P. A. Matthews and A. L. Cullen. (*Proc. IEE*, Part C, vol. 103, pp. 449–456; September, 1956.) Discussion indicates that the principal component of the electric field vector can be evaluated with sufficient accuracy from the scalar theory. Measurements of the transverse component of the electric field made by a perturbation method using a spinning dipole, give results in good agreement with theory. The 180° phase shift associated with passage of a wave through the focal plane is considered; the related change in wavelength near the focus is verified experimentally.

## AUTOMATIC COMPUTERS

681.142 359

**Transistor/Magnetic Analogue Multiplier**—G. L. Keister. (*Electronics*, vol. 29 pp. 160–163; October, 1956.) A four-quadrant voltage multiplier using magnetic cores and junction-transistor switches is described, giving an output linear within ±3 per cent. Multiplication is performed by averaging a square wave of which the amplitude is proportional to one input voltage and the phase to the other.

681.142:621.318.134 360

**Design of the Components of a Fast-Acting Store using Ferrite Ring Cores**—H. Gillert. (*Nachrichtentech. Z.*, vol. 9, pp. 250–252; June, 1956.) The requirements for transmission between line selector and lines and between column selector and columns in matrix storage devices are investigated; design data for the selector-core windings are hence derived. Some details are given of the storage unit in the Darmstadt computer DERA.

681.142:621.383 361

**Photoelectric Analogue Function Generator**—R. A. Sinker. (*Electronics*, vol. 29, pp. 178–181; October, 1956.) Three-dimensional data stored as density variations on photographic film are read by means of a cr beam and photomultiplier. Satisfactory accuracy for function generation is obtained by providing a grey-scale standard deflection feedback loop and a servo feedback loop controlling light intensity.

## CIRCUITS AND CIRCUIT ELEMENTS

621.318.424:621.3.011.3 362

**On the Inductance of Iron-Cored Coils**—P. Hammond. (*Proc. IEE*, Part C, vol. 103, pp. 249–259; September, 1956.) "The magnetic field in the neighborhood of coils threaded through holes in iron cylinders is calculated. The results are relevant to the estimation of the self-inductance of rotating machines and transformers and form the basis for an examination of the soundness of approximate formulas used in the design of such apparatus."

621.318.57 363

**Precision Electronic Switching with Feedback Amplifiers**—C. M. Edwards. (PROC. IRE, vol. 44, pp. 1613–1620; November, 1956.) Circuits are described in which precise control of voltage or current levels is achieved by using a high-gain feedback amplifier to minimize differences and nonlinearities in the circuit elements. Applications in analogue computers, signal-comparing devices and communication systems are indicated.

621.319.45 364

**Tantalum Capacitors use Solid Electrolyte**—D. A. McLean. (*Electronics*, vol. 29, pp. 176–177; October, 1956.)

621.372 365

**Nyquist's Stability Criterion**—E. A. Freeman and J. F. Meredith. (*Wireless Eng.*, vol. 33, pp. 290–294; December, 1956.) A proof based on Laplace-transform calculus is presented; the analysis is applicable to multiloop as well as single-loop feedback systems.

621.372:534.213-8 366

**On the Measurement of Attenuation in Ultrasonic Delay Lines**—M. Redwood and J. Lamb. (*Proc. IEE*, Part B, vol. 103, pp. 773–780; November, 1956.) "A theoretical and experimental study has been made of the effects of coupling films on the propagation of compressional waves from a transducer to a solid medium. In practice it has been found that "wringing" the transducer to the specimen with oil as a coupling medium produces a film of nonuniform thickness. Although the variations in thickness are of the order of a wavelength of light, these variations are important at ultrasonic frequencies in the region of 50 mc and above. Conditions are described under which such films can lead to the propagation of a predominantly first-order mode in the specimen, resulting in an exponential decay of the amplitudes of successive reflections, with a consequent improved accuracy of attenuation measurement."

621.372.412:549.514.51 367

**Frequency-Temperature-Angle Characteristics of AT-Type Resonators made of Natural and Synthetic Quartz**—R. Bechmann. (PROC. IRE, vol. 44, pp. 1600–1607; November, 1956.) Differences between the frequency/temperature characteristics and optimum orientation angles of natural and synthetic quartz resonators are discussed on the basis of detailed experimental and analytical studies. Synthetic quartz with deliberately modified properties can be produced by introducing other substances during the growing process. Measurements are reported on AT-type resonators cut from quartz grown in an alkaline solution containing germanium oxide; the third-order temperature coefficient is noticeably reduced.

621.372.412:621.372.54 368

**The Use of Ethylene Diamine Tartrate for Piezoelectric Filter Elements**—J. Birch, A. G. Frith, A. C. L. Ferguson, R. H. A. Miles and J. F. Werner. (*Proc. IEE*, Part C, vol. 103, pp. 420–427; September, 1956.) "Crystal resonators suitable for use in telephone system channel filters have been made from ethylene diamine tartrate (edt). Methods used in the fabrication and mounting of these units are described. The circuits used in the determination of the electrical parameters are given, and the results of a large number of measurements are outlined. Freedom from unwanted resonances over the desired frequency bands combined with the required characteristics has been achieved by suitably dimensioning the plates, and by applying electrodes to a part of the crystal surface only. Increased activity for crystals mounted in air can be obtained by the use of suitably placed reflectors."

621.372.45 369

**A Negative Resistance for Direct and Alternating Current**—L. Waldmann and R. Bieri. (*Z. Naturf.*, vol. 10a, pp. 814–820; November, 1955.) The negative resistance is constituted by a symmetrical two-terminal tube circuit comprising in its simplest form a twin triode with separate anode resistances. By connecting further twin triodes in parallel and providing cathode resistances, a resistance of about −3 kΩ is obtained, constant to within 1.5 per cent at voltages up to ±12 v. The stability limits of the arrangement are discussed. Applications in the field of measurements and analogue technique are indicated.

621.372.5 370

**A Method of Analysing the Performance of Tandem-Connected Four-Terminal Networks**—P. W. Seymour. (*Proc. IRE, Aust.*, vol. 17, pp. 249–255; July, 1956, *J. Brit. Instn Radio Engrs.*, vol. 16, pp. 555–562; October, 1956.) A graphical technique is presented.

621.372.5 371

**Some Properties of the Transfer Function of Unbalanced RC Networks**—I. Cederbaum. (*Proc. IEE*, Part C. vol. 103, pp. 400–406: September, 1956.)

621.372.5 372

**A Transformation Diagram for Lossy Quadripoles**—F. Gemmel. (*Arch. Elekt. Übertragung*, vol. 10, pp. 265–267; June, 1956.) A graphical method is described which, using measured input and termination data, represents separately the transformation of the lossy and loss-free components. The input impedance corresponding to any required terminating impedance can thus be determined.

621.372.5:621.374.5:621-52 373

**The Continuous Delay-Line Synthesizer as a System Analogue**—J. H. Westcott. (*Proc. IEE*, Part C, vol. 103, pp. 357–366; September, 1956.) The use of a tapped continuous delay line for simulating transfer functions of linear systems is described. The accuracy of the device is estimated with particular reference to servo systems.

621.372.5:621.374.5:621-52 374

**Properties of a Feedback-System Analogue based on a Discontinuous-Delay-Line Synthesizer**—R. M. F. Houtappel. (*Proc. IEE*, Part C, vol. 103, pp. 367–377; September, 1956.) A simple method is presented for determining the number of sections required for a discontinuous delay line to simulate with a given degree of accuracy the transfer function of a network considered as part of a feedback system.

621.372.5:621.375.2 375

**Determination of Optimum Linear-Network Parameters based on Time Characteristics**—R. Kulikowski. (*Archiwum Elektrotech.*, vol. 4, pp. 323–346. English summary, pp. 345–346; 1955.) The method of analysis used is based on minimization of the deviation between the input and output functions considered in a particular function space. Several amplifier circuits are discussed in detail, with emphasis on pulse response.

621.372.5.029.3.018.783 376
**Compensation Method of Reducing Nonlinear Distortion**—A. A. Gorbachev. (*Radiotekhnika, Moscow*, vol. 11, pp. 67–74; April, 1956.) The principles of a method of compensating distortion by means of distortion voltages or currents fed in phase opposition are considered theoretically. A block diagram of the arrangement is given. Results of an experimental verification in an af power amplifier are tabulated and indicate, in this particular case, a decrease of the coefficient of nonlinearity at the load from 13 per cent to 1.4 per cent at 400 cps, with smaller improvements at higher and lower frequencies.

621.372.54 377
**Filter Design by Synthesis**—G. De Lotto and M. Trinchieri. (*Alta Frequenza*, vol. 25, pp. 233–274; June/August, 1956.) The application of Darlington's method (*J. Math. Phys.*, vol. 18, pp. 257–353; September, 1939) is detailed. By modifying the attenuation function used in the calculations, compensation for losses in the circuit, and hence a closer approach to the ideal response curve, can be achieved. A numerical example and useful tables are included.

621.372.54 378
**Comparison between the Image-Parameter Method and the Modern Insertion-Loss Method of Filter Design**—V. Fetzer. (*Arch. Elekt. Übertragung*, vol. 10, pp. 225–240; June, 1956.) The two design methods are outlined; discussion shows that the insertion-loss method is superior, as it permits closer conformity to the required tolerances and a greater freedom in the selection of parameters. The larger amount of calculation necessary can, in special cases, be reduced by reference to a filter catalogue [351 of 1956 (Glowatzki)]. Further applications of the modern method are mentioned. 76 references.

621.372.54:621.315.212 379
**V.H.F. Band-II Harmonic Filter**—B. M. Sosin. (*Marconi Rev.*, vol. 19, pp. 89–96; 3rd Quarter, 1956.) The design and characteristics of a filter for use in transmitters are described. The filter is of the varying-impedance type [see *e.g.*, 2338 of 1954 (Yin and Foley)] and forms part of a $3\frac{1}{4}$-in-diameter 51.5-Ω coaxial feeder run.

621.372.57.029.6 380
**The Equivalent Noise Quadripole Treated as a Wave Quadripole**—H. Bauer and H. Rothe. (*Arch. Elekt. Übertragung*, vol. 10, pp. 241–252; June, 1956.) Sources of noise voltage and current are interpreted as waves, to make the equivalent network [2665 of 1956 (Rothe and Dahlke)] applicable to uhf conditions. The relation between matrices based on wave parameters and impedance concepts is derived.

621.372.6 381
**New Hybrid Devices for Combining and Distributing Electric Power at High Frequencies**—A. A. L'vovich. (*Radiotekhnika, Moscow*, vol. 11, pp. 36–43; April, 1956.)

621.372.6 382
**Analysis of Linear $n$-Port Networks**—I. Cederbaum. (*Proc. IEE*, Part C, vol. 103, pp. 267–271; September, 1956.) Analysis is presented for the general case of the nonreciprocal $n$-port linear network; using a symbolic notation, a unique expression for all network functions is derived. Examples of the application of the method are given.

621.372.6:621.396.677.3 383
**$N$-Terminal Networks**—A. Bloch. (*Wireless Eng.*, vol. 33, pp. 295–300; December, 1956.) Generalized two-terminal-network theorems are used to analyze the directive properties of antenna arrays.

621.373:621.316.729 384
**Synchronization of Oscillators by Sloping [-edge] Radio [-frequency] Pulses**—E. S. Voronin and R. V. Khokhlov. (*Radiotekhnika i Elektronika*, vol. 1, pp. 79–87; January, 1956.) A theoretical analysis is presented.

621.373.2.029.6 385
**Researches into Spark Generation of Microwaves**—M. H. N. Potok. (*Proc. IEE*, Part B, vol. 103, pp. 781–787; November, 1956.) The wide frequency band generated by spark generators permits desired ranges to be selected by means of filters. Waveguide filters of post and iris type are suitable for this purpose. Measurements of wavelength and bandwidth are reported; a Boltzmann interferometer was used, with a cro and recording camera.

621.373.4:621.396.61 386
**An Analysis of Pulse-Synchronized Oscillators**—G. Salmet. (PROC. IRE, vol. 44, pp. 1582–1594; November, 1956.) Problems in the design of high-precision variable-frequency oscillators, as required for communication transmitters, are discussed. The principles of operation of the impulse-governed oscillator [see *e.g.*, 767 of 1951 (Hugenholtz)] are explained. Use of a modified form of the circuit, termed a "phase follower," eliminates the tendency to instability as well as other undesirable characteristics of the IGO.

621.373.421 387
**Theory of the Triode Valve Oscillator with Feedback**—V. A. Zore. (*Zh. Tekh. Fiz.*, vol. 26, pp. 181–192; January, 1956.) Continuation of a previous paper (2480 of 1949). A general design method is given taking into account the effect of grid current and of electron transit time. The condition of self-excitation and the dependence of the frequency generated on the anode voltage are investigated. A formula is derived for calculating the limiting wavelengths.

621.373.421:621.314.7 388
**A Transistor RC Oscillator**—G. Francini. (*Alta Frequenza*, vol. 25, pp. 198–210; June /August, 1956.) The substitution of a transistor for a triode is investigated in the RC oscillator with twin-T feedback network described *e.g.* in 3198 of 1955 (Tucker). A single transistor without matching transformer is sufficient because of the low attenuation and high selectivity of the network even at very low frequencies where its impedance rises to about 1MΩ.

621.373.421.14:621.372.413 389
**Frequency Control in the 300–1200-Mc/s Region**—D. W. Fraser and E. G. Holmes. (PROC. IRE, vol. 44, pp. 1531–1541; November, 1956.) Oscillators controlled by coaxial-cavity resonators are discussed. Frequency stability can be increased ten- to twenty-fold by connecting a small capacitor in series with the resonator. Using a Type-6AF4 tube with an anode voltage of 50–60V, the variation of frequency with voltage is 0.3 cps per mc per volt in the 600-mc region: this compares favorably with overtone-crystal oscillators in their upper frequency range. For frequencies between 700 mc and 1 kmc a control system comprising two cavities arranged end to end is found satisfactory with a pencil triode Type-5876; the series capacitor is constituted by an iris between the two cavities. A resonator tunable over a range of ±15 mc around 600 mc is described. Variations with temperature are minimized by the design.

621.373.423:621.316.729 390
**Mutual Synchronization of Reflex Klystrons**—R. V. Khokhlov. (*Radiotekhnika i Elektronika*, vol. 1, pp. 88–97; January, 1956.) The possibility of a continuous transition from control by one klystron to control by the other is shown by an analysis of the solutions of van der Pol's equations describing the synchronization process.

621.375.2 391
**Negative-Impedance Amplifiers**—W. Nowicki. (*Archiwum Elektrotech.*, vol. 4, pp. 279–322. French summary, pp. 318–322, 1955.) An analysis is made of the performance of amplifiers in which negative impedance is introduced by positive feedback. Frequency distion is greater than for other types of line amplifier; stability also is inferior. Such amplifiers are nevertheless useful in certain circumstances since they are very economical of copper.

621.375.221 392
**Wide-Band Multistage Amplifiers with Gaussian Characteristics**—W. Golde. (*Archiwum Elektrotech*, vol. 4, pp. 215–246. English summary, pp. 245–246; 1955.) Analysis is based on use of the Taylor approximation to the Gaussian function. Typical low-pass and band-pass circuits are examined. The response of such amplifiers to step signals is nearly monotonic overshoots being ≯1 per cent for the number of stages commonly used. A comparison of asynchronous and synchronous types indicates that the former give better pulse response.

621.375.227 393
**A Differential Amplifier of high Input Impedance for Suppression of [unbalanced] Earth Voltages**—A. P. Bolle and D. Capel. (*PTT-Bedrijf*, vol. 7, pp. 58–63; July, 1956.) The circuit described covers a frequency band from 10 kc to 1 or 2 mc. Consideration is given to the rejection factor [see also 362 of 1956 (Klein)], which varies with frequency and according to the adjustment of the balancing net work.

621.375.23 394
**Nonlinear-Amplifier Design for Pulse-Height Analysers**—G. W. Hutchinson. (*Rev. Sci. Instr.*, vol. 27, pp. 592–596; August, 1956.) "The ratio of maximum to minimum pulse sizes which can be simultaneously recorded by a kick sorter is limited by the curvature of the characteristic of its biased amplifier over a finite region of input potentials. The extent of this region can be reduced from a few volts to about 10 mV by including the nonlinear elements of the amplifier in its negative feedback loop. An amplifier is described which embodies this principle and also provides a discriminator and gate circuit."

621.375.4:621.314.7 395
**Gain Chart for Transistor Amplifiers**—G. H. Myers. (*Electronics*, vol. 29, pp. 224, 226; October, 1956.)

621.375.4:621.314.7 396
**Common-Emitter Transistor Video Amplifiers**—G. Bruun. (PROC. IRE, vol. 44, pp. 1561–1572; November, 1956.) Design theory applicable to alloy-junction transistors makes use of the hybrid-Π equivalent network. The bilateral nature of the transistor is taken into account by adding a "Miller" capacitance term to the diffusion capacitance. Gain-bandwidth products and optimum load resistances are determined for cascaded stages. The production of maximally flat frequency response in a single stage by means of a capacitor shunting the feedback resistor or by means of inductances in the interstage couplings is discussed.

621.375.5 397
**Resonant Dielectric Amplifier**—M. Vadnjal. (*Alta Frequenza*, vol. 25, pp. 211–232; June /August, 1956.) The solution of a second-order

differential equation yields an expression for the amplification in terms of all the circuit parameters. Hence the optimum voltage and power amplification, the distortion, and the frequency response are easily determined [see also 2903 of 1954 (Penney *et al.*)]. Experimental results confirm the validity of this analysis.

## GENERAL PHYSICS

537.312.62 398

**Magnetic Energy and Electron Inertia in a Superconducting Sphere**—E. G. Cullwick. (*Proc. IEE*, Part C. vol. 103, pp. 441–446; September, 1956.)

537.5 399

**Polarization of the Continuous Spectrum in a Gas Discharge**—B. V. Paranjape. (*Proc. Phys. Soc.*, vol. 69, pp. 765–768; July 1, 1956.) Arguments based on the velocity distribution of the electrons are advanced to show that the continuous part of the spectrum of a gas discharge may be expected to be polarized.

537.5 400

**Electronic Motion in Gases and the Method of Free Paths**—L. G. H. Huxley. (*Proc. Phys. Soc.*, vol. 69, pp. 769–771; July 1, 1956.) Discussion on statements made in a paper by Jancel and Kahan (*The Physics of the Ionosphere*, conference report, 1955, p. 365).

537.533 401

**Electron Emission from Zinc Crystals subjected to Plastic Deformation**—J. Lohff. (*Z. Phys.*, vol. 145, pp. 504–507; June 15, 1956.)

537.533:537.534.8:537.525.8 402

**Auger Ejection of Electrons from Barium Oxide by Inert Gas Ions and the Cathode Fall in Normal Glow Discharges**—Y. Takeishi. (*J. Phys. Soc. Japan*, vol. 11, pp. 676–689; June, 1956.) Calculated total yields and energy distribution for He, Ne, Ar, Kr, and Xe positive ions are high compared with those observed by Hagstrum (681 of 1955) for emission from tungsten; the values for $Ar^+$ agree with those observed by Varney (2357 of 1954). Cathodic phenomena observed were in agreement with theory.

537.533:537.58 403

**Thermionic Emission Constants of Iridium**—D. L. Goldwater and W. E. Danforth. (*Phys. Rev.*, vol. 103, pp. 871–872; August 15, 1956.) Measurements on ribbon specimens subjected to prolonged cleaning are reported; the values found for the constants $A$ and $\phi$ are respectively 170 A/cm² per (deg C)² and 5.40 v. The spectral emissivity is 0.33 at $\lambda = 0.65\mu$.

537.533:621.385.032.21 404

**Field Emission and Field-Emission Cathodes**—Zernov and Elinson. (See 637.)

537.534 405

**Mass-Spectrometer Investigations of the Field Emission of Positive Ions**—M. C. Inghram and R. Gomer. (*Z. Naturf.*, vol. 10a, pp. 863–872; November, 1955.) Preliminary experiments indicate that the combination of the field-emission source and the mass spectrometer is likely to be useful for investigating various phenomena such as adsorption processes etc.

537.56 406

**Theory of Plasma Resonance**—P. A. Wolff. (*Phys. Rev.*, vol. 103, pp. 845–850; August 15, 1956.) "Starting from the Boltzmann transport equation, a formula is derived for the rate of change of electron density in a gas plasma. With its aid a study is made of the oscillations of electron density (about the steady state value) in a discharge confined between parallel plates. The oscillations are described in terms of a set of normal modes characteristic of the plasma under study. An expression is obtained for the impedance of the discharge as a function of frequency: for the one case calculated in detail, this formula gives a resonance in power absorption at a frequency of 0.7 of the plasma resonance frequency corresponding to the central electron density."

537.56 407

**Helmholtz Instability of a Plasma**—T. G. Northrop. (*Phys. Rev.*, vol. 103, pp. 1150–1154; September 1, 1956.)

537.56 408

**The Extraction of Ions from Plasmas and Plasma-Like System**—W. Fischer and W. Walcher. (*Z. Naturf.*, vol. 10a, pp. 857–863; November, 1955.) Report of an experimental and theoretical investigation of the extraction mechanism; conclusions are drawn regarding the limits of applicability of probe theory and of the term "plasma."

537.56 409

**Recording Decay of Electron Density in Ionized Gases**—G. E. Deakins and C. M. Crain. (*Rev. Sci. Inst.*, vol. 27, pp. 606–608; August, 1956.) The electron density was measured and recorded photographically at times up to 500 μs after removal of a 24.5-mc ionizing voltage, using a cavity-resonator refractometer operating at about 9.4 kmc. Observations indicate that for commercial neon 90 per cent of the decay occurred within the first 250 μs.

537.56:538.56:538.6 410

**The Boltzmann Equation and the One-Fluid Hydromagnetic Equations in the Absence of Particle Collisions**—F. G. Chew, M. L. Goldberger, and F. E. Low. (*Proc. Roy. Soc. A*, vol. 236, pp. 112–118; July 10, 1956.) "Starting from the Boltzmann equation for a completely ionized dilute gas with no interparticle collision term but a strong Lorentz force, an attempt is made to obtain one-fluid hydromagnetic equations by expanding in the ion mass-to-charge ratio. It is shown that the electron degrees of freedom can be replaced by a macroscopic current. but true hydrodynamics still does not result unless some special circumstance suppresses the transport of pressure along magnetic lines of force. If the longitudinal transport of pressure is ignored, a set of self-contained one-fluid hydromagnetic equations can be found even though the pressure is not a scalar."

537.562 411

**The Dependence of the Electronic Recombination Coefficient on Temperature and Pressure**—I. L. Aptekar' and B. L. Timan. (*Zh. Tekh. Fiz.*, vol. 26, pp. 343–347; February, 1956.) Electronic recombination in gases is examined on the basis of statistical relations. The variation of the recombination coefficient thus determined coincides with that determined experimentally for ionic recombination in air.

537.562 412

**The Behaviour of Ion Columns in Pure Gases**—A. Müller. (*Z. Phys.*, vol. 145, pp. 469–485; June 15, 1956.) Measurements of recombination and diffusion in oxygen and in inert gases are reported.

538.3 413

**Some Problems relating to the Application of Hertzian Vectors**—K. Bochenek. (*Archiwum Elektrotech.*, vol. 4, pp. 247–268; English summary, pp. 266–268; 1955.) Two theorems on em fields are presented: in the first, the field is described by only one vector, electric or magnetic; in the second it is described by two vectors.

538.566:535.4]+534.2 414

**First Correction to the Geometric-Optics Scattering Cross-Section from Cylinders and Spheres**—S. I. Rubinow and T. T. Wu. (*J. Appl. Phys.*, vol. 27, pp. 1032–1039; September, 1956.) "The total scattering cross section in the short-wavelength limit is considered in this paper. The problems treated include diffraction of a plane electromagnetic wave by a conducting cylinder (two possible polarizations) or a conducting sphere, acoustic scattering by a rigid sphere, and quantum-mechanical scattering by an impenetrable sphere."

538.566:535.42/.43]+621.396.11+621.396.67 +[621.372.8:538.63 415

**Symposium on Electromagnetic Wave Theory**—IRE Trans, vol. AP-4, pp. 190–586; July, 1956.) Report of a conference held at the University of Michigan in June 1955, giving the text of 45 papers and abstracts of 54 others, together with some discussion; an author index is included. The main material is grouped under the headings 1) boundary-tube problems of diffraction and scattering theory, 2) forward and multiple scattering, 3) antenna theory and microwave optics, and 4) propagation in doubly-refracting media.

538.569.4:535.33 416

**Recording Magnetic-Resonance Spectrometer**—M. W. P. Strandberg, M. Tinkham, I. H. Solt, Jr, and C. F. Davis, Jr. (*Rev. Sci. Instr.*, vol. 27, pp. 596–605; August, 1956.) "Apparatus especially designed for studying electron paramagnetic resonance is described and discussed. A magnet of novel yokeless design is presented. Field stabilization and modulation procedure is considered. The microwave sample cavity is analyzed to determine conditions for optimum operation. The klystron stabilization problem is examined. Appropriate lumped circuits for low-frequency operation are described. The signal amplification and presentation system is treated in detail."

## GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 417

**Flux Measurements of Discrete Radio Sources at Frequencies below 30 Mc/s**—H. W. Wells. (*J. geophys. Res.*, vol. 61, pp. 541–545; September, 1956.) Measurements on Taurus A, Virgo A, Cygnus A and Cassiopeia A are reported; the amplitude ratios between pairs of these sources have different values at 26.75 and 18.5 mc.

523.16 418

**The Measurement of the Distance of Radio Sources by Interstellar Neutral Hydrogen Absorption**—D. R. W. Williams and R. D. Davies. (*Phil. Mag.*, vol. 1, pp. 622–636; July, 1956.) Three different methods are discussed for determining the distances from the absorption spectra. The technique has been applied to the intense sources in Cassiopeia, Cygnus, Taurus, and Sagittarius; some results are presented.

523.16 419

**The Low-Frequency Spectrum of the Cygnus (19 N4A) and Cassiopeia (23 N5A) Radio Sources**—R. J. Lamden and A. C. B. Lovell. (*Phil. Mag.*, vol. 1 pp. 725–737; August, 1956.) Intensity measurements have been made at frequencies of 16.5, 19.0, 22.6 and 30.0 mc, using the 218-ft-aperture transit radio telescope at Jodrell Bank. The intensity of both sources falls abruptly at frequencies below 22 mc; this is probably due to absorption in an interstellar HII region of average density and extent.

523.16:551.510.535 420

**Low-Angle Fluctuations of the Radio Star Cassiopeia as Observed at Ithaca, N.Y., and their Relation to the Incidence of Sporadic E**—B. Dueño. (*J. Geophys. Res.*, vol. 61, pp. 535–540; September, 1956.) Observations over the period September 1954–August 1955 are reported. Correlation between the radio-star fluctuations and the occurrence of $E_s$ was good during midwinter and poor afterwards. The $E_s$ layer would be expected to exert its greatest influence over the incoming radiation from Cassiopeia during midwinter, since the radiation then pierces through the ionosphere at a height of 400 km, about 1500 km north of Ithaca; at this point the normal sun-controlled forms of ionization would be at a minimum.

523.16:621.396.677 421
**Radio Astronomy and the Jodrell Bank Telescope**—A. C. B. Lovell. (*Proc. IEE*, Part B, vol. 103 pp. 711–721; November, 1956.) Text of the 47th Kelvin lecture. The 250-ft-aperture steerable radio telescope is to be used for studying both galactic and extragalactic sources of rf radiation over a wide frequency range. It will also be used as a combined transmitter and receiver for studies of the moon, and possibly of the planets. Very faint meteors and various solar/terrestrial relationships of particular importance to the International Geophysical Year will also be investigated. 57 references.

523.16:621.396.822:523.75 422
**Absorption of Cosmic Radio Noise at 22.2 Mc/s Following Solar Flare of February 23rd 1956**—S. E. Forbush and B. F. Burke. (*J. Geophys. Res.*, vol. 61, pp. 573–575; September, 1956.) Absorption effects noted in records obtained at an observatory near Washington are briefly discussed.

523.5:621.396.96 423
**Diurnal Variations in Forward-Scattered Meteor Signals**—C. O. Hines. (*J. Atmos. Terr. Phys.* vol. 9. pp. 229–232; October, 1956.)

523.75:551.510.535:621.396.11.029.45 424
**V.L.F. Phase Shifts associated with the Disturbance of February 23rd 1956**—Pierce. (See 557.)

550.3 425
**A Method for Drawing the Great-Circle Path between Any Two Points on Earth**—J. H. Meek. (*J. Geophys. Res.*, vol. 61, pp. 445–448; September, 1956.)

550.38:551.594.6 426
**Arc Lengths along the Lines of Force of a Magnetic Dipole**—S. Chapman & M. Sugiura. *J. Geophys. Res.*, vol. 61, pp. 485–488; September, 1956.) Formulas and tables for the arc length along the lines of force of a magnetic dipole are given with reference to the earth treated as a sphere. These tables may prove useful in connection with the study of radio whistlers and of the motion of charged particles along the lines of geomagnetic force.

550.380.8+550.37 427
**Constructional Details of the Inductive Magnetic-Pulsation and Earth-Current Measuring Equipment at the Geomagnetic Observatory at Fürstenfeldbruck [W. Germany]**—K. Burkhart. (*Geofis. Pura Appl.*, vol. 33, pp. 78–85; January–April, 1956. In German.)

550.380.8:621.317.444 428
**High-Altitude Measurements of the Earth's Magnetic Field with a Proton Precession Magnetometer**—Cahill and Van Allen. (See 526.)

550.385 429
**Mean Results of Geomagnetic Observations in Tromsø, Norway, for the Years 1930–50**—J. Frøshaug. (*J. Geophys. Res.* vol. 61, pp. 435–444; September, 1956.) Summarized tables of results are presented; graphs show short-term and long-term variations.

550.385:550.37 430
**The Origin of Earth Currents and their Probable Influence on the Earth's Magnetic Field**—K. Burkhart. (*Geofis. Pura Appl.*, vol. 33, pp. 49–77; January–April, 1956. In German.) For short distances between measuring points a definite relation between earth currents and magnetic variations is generally confirmed. To explain discrepancies over longer distances, not only the subsoil resistance within the area of measurement but the total earth current circuit must be considered. Anomalies in Japan and North Germany are analyzed in relation to the gradient of conductivity normal to the induced emf. Observations agree satisfactorily with theory based on inductively coupled circuits

551.5 431
**Pulsed Searchlighting the Atmosphere**—S. S. Friedland, J. Katzenstein, and M. R. Zatzick. (*J. Geophys. Res.*, vol. 61, pp. 415–434; September, 1956.) An optical reflection technique which has been used to measure the density profile of the upper atmosphere up to a height of 40 km uses a 50-Mlm light pulse of duration 20$\mu$s from a lamp at the focus of a 60-in. mirror, with a photomultiplier at the focus of a similar mirror. The results are in agreement with those obtained by previous investigators and by theoretical methods.

551.5 432
**Circulation in the Upper Atmosphere**—P. S. Pant. (*J. Geophys. Res.*, vol. 61, pp. 459–474; September, 1956.) A study covering the height range 20–100 km over the northern hemisphere. Summer and winter temperatures computed from the wind field, using rocket mean-pressure data and assuming that the wind field is geostrophic, agree with direct observations. Possible causes of observed seasonal temperature changes are discussed. 48 references.

551.51 433
**Gravitational and Thermal Oscillations in the Earth's Upper Atmosphere**—M. L. White. (*J. Geophys. Res.*, vol. 61, pp. 489-499; September, 1956.) Analysis given previously [1404 of 1956 (Sen and White)], for the case where heat is introduced only at the base of the atmosphere, is extended to apply to cases where heat is introduced at any region. Expressions obtained for the velocity components and pressure variation at any height are identical with the corresponding equations for the purely gravitational case, except for a new interpretation of the wave function. Expressions are also obtained for the vertical and horizontal displacements and for the amplification over equilibrium tide at any level. Results of the analysis are compared with those obtained by other workers.

551.510.535 434
**The Ionization of the E Layer**—H. Pichler. (*Geofis. Pura Appl.*, vol. 33, pp. 146–152; January–April, 1956. In German.) The effect of geographical location, season, and solar activity on the relation between the critical frequency $f_1E$ and the position of the sun is discussed. Comparison with observations shows that formulas previously derived *e.g.*, by Harnischmacher (1682 of 1950) are inadequate, probably owing to oversimplification, and that more extensive investigations are needed to clarify the correlation.

551.510.535 435
**The World-Wide Height Distribution of the $F_2$ Layer**—R. Eyfrig, E. Harnischmacher, and K. Rawer. (*Geofis. Pura Appl.*, vol. 33, pp. 153–162; January–April, 1956. In German.) An analysis is made of (M3000)$F_2$ maps covering America from high north magnetic latitudes to the magnetic equator, for years of sunspot minimum and maximum respectively. Generally a daily minimum occurs on the magnetic equator and a maximum at a geographical latitude of 40°N, the latter being less pronounced in summer than in winter. A sunrise maximum occurs irrespective of latitude. Increased solar activity is correlated with decreased M values and accentuation of the equatorial minimum.

551.510.535 436
**Physical Properties of the Atmosphere from 90 to 300 km**—H. K. Kallmann, W B. White, and H. E. Newell, Jr. (*J. Geophys. Res.* vol. 61, pp. 513–524; September, 1956.) A model atmosphere based on recent experimental and theoretical studies has the following features: molecular oxygen begins to dissociate at a height of about 90 km; at about 130 km about 30 per cent of $O_2$ is still in the undissociated state; molecular nitrogen begins to dissociate at about 220 km; the concentrations of $O_2$ and $N_2$ decrease exponentially with increasing height; the temperature ceases to vary with height at about 360 km. This model gives lower temperatures and densities throughout the ionosphere than have been deduced previously.

551.510.535 437
**Daytime Measurement of Positive and Negative Ion Composition to 131 km by Rocket-Borne Spectrometer**—C. Y. Johnson and J. P. Heppner. (*J. Geophys. Res.*, vol. 61, p. 575; September, 1956.) Negative ions of mass numbers 46, 32, 29, 22, and 16 were detected, but no positive ions. For night-time measurements, see 1409 of 1956.

551.510.535 438
**Neutral Gas Composition of the Upper Atmosphere by a Rocket-Borne Mass Spectrometer**—E. B. Meadows and J. W. Townsend, Jr. (*J. Geophys. Res.*, vol. 61, pp. 576–577; September, 1956.) Measurements at heights up to 141.6 km are reported. In addition to the usual constituents of air, components with mass numbers 46 and 23 were detected in the neighborhood of 85 km. See also 437 above.

551.510.535 439
**Resonance Scattering by Atmospheric Sodium: Part 2—Nightglow Theory**—J. W. Chamberlain and B. J. Negaard. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 169–178; October, 1956.) Continuation of previous discussion [3728 of 1956 (Chamberlain)]. Three models of the nightglow layer are considered in which the primary excitation is concentrated at the bottom of the layer, evenly distributed through the layer. or concentrated at the top of the layer.

551.510.535 440
**Resonance Scattering by Atmospheric Sodium: Part 3—Supplementary Considerations**—D. M. Hunten. (*J. Atmos. Terr. Phys.*, vol. 9, pp 179–183; October, 1956.) For the excitation mechanism assumed, the best model of the sodium layer is that in which the excitation is concentrated at the bottom. Part 2: 439 above.

551.510.535:523.75 441
**Ionospheric Changes associated with the Solar Event of 23 February 1956**—C. M. Minnis, G. H. Bazzard, and H. C. Bevan. (*J. Atmos. Terr. Phys.* vol. 9, pp. 233–234; October, 1956.) A discussion based on observations at Singapore, Inverness, and Slough. A magnetic crochet and intense ionospheric absorption were observed characteristic of an important solar flare. Effects associated with the incidence of high-speed and low-speed particles with the incidence of high-speed and low-speed particles and wave radiation are distinguished.

551.510.535:523.78 442
**The Response of the Ionosphere to the Solar Eclipse of 30 June 1954 in Great Britain**—C. M. Minnis. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 201–209; October, 1956.) "Vertical-incidence measurements of virtual height and critical frequency were made at two points

Woodruff. (*Phys. Rev.*, vol. 103, pp. 1159–1166; September 1, 1956.)

537.311.33:546.28 479
**Line Broadening of an Impurity Spectrum in Silicon**—D. Sampson and H. Margenau. (*Phys. Rev.*, vol. 103, pp. 879–885; August 15, 1956.) A simplification of the method of calculation presented by Lax and Burstein (1086 of 1956).

537.311.33:546.28 480
**Nuclear Magnetic Resonance of Si$^{29}$ in *n*- and *p*-Type Silicon**—R. G. Shulman and B. J. Wyluda. (*Phys. Rev.*, vol. 103, pp. 1127–1129; August 15, 1956.)

537.311.33:546.281.26 481
**The Effect of Impurities on the Electrical Conductivity and Nonlinearity of Silicon Carbides**—L. I. Ivanov and V. I. Pruzhinina-Granovskaya. (*Zh. Tekh. Fiz.*, vol. 26, pp. 220–231; January, 1956.) A method is described for determining the nonlinearity and resistance of powder semiconductors, and the dependence of these two parameters on the granular composition of the powders and on the pressure is discussed. Experiments have shown that impurities greatly affect the properties of silicon carbides.

537.311.33:546.289 482
**Visual Evidence of Inversion Layers on Semiconductor Materials**—P. Levesque. (*J. Appl. Phys.*, vol. 27, pp. 1104–1105; September, 1956.) Observations on etched Ge specimens, made by a technique involving deposition of a $BaTiO_3$ layer, are briefly reported.

537.311.33:546.289 483
**An Investigation of the Resistance Variation of Germanium in a Magnetic Field**—V. I. Stafeev and V. M. Tuchkevich. (*Zh. Tekh. Fiz.*, vol. 26, pp. 273–276; February, 1956.) The experimental curves $\Delta\rho/\rho_0$ vs magnetic field for different temperatures show that the variation is nonlinear both for strong and weak fields, the power indices depending on temperature. At relatively low temperatures, of the order of 170°K, $\Delta\rho/\rho_0$ varies more slowly than the square of the mobility. These findings are at variance with existing theory. For all *p*-type specimens the temperature variation curve exhibits a maximum, as predicted by theory.

537.311.33:546.289 484
**The Diffusion of Impurities in Germanium**—B. I. Boltaks. (*Zh. Tekh. Fiz.*, vol. 26, pp. 457–474; February, 1956.) A comprehensive review of the experimental data, divided into the following sections: 1) methods for measuring the diffusion coefficients in semiconductors; 2) diffusion and electrical activity of impurities in Ge; 3) temperature dependence of the diffusion coefficients of impurities in Ge, and 4) diffusion and solubility of impurities in Ge. A close connection between the diffusion and solubility of impurities and the electrical properties of Ge is indicated. The diffusion and solubility exhibit a number of important peculiarities which do not exist in metals and alloys. 48 references.

537.311.33:546.289 485
**Field-Induced Conductivity Changes in Ge**—H. C. Montgomery and W. L. Brown. (*Phys. Rev.*, vol. 103, pp. 865–870; August 15, 1956.) Measurements were made on filament specimens with fields of strength up to $10^5$V/cm normal to the surface. From observations of the minimum in the conductance curve, information can be derived on surface states as a function of surface potential. Data obtained using various gaseous ambients are compared with the contact-potential observations of Brattain and Bardeen (1698 of 1953). Measurements at temperatures down to 170°K give information about properties of "fast" surface states. Phase-shift loops observed at frequencies around 1 kc are interpreted in terms of minority-carrier lifetime.

537.311.33:546.289 486
**Acceptors Quenched into Germanium**—S. Mayburg. (*Phys. Rev.*, vol. 103, pp. 1130–1131; August 15, 1956.) Discrepancies between results obtained previously by the author (148 of 1955) and other workers [1765 of 1956 (Hopkins and Clarke) and 2793 of 1956 (Logan)] are explained by taking note of dislocation effects.

537.311.33:546.289 487
**Use of Infrared Absorption in Germanium to determine Carrier Distributions for Injection and Extraction**—N. J. Harrick. (*Phys. Rev.*, vol. 103. pp. 1173–1181; September 1, 1956.) Continuing the work reported previously (2109 of 1956), measurements of infrared absorption have been used to determine longitudinal distributions of free carriers for the cases of field-opposed and field-aided injection and extraction.

537.311.33:546.289 488
**Elastogalvanomagnetic Effect and Intervalley Scattering in *n*-Type Germanium**—R. W. Keyes. (*Phys. Rev.*, vol. 103, pp. 1240–1245; September 1, 1956.) "Starting from a multivalley model of a cubic semiconductor, a calculation of the effect of elastic strain on certain galvanomagnetic effects is carried out. It is found that the effects are sensitive to the strength of the intervalley scattering. An experiment on Ge to which the calculation is applicable is described. It is concluded from a comparison of the theory and the experiment that the coupling constant which characterizes the coupling of the electrons to the intervalley phonons in the model of Herring [2642 of 1955] is considerably smaller than the coupling constant to the acoustic phonons.

537.311.33:546.289 489
**Carrier Accumulation in Germanium**—J. B. Arthur, A. F. Gibson, and J. B. Gunn. (*Proc. Phys. Soc.*, vol. 69. pp. 697–704; July 1, 1956.) An account is given of experiments on junctions between lightly and heavily doped regions of crystals of the same conductivity type; these are designated L-H junctions. They can be made largely impermeable to minority carriers but permeable to majority carriers, and high concentrations of minority carriers can be accumulated at such junctions.

537.311.33:546.289 490
**Current Gain at L-H Junctions in Germanium**—J. B. Arthur, A. F. Gibson, and J. B. Gunn. (*Proc. Phys. Soc.*, vol. 69, pp. 705–711; July 1, 1956.) The relatively low permeability to minority carriers of junctions between lightly and heavily doped crystals of the same type (see 489 above) can be considered as a reduction of the minority carrier mobility, and gives rise to enhanced current gain in certain structures, for example filamentary transistors. Experiments confirming the existence of the effect and indicating its magnitude are described.

537.311.33:546.289:535.215:538.639 491
**The Transverse Photomagnetic Effect in *n*- and *p*-Type Germanium**—I. K. Kikoin and Yu. A. Bykovski. (*C.R. Acad. Sci. U.R.S.S.*, vol. 109, pp. 735–736; August 1, 1956. In Russian.) The effect investigated is the following: if a semiconductor in a magnetic field $H$, with components $H_z$ and $H_y$ in the $x$ and $y$ directions, is illuminated by light in the direction of the $y$ axis, then in addition to the emf proportional to $H$ along the $z$ axis an emf is produced in the $x$ direction proportional to $H_zH_y$. The emf $E$ is plotted against $H$ for *p*-type Ge at temperatures of +20°, −20°, and −100°C, and against temperature for *p*- and *n*-type Ge at $H=20000$ oersted.

537.311.33:546.482.21 492
**Redistribution of the Electron Density in Cadmium Sulphide Crystals associated with Changes in its Electrical Conductivity**—Yu. N. Shuvalov. (*C.R. Acad. Sci. U.R.S.S.*, vol. 109 pp. 753–756; August 1, 1956. In Russian.) Results of an X-ray investigation of hexagonal crystals are reported.

537.311.33:546.682.19 493
**The Electrical Properties of InAs**—D. N. Nasledov and A. Yu. Khalilov. (*Zh. Tekh. Fiz.*, vol. 26, pp. 251–254; February, 1956.) An investigation was made over the temperature range 1.3°–673°K and in magnetic fields up to 33,000 oersted. The main results are as follows: 1) conductivity is practically independent of temperature from 1.3°K to room temperature; above this temperature it rapidly increases; 2) the concentration of carriers at room temperature is about $10^{17}$/cm$^3$; 3) in a magnetic field of 20,000 oersted, resistance increases fourfold at 400°C; 4) mobility increases greatly at high temperatures and reaches 120,000 cm per V/cm.

537.311.33:546.682.86 494
**Electrical Properties of InSb**—D. N. Nasledov and A. Yu. Khalilov. (*Zh. Tekh. Fiz.*, vol. 26, pp. 6–14; January, 1956.) Specimens with *n*- and *p*-type conductivity were investigated over a range of temperatures from 1.3° to 673°K. Curves show the variation of conductivity with temperature and the effect of a magnetic field of strength 20,000 oersted. The Hall effect was also investigated for various compositions, using magnetic field intensities up to 33,000 oersted. A theoretical interpretation of the results is given.

537.311.33:546.682.86 495
**Indirect Transitions in Indium Antimonide**—R. F. Potter. (*Phys. Rev.*, vol. 103, pp. 861–862; August 15, 1956.) "An interpretation of published infrared absorption data for InSb is made. It is proposed that the absorption beyond the main optical edge is due to indirect transitions involving both optical mode and low-energy acoustical mode phonons."

537.311.33:546.817.221 496
**Calculation of the Cohesive Energy of Zincblende**—S. Asano and Y. Tomishima. (*J. Phys. Soc. Japan*, vol. 11, pp. 644–653; June, 1956.)

537.311.33:546.817.221:539.234 497
**Surface Structure of Thin Layer of PbS obtained by Evaporation in Vacuum**—R. Ya. Berlaga, M. I. Rudenok, and L. P. Strakhov. (*Zh. Tekh. Fiz.*, vol. 26, pp. 3–5; January, 1956.)

537.311.33:621.396.822 498
**Shot Noise in Semiconductors**—L. Ya. Pervova. (*Radiotekhnika i Elektronika*, vol. 1, pp. 98–105; January, 1956.) An expression is derived for the spectral density of shot noise in an arbitrary length of semiconductor taking into account the true lifetimes, $\tau$, of electrons in the conduction zone (instead of the average lifetime) and the length $L$ of the semiconductor. For $\gamma \ll L$, where $\gamma$ is the free path of the electron, the noise spectral density is proportional to $1/\omega$ in the range $\omega\tau \simeq 1$ to $\omega\tau \approx 10^3$; for $\omega\tau < 1$, the noise intensity is independent of frequency, for $\omega\tau > 10^3$ the noise varies as $1/\omega^2$. For $\gamma > L$ the spectral density remains nearly constant up to $\omega\tau \approx \lambda/L$, and is proportional to $1/\omega$ for $\omega\tau > \lambda/L$.

538.22 499
**The Magnetic Properties of FeSe$_x$ with NiAs Structure**—T. Hirone and S. Chiba. (*J. Phys. Soc. Japan*, vol. 11, pp. 666–670; June, 1956.) In the range 48.8 per cent–53.1 per cent Se, at room temperature, the compound is made up of two phases, $\alpha$ and $\beta$; the $\alpha$ phase

has PbO structure and is weakly paramagnetic or antiferromagnetic, while the $\beta$ phase has NiAs structure and is ferrimagnetic.

538.221 500
**A Symposium on Magnetism.—London, September 1955**—A. C. Lynch and J. Watkins. (*Brit. J. Appl. Phys.*, vol. 7, pp. 236–242; July, 1956.) Summaries are presented of papers and discussion on theoretical studies and applications, particularly of ferrites, and including the use of neutron diffraction techniques.

538.221 501
**Curvature of the Bloch Wall as an Elementary Process in Reversible Variations of Magnetization (Initial Permeability and $\Delta E$-Effect)**—M. Kersten. (*Z. Angew. Phys.*, vol. 8, pp. 313–322; July, 1956.) The observed temperature variations of initial permeability and $\Delta E$-effect can be explained if cylindrical rather than cushion-type or spherical walls are assumed.

538.221 502
**The Magnetic Properties of Chromium-Tellurium-Selenium Systems**—I. Tsubokawa. (*J. Phys. Soc. Japan*, vol. 11, pp. 662–665; June, 1956.) Properties of compounds having the formula $Cr(Te_{1-x}Se_x)$ are discussed. They are ferromagnetic in the range $0<x<0.8$; outside this range they are paramagnetic or antiferromagnetic.

538.221 503
**Effect of Magnetic Field Strength during Condensation on the Coercivity and Form of Vapour-Deposited Iron**—A. J. Griest, J. F. Libsch, and G. P. Conard. (*J. Appl. Phys.*, vol. 27, pp. 1022–1024; September, 1956.) The size, shape, and coercivity of Fe particles deposited from vapor on to a Cu surface were all influenced by the temperature of condensation and the strength of the applied magnetic field. The coercivity increased with the field strength, but the amount of the increase decreased as the condensation temperature was raised. The observed coercivity values of 30–120 oersted were related to strain in the deposited material.

538.221:621.318.122 504
**Magnetic After-Effect in $Ni_2Mn$ Alloy**—T. Taoka. (*J. Phys. Soc. Japan*, vol. 11, pp. 537–547; May, 1956.) Continuation of work described previously [2042 of 1955 (Taoka and Ohtsuka)]. Experimental results are reported in detail. An electron-microscope study confirms that alloys showing marked after-effect consist of numerous single-domain particles of prolate ellipsoidal form embedded in a matrix.

538.221:621.318.122 505
**Study of Magnetic Annealing on $Ni_3Fe$ Single Crystal**—S. Chikazumi. (*J. Phys. Soc. Japan*, vol. 11, pp. 551–558; May, 1956.)

538.221:621.318.124 506
**Investigations on Barium Ferrite Magnets**—K. J. Sixtus, K. J. Kronenberg, and R. K. Tenzer. (*J. Appl. Phys.*, vol. 27, pp. 1051–1057; September, 1956.) Report of an experimental investigation of the relation between sintering temperature, particle size and coercive force in $BaO{:}6Fe_2O_3$. With particles of linear dimensions $>10^{-3}$ cm, domain-wall movements are observed on application of a magnetic field; for dimensions $<5\times10^{-4}$ cm the observations indicate reversals of magnetization by the rotation process. With particles of intermediate size, both processes occur.

538.221:621.318.13 507
**Some Properties of the Coercive Force in Soft Magnetic Materials**—D. S. Rodbell and C. P. Bean. (*Phys. Rev.*, vol. 103, pp. 886–895; August 15, 1956.) A discussion in terms of domain-wall movements indicates that the coercive force is composed of two terms, one characteristic of the bulk material and the other arising from the pinning of domain walls at the surfaces. The observed increase in coercive force with decrease in thickness of sheet specimens is thus explained. See also 1126 of 1956.

538.221:621.318.13.014.4 508
**The Eddy-Current Anomaly in Electrical Sheet Steel**—H. Aspden. (*Proc. IEE*, Part C, vol. 103, pp. 272–278; September, 1956.) "A theory which accounts for the well-known discrepancy between predicted eddy-current losses in electrical sheet steels and the experimentally observed values is presented. The anomaly is shown to be due partly to the magnetic inhomogeneity arising from ferromagnetic domain structure and partly to a time-lag effect caused by the finite speed of domain boundary movements. A new experimental approach to the study of the eddy-current anomaly is described. This involves the use of a method of measuring the anomaly factor as it applies instantaneously at a point in a magnetization cycle."

538.221:621.318.13.014.4 509
**Magnetic Time-Lag Effects in Solid Steel Cores**—H. Aspden. (*Proc. IEE*, Part C, vol. 103, pp. 279–285; September, 1956.) The discrepancy between theoretical and actual eddy-current effects in thick steel cores is found to be attributable to time-lag in the magnetization process, the magnitude of which decreases with increasing frequency, its value being about $10^{-5}$s at 10 kc.

538.221:621.318.134 510
**Cation Distributions in Ferrospinels. Theoretical**—H. B. Callen, S. E. Harrison, and C. J. Kriessman. (*Phys. Rev.*, vol. 103, pp. 851–856; August 15, 1956.) The thermodynamic relation between the ion distribution and the nonthermal part of the internal energy function is developed. The results are used in a separate paper [*ibid.*, pp. 857–860 (Kriessman and Harrison)] to interpret observations on Mg-Mn ferrites.

538.221:621.318.134 511
**On the Uniaxial Anisotropy Induced by Magnetic Annealing in Ferrites**—S. Taniguchi and M. Yamamoto. (*J. Phys. Soc. Japan*, vol. 11, pp. 604–605; May, 1956.)

538.221:621.318.134:621.372.8 512
**Relations between the Structure of Ferrites and Conditions for Their Resonance in Waveguides. Unidirectional Guides**—J. Suchet. (*Onde Élect.*, vol. 36, pp. 508–519; June, 1956.) An examination is made of the influence on the behavior of ferrites of their composition and method of preparation, and the unidirectional propagation characteristics of waveguides enclosing Ni-Zn and Ni-Al ferrites are calculated.

548.0 513
**Observations of Dislocation Glide and Climb in Lithium Fluoride Crystals**—J. J. Gilman and W. G. Johnston. (*J. Appl. Phys.*, vol. 27, pp. 1018–1022; September, 1956.)

621.315.61+621.318.1+621.315.3 514
**Materials for Electronics**—J. Markus and D. A. Findlay. (*Electronics*, vol. 29, pp. 185–216; October, 1956.) A survey of U. S. products, including insulating and magnetic materials, wires, and solders.

621.315.612.6 515
**Thermal Expansion of Binary Alkali Silicate Glasses**—H. F. Shermer. (*J. Res. Nat. Bur. Stand.*, vol. 57, pp. 97–101; August, 1956.)

621.315.616:[537.531.9+539.166.9 516
**Radiation-Induced Conductivity in Polyethylene and Teflon**—R. A. Meyer, F. L. Bouquet, and R. S. Alger. (*J. Appl. Phys.*, vol. 27, pp. 1012–1018; September, 1956.) Conductivity changes induced by bombardment with X rays and gamma rays have been studied as a function of time, temperature, specimen geometry, exposure rate, and applied electric field. Observed photocurrents were directly proportional to exposure rate and field. The conductivity increased by a factor of about $10^3$ during irradiation and the photocurrent was nearly independent of temperature over the range 78°–273°K.

## MATHEMATICS

517 517
**Incomplete Bessel and Struve Functions**—W. H. Steel and J. Y. Ward. (*Proc. Camb. Phil. Soc.*, vol. 52, pp. 431–441; July, 1956.) "Some properties are given of the incomplete Bessel and Struve functions defined by a Poisson-type integral. These functions are tabulated for the orders 0 and 1."

517:518.2 518
**Laguerre Functions: Tables and Properties**—J. W. Head and W. P. Wilson. (*Proc. IEE*, Part C, vol. 103, pp. 428–440; September, 1956.) "The first 21 Laguerre functions have been tabulated to four decimal places (or two significant figures when this is more accurate) for values of the argument 0(0.1)1(0.2)3(0.5)6(1)14(2)40(5)100. The practical applications of these functions are briefly considered. The zeros of these Laguerre functions are also tabulated to 5 places, the last of which is not reliable except in the case of the smallest zero for orders above 6. Differentiation and integration of Laguerre functions, orthonormal properties and addition formulas are briefly discussed. Finite or infinite series of Laguerre functions whose sums are well-known functions such as products of polynomials and circular or exponential functions, step and delta functions, Bessel and Kelvin functions, and gamma and error functions are stated."

512.9 519
**Champs de Vecteurs et de Tenseurs [Book Review]**—E. Bauer. Masson, Paris, 204 pp., 2200 fr. (*Phil. Mag.*, vol. 1, p. 691; July, 1956.) A straightforward account of vectors, tensors and electromagnetism, in a form ready for application to engineering problems.

517 520
**Numerical Analysis [Book Review]**—Z. Kopal. Chapman and Hall, London, 556 pp., 63s. 1955, (*Brit. J. Appl. Phys.*, vol. 7, pp. 269–270; July, 1956.) A comprehensive handbook; emphasis is on the application of numerical techniques to problems of infinitesimal calculus in a single variable.

## MEASUREMENTS AND TEST GEAR

621.314.7.001.4(083.74) 521
**IRE Standards on Solid-State Devices: Methods of Testing Transistors, 1956.**—(PROC. IRE, vol. 44, pp. 1542–1561; November, 1956.) Standard 56 I.R.E. 28. S2.

621.317.3.029.6:621.372.5 522
**Microwave Measurements with a Lossy Variable Termination**—H. M. Altschuler and A. A. Oliner. (*Proc. IEE*, Part C, vol. 103, pp. 392–399; September, 1956.) The problem is treated by considering the lossy termination as the combination of a purely lossy quadripole ("reflection-coefficient transformer") with a loss-free termination. The measurement data can then be analyzed as for the familiar case of an actual loss-free termination.

621.317.32:537.525.8 523
**"Glo-Ball" Development**—J. F. Steinhaus. (*Rev. Sci. Inst.*, vol. 27, pp. 575–580; August, 1956.) The "glo-ball" is a thin-walled partially evacuated glass sphere containing helium, used for investigating electric field distributions in resonant cavities; spheres with diameters ranging between $\frac{1}{4}$ inch and 1 inch have been used in studying linear accelerators. The indication of the field strength is given by ionization of the

gas in the device. Technique for producing "glo-balls" with stable and relatively low ionization thresholds is discussed.

621.317.382:621.317.784:538.632:537.311.33 524
**The Application of the Hall Effect in a Semiconductor to the Measurement of Power in an Electromagnetic Field, and the Design of Semiconductor Wattmeters for Power-Frequency and Audio-Frequency Applications**—H. E. M. Barlow. (*Proc. IEE*, Part B, vol. 103, p. 710; November, 1956.) Discussion on 1735 and 1744 of 1955.

621.317.431 525
**Magnetic-Switch B/H Loop Tracer**—W. Geyger. (*Electronics*, vol. 29, pp. 167–169; October, 1956.) Magnetic switching of Si-diode chopper circuits enables dynamic hysteresis loops to be traced at frequencies up to 20 kc.

621.317.444:550.380.8 526
**High-Altitude Measurements of the Earth's Magnetic Field with a Proton Precession Magnetometer**—L. J. Cahill, Jr, and J. A. Van Allen. (*J. Geophys. Res.*, vol. 61, pp. 547–558; September, 1956.) An instrument of the Packard-Varian type (*Phys. Rev.*, vol. 93, p. 941; February 15, 1954), has been used for measurements at heights up to 100,000 ft. The equipment, including magnetic head, amplifier, and telemetering transmitter, is carried by a plastic balloon. The results indicate that the field strength decreases with increasing height, and that there are large variations with geographical position. An instrument that can be carried to a height of 150 km by rocket has been constructed.

621.317.7:621.373.4 527
**Phase Generator for Tropospheric Research**—R. W. Hubbard and M. C. Thompson, Jr. (*Electronics*, vol. 29, pp. 220–223; October, 1956.) A laboratory-standard instrument based on earlier work [*Rev. Sci. Instr.*, vol. 26, pp. 617–618; June, 1955 (Thompson)], gives pulsed or sinusoidal signals with phase displacements up to 360° in steps of 2°.

621.317.7.082.64:621.316.86 528
**The Indirectly Heated Thermistor as a Precise A.C.-D.C. Transfer Device**—F. C. Widdis. (*Proc. IEE*, Part B, vol. 103, pp. 693–703. Discussion, pp. 707–709; November, 1956.) Methods are described for eliminating thermal drift in indirectly heated thermistors used as transfer devices for measurements of voltages and currents, thus rendering the device suitable for precision use over a wide range of audio frequencies. The sensitivity is high, but this advantage is to some extent outweighed by the slow response. dc reversal errors due to Peltier and Thomson effects are small.

621.317.733:621.317.335.3 529
**A Bridge for the Measurement of Permittivity**—A. M. Thompson. (*Proc. IEE*, Part B, vol. 103, pp. 704–707. Discussion, pp. 707–709; November, 1956.) "The direct admittance of a 3-terminal capacitor with the sample as dielectric is measured as a complex capacitance, the two components being indicated directly on two 3-terminal variable air capacitors. In addition to these the bridge network comprises transformer ratio-arms and an amplifier whose output voltage is in quadrature with that of the transformer. The bridge operates at ten fixed frequencies from 30 to $10^6$ rad/sec."

621.317.733:621.317.4:538.221 530
**Investigation of an Alternating-Current Bridge for the Measurement of Core Losses in Ferromagnetic Materials at High Flux Densities**—I. L. Cooter and W. P. Harris. (*J. Res. Nat. Bur. Stand.*, vol. 57, pp. 103–112; August, 1956.) Accurate values can be obtained at high flux densities if a correction term is included derived from the harmonic components of the exciting current.

621.317.755 531
**Equipment for the Continuous Vectorial Display of Alternating Voltages in the Frequency Range 5 kc/s to 3 mc/s**—E. C. Pyatt. (*J. Brit. IRE*, vol. 16, pp. 563–567; October, 1956.) A modified form of the equipment described previously (1826 of 1956).

621.317.755 532
**Polar-Coordinate Oscilloscopes**—G. F. Craven. (*Electronic Eng.*, vol. 28, pp. 422–425; October, 1956.) A circular-timebase circuit with radial modulation is described. using a three-phase Magslip generator.

621.317.755 533
**VIAC—a Variable-Interval Automation Controller**—M. L. Klein, H. C. Morgan, and J. R. Wood. (*Electronic Eng.*, vol. 28, pp. 425–429; October, 1956.) Designed for use in high-speed oscillograph recording, the unit described permits the actual running time of the recording system to be restricted to periods of particular interest, so as to conserve paper without loss of important information.

621.317.761 534
**A Precision Frequency Meter Produced by the Noiseau Measurement Center**—J. Boulin. (*Onde Élect.*, vol. 36, pp. 532–540; June, 1956.) Description of receiver-type equipment designed by the Direction des Services Radioélectriques des P.T.T. to cover the frequency range 3–27 mc, using harmonics of 10 kc calibrated by comparison with a 100-kc standard; an accuracy of $10^{-7} \pm 1$ cps is achieved.

621.317.784.029.4 535
**An Audio-Frequency Dynamometer Wattmeter**—A. H. M. Arnold and J. J. Hill. (*Proc. IEE*, Part C, vol. 103, pp. 325–333; September, 1956.)

621.317.79:537.228.1 536
**Apparatus for Investigating the Piezoelectric Properties of Crystals**—M. I. Yaroslavski, R. M. Lyutenberg, and V. N. Chernyshov. (*Zh. Tekh. Fiz.*, vol. 26 pp. 439–441; February, 1956.) The operation of the apparatus described is based on the fact that vibrations of a piezoelectric crystal continue for a time after the cessation of an excitation pulse.

## OTHER APPLICATIONS OF RADIO AND ELECTRONICS

538.566.029.6:61 537
**Hazards due to Total Body Irradiation by Radar**—H. P. Schwan and K. Li. (PROC. IRE, vol. 44, pp. 1572–1581; November, 1956.) The modes of propagation of em waves in the various tissues of the human body and the resulting development of heat are discussed. At frequencies $\ll$1kmc or $>$3kmc about 40 per cent of incident radiation is absorbed at a body/air interface; over the intervening frequency range the amount absorbed varies between 20 per cent and 100 per cent. Radiation at frequencies $<$1kmc must be guarded against particularly, since, these frequencies produce deep heating and may not be detected by the sensory organs in the skin. It is suggested that an irradiation level of 0.01W/cm² is tolerable.

621-52:061.3 538
**Progress in Mechanics under the Combined Influence of Mechanical and Electronic Developments**—(*Onde Élect.*, vol. 36, pp. 567–675; July, 1956.) The main part of this issue comprises 13 papers dealing with theoretical, practical, and economic aspects of automation, the use of automatic control apparatus on railways, and specific examples of electronic applications in the testing of fuels and engines, the manufacture of cables, and the classification and retrieval of documents. Several of these papers were presented at a conference held in Paris on November 5, 1955, under the above title.

621.317.79:531.771 539
**An Electronic Governor**—S. C. Hine. (*Electronic Eng.*, vol. 28, pp. 448–449; October, 1956.) A tachometer-generator driven by the motor to be governed feeds a frequency discriminator whose output controls the power suppled to the motor.

621.317.79:616-1-9 540
**A Sensitive Quick—Reacting Cardiotachometer**—M. Manzotti. (*Electronic Eng.*, vol. 28, pp. 446–448; October, 1956.)

621.384.6 541
**Theory of Strong Focusing [in particle accelerators]**—E. L. Burshtein and L. S. Solov'ev. (*C.R. Acad. Sci. U.R.S.S.*, vol. 109, pp. 721–724; August 1, 1956. In Russian.)

621.384.6 542
**Experimental Test of the Fixed-Field Alternating-Gradient Principle of Particle Accelerator Design**—L. W. Jones, K. M. Terwilliger, and R. O. Haxby. (*Rev. Sci. Instr.*, vol. 27, pp. 651–652; August, 1956.)

621.384.612:538.56.029.6 543
**Coherent Radiation from Electrons in the Synchrotron at Centimetre Wavelengths**—A. M. Prokhorov. (*Radiotekhnika i Electronika*, vol. 1, pp. 71–78; January, 1956.) An experimental investigation is reported of the radiation from 5-mev electrons traveling in an orbit with a radius of 8 cm. The measured radiated power at 2 and 3 cm λ was about $10^{-6}$ w, in agreement with calculations. It is suggested that an increase in the radiated power by a factor of $10^6$ is possible, and hence powers of 1 w at 1 cm λ, $10^{-2}$ w at 1 mm λ and $10^{-4}$ w at 0.1 mm λ should be obtainable.

621.385.833 544
**Advance Calculation of Magnetostatic Electron-Optical Imaging Systems**—B. V. Borries and F. Lenz. (*Optik, Stuttgart*, vol. 13, pp. 264–276; 1956.) The basic problem in permanent-magnet lenses for electron microscopes is to combine maximum refractivity with minimum external dimensions. Methods are presented for calculating the gap flux from the geometrical arrangement of the magnets and the soft-iron parts. Control of refractivity by displacement of iron parts within the system is discussed.

621.385.833 545
**A Model permitting Rigorous Calculation of Electronic Immersion Cylindrical Lense**—H. Grümm. (*Optik, Stuttgart*, vol. 13, pp. 277–288; 1956.) The model involves fields which are conveniently investigated by introduction of elliptic coordinates.

621.387:621.398 546
**The Dekatron in a Digital Data Transmission System**—G. Shand and C. Dean. (*J. Brit IRE*, vol. 16, pp. 533–542; October, 1956.)

621.387.4:519.2 547
**On a Probability Problem arising in the Theory of Counters**—L. Takács. (*Proc. Camb. Phil. Soc.*, vol. 52, pp. 488–498; July, 1956.)

621.396.91:621.398 548
**New Design Aspects of Boomerang Radiosondes for Upper Air Research**—H. J. Albrecht. (*Geofis. Pura Appl.*, vol. 33, pp. 121–145; January–April, 1956. In English.) Description of an electronically controlled glider carrying upper-air research equipment. The ascent is by means of normal balloons, and after release the glider is guided in the direction of the base transmitter. The history and details of its aerodynamic and electronic design are given, together with possible applications. The tech-

nique affords safe and prompt return of equipment and simplification of the calibration procedure; a saving of about 60 per cent over normal radio-sonde systems is claimed. 30 references.

621.398:621-52 549
**Integral Control with Torque Limitation**—J. C. West and M. J. Somerville. (*Proc. IEE*, Part C, vol. 103, pp. 407–419; September, 1956.) "The behavior of a basic remote-position-control system with velocity-feedback stabilization and integral-of-error control is investigated, taking torque limitation of the servo motor into account. It is shown that the transient response deteriorates as the magnitude of the input signal is increased and becomes unstable above a critical value."

621.56:537.311.33:537.322.1 550
**Thermoelectric Microrefrigerators**—E. K. Iordanishvili and L. S. Stil'bans. (*Zh. Tekh. Fiz.*, vol. 26, pp. 482–483; February, 1956.) A brief preliminary report on experiments with semiconductor elements for refrigeration.

681.81:621.373.431 551
**The Univox**—A. Douglas. (*Electronic Eng.*, vol. 28. pp. 434–437; October, 1956.) An electronic musical instrument is described using a hard-tube sawtooth generator of high constancy of tuning to produce the tones.

## PROPAGATION OF WAVES

621.396.11+[538.566:535.42/.43]+621.396.67 +[621.372.8:538.63 552
**Symposium on Electromagnetic Wave Theory**—(See 415.)

621.396.11:621.396.65 553
**Tropospheric Scatter Propagation**—J. A. Saxton. (*Wireless World*, vol. 62, pp. 587–590; December, 1956.) A nonmathematical review is presented of the theoretical background, and practical aspects of point-to-point tropospheric-scatter radio links are discussed.

621.396.11 554
**The Present State of Research in the Field of Tropospheric-Scatter Propagation**—J. Grosskopf. (*Nachrichtentech. Z.*, vol. 9. pp. 272–279; and 315–329; June/July, 1956.) Summary of results of theoretical and experimental work carried out mainly in the U.S.A.

621.396.11 555
**Useful Bandwidth in Scatter Transmission**—J. P. Voge. (Proc. IRE, vol. 44, pp. 1621–1622; November, 1956.) Discussion of recent publications on the subject, with reference to previous work by the author (*e.g.*, 510 and 3666 of 1954).

621.396.11.029.45:551.510.535 556
**The Polarization of Very Long Radio Waves Reflected from the Ionosphere at Oblique Incidence**—W. C. Bain and C. B.-I. Glass. (*Proc. IEE*, Part C, vol. 103, pp. 447–448; September, 1956.) A re-examination of the original evidence [2871 of 1952 (Bain *et al.*)], coupled with a study of additional data, indicates that the mean value of the ratio between the vertical and horizontal electric fields is 5.

621.396.11.029.45:551.510.535:523.75 557
**V.L.F. Phase Shifts associated with the Disturbance of February 23rd 1956**—J. A. Pierce. (*J. Geophys. Res.*, vol. 61, pp. 475–483; September, 1956.) Observations at Cambridge, Mass., on 16-kc signals from Rugby are discussed. During the disturbance which followed the solar flare at about 0334 U.T., the phase of the received wave was advanced by about 250°. In contrast to the effects observed with sudden ionospheric disturbances in the daytime, which appear to involve phase-shift as well as height variations of the reflecting layer, the night-time observations can be interpreted on the basis of height variations only. The phenomena are discussed in relation to reports of associated observations [*e.g.*, 2716 of 1956 (Ellison and Reid)].

621.396.029.55 558
**Rapid Method of M.U.F. Prediction**—R. Gea Sacasa. (*Rev. Telecomunicación, Madrid*, vol. 11, pp. 54–59; June, 1956.) Based on the "Spanish Method" (see *e.g.*, 3727 of 1955), nomograms (with captions in Spanish and English) are presented for the rapid assessment of muf and optimum working frequencies for broadcasting, standard-frequency transmissions, amateur, and mobile marine services. The Madrid-New York hour chart is shown with examples to explain its use.

621.396.11.029.55:551.510.535 559
**The Intensity of the So-Called Pedersen Ray**—K. Rawer. (*C.R. Acad. Sci., Paris*, vol. 243, pp. 797–798; September 10, 1956.) A theoretical calculation indicates that while the geometrical attenuation of the Pedersen ray reflected from the ionosphere is greater than that of the normal ray, absorption in the ionospheric layers is greater for the normal ray. The Pedersen ray may in some circumstances be the stronger of the two.

621.396.11.029.6 560
**Field-Strength Calculations for V.H.F. and U.H.F. Transmitters over Level Ground**—H. van der Hak. (*PTT-Bedrijf*, vol. 7, pp. 64–68; July, 1956.) Formulas are derived to simplify calculations for the interference and diffraction regions, and numerical examples clarify the use of 13 nomograms included as loose sheets with the journal.

621.396.11.029.62 561
**Some Aircraft Measurements of Beyond-the-Horizon Propagation Phenomena at 91.3 Mc/s**—B. J. Starkey. (*Proc. IEE*, Part B, vol. 103, pp. 761–763; November, 1956.) "Field-strength measurements at distances extending far beyond the horizon from a transmitter on a frequency of 91.3 mc have been carried out in an aircraft flying at a height of 10,000 ft. The analysis of the results obtained and their correlation with meteorological data suggest that many phenomena of long-distance propagation could possibly be explained by the simple hypothesis of specular reflection from temperature-inversion layers at the tropopause."

## RECEPTION

621.396.621:621.396.822 562
**Interference Stability of a Filter Autocorrelator Receiver for Pulse Signals**—V. I. Chaikovski. (*Radiotekhnika, Moscow*, vol. 11, pp. 24–30; April, 1956.) The output signal/noise ratio is calculated for fluctuation-type interference in the input of a correlation-type receiver using a low-pass filter in the averaging circuit.

621.396.621.029.55:621.314.7 563
**Transistor Superregenerative Circuits**—D. F. Page. (*Wireless World*, vol. 62, pp. 606–609; December, 1956.) Two receivers suitable for operation at frequencies of 15–25 mc are described. A surface-barrier-transistor oscillator is used 1) in a self-quenching circuit or 2) in a linear circuit in conjunction with a separate transistor quenching oscillator. Circuit diagrams and a table of typical component values are given.

621.396.621.54 564
**A General-Purpose Communication Receiver**—J. B. Rudd and J. B. Stacy. (Proc. IRE, *Aust.*, vol. 17 pp. 207–217; June, 1956.) Description of a 14-tube superheterodyne receiver designed to meet the specification embodied in the U.K. Merchant Shipping (Radio) Rules, 1952. The rf ranges are 12–60 kc and 100 kc–30mc, and the bandwidth is adjustable between 500 cps and 8 kc in four steps. The turret wave-change unit includes a special mechanism giving a smooth change with positive and precise locking. Additional facilities such as crystal tuning control can be incorporated.

621.396.621.54 565
**A Sideband-Mixing Superheterodyne Receiver**—M. Cohn and W. C. King. (Proc. IRE, vol. 44, pp. 1595–1599; November, 1956.) "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."

621.396.621.54:621.317.4 566
**Transistorized Receiver for Vehicular Radio**—S. Schwartz. (*Electronics*, vol. 29, pp. 217–219; October, 1956.) Description of a multiple-superheterodyne military communications receiver with transistors from the IF stage onwards.

621.396.8:621.376.3 567
**F.M. Multi-path Distortion**—M. G. Scroggie. (*Wireless World*, vol. 62, pp. 578–582, December, 1956.) With a path difference of several miles between two signal components, the comparatively small frequency changes due to fm are sufficient to cause maximum signal fluctuations. These lead to distortions, of magnitude depending on the proportion of the AM which the limiter fails to remove and on the ratio of the signal components. Experiments carried out by the B.B.C. show that as little as 5 per cent signal delayed by 10 miles can cause unpleasant distortion in a typical receiver. The effect of undesired ph.m. is small in the case considered. Remedies suggested are 1) improved amplitude limiting and 2) the use of a direction antenna.

621.396.8.029.62 568
**Polarization Discrimination in V.H.F. Reception**—J. A. Saxton and B. N. Harden. (*Proc. IEE*, Part B, vol. 103, pp. 757–760; November, 1956.) "An account is given of measurements in the band 40–200 mc of the discrimination likely to be achievable between common-frequency transmissions by the use of orthogonal polarizations. It is shown that the discrimination is determined primarily by the topographical nature of the receiving site, that it is substantially independent of distance from the transmitter and of frequency in the band under consideration, and that the median value is about 18 db. The perturbing effects of pick-up on the feeder and of receiving antenna misalignment are discussed.

621.396.822:551.594.6 569
**An Investigation of Atmospheric Radio Noise at Very Low Frequencies**—Horner and Harwood. (See 445.)

621.396.822:621.376.23 570
**An Experimental Study of Intensity Spectra after Half-Wave Rectification of Signals in Noise**—G. E. Fellows and D. Middleton. (*Proc. IEE*, Part C, vol. 103, pp. 243–248; September, 1956.) "Measurement has been made of the spectral intensity of the harmonic

zones which exist at the output of a nonlinear device fed by narrow-band noise and an unmodulated carrier. The spectral shape, the maximum spectral intensity, and the carrier component in each of the first six harmonic zones ($l=0 \cdots 5$) have been determined both theoretically and experimentally for a half-wave $\nu$ *th*-law rectifier with $\nu=\frac{1}{2}$, 1 and 2, for a wide range of input carrier/noise ratios. The measuring equipment is discussed in the paper, the theoretical and experimental results are compared, and a number of computed results of practical interest are presented."

## STATIONS AND COMMUNICATION SYSTEMS

621.376.56:621.396.5 571

**Transmitting System uses Delta Modulation**—R. B. Watson and O. K. Hudson. (*Electronics*, vol. 29, pp. 164–166; October, 1956.) A system suitable for voice communication uses a transmitter in which positive or negative pulses from a generator are applied via one of two gates respectively to a local detector; the stepped output from the detector is compared with the modulating signal, the difference voltage controlling the opening of the gates. A pulse repetition rate of $10^5$/s is adequate. The remote receiver incorporates a detector similar to that in the transmitter.

621.396.3.029.55 572

**Investigation of an Eight-Channel Telegraphy System with Automatic Error Correction**—M. Corsepius, H. Logemann, and K. Vogt. (*Nachrichtentech. Z.*, vol. 9. pp. 306–309; July, 1956.) By combining two sets of T.O.M. (Teletype-on-Multiplex) equipment [see also 2533 of 1956 (Corsepius and Vogt)] working on Fl-Duoplex, an eight-channel system with automatic error correction is obtained. Results of tests in both directions over two s.w. radio links are satisfactory.

621.396.41:621.376.3:621.3.018.78 573

**Distortion in Frequency-Modulation Systems due to Small Sinusoidal Variations of Transmission Characteristics**—R. G. Medhurst and G. F. Small. (PROC. IRE, vol. 44, pp. 1608–1612; November, 1956.) Curves are plotted relating intermodulation distortion in a 600-channel frequency-division transmission system to the amplitude, periodicity, and location of a small sinusoidal variation of either the group-delay or the amplitude characteristic of the system.

621 396.41:621.396.822 574

**Method to Calculate Intermodulation Noise**—A. P. Bolle. (*PTT-Bedrijf*, vol. 7, pp. 51–57; July, 1956. In English.) This method is applicable to multichannel frequency-division am telephony systems and to fm systems based on them. The noise at zero relative level is determined by complex integration using the characteristic functions of the energy spectra.

621.396.5 575

**Single-Sideband Communication Equipment**—J. L. Delvaux and M. Byk. (*Onde Élect.*, vol. 36, pp. 520–531; June, 1956.) Equipment manufactured in France is described. A limit of 50 cps is set for the permissible frequency deviation, corresponding to a stability of $10^{-5}-5\times10^{-6}$ at $5-10$ mc; afc is dispensed with. For distances up to 1000 km a transmitter power of 10 w is sufficient; a 50-w transmitter gives a range of 2000 km.

621.396.65:621.396.11 576

**Tropospheric Scatter Propagation**—Saxton (See 553.)

## SUBSIDIARY APPARATUS

621-526 577

**An Approximate Method for Finding the 'Best Linear Servo Mechanism'**—H. H. Rosenbrock. (*Proc. IEE*, Part C, vol. 103, pp. 260–266; September, 1956.) An elementary graphical method is described for evaluating Wiener's solution for the servomechanism which minimizes the rms error when the system is following a signal affected by noise (*Extrapolation, Interpolation, and Smoothing of Stationary Time Series*, 1950). The minimum phase corresponding to a given attenuation characteristic is determined as an intermediate result of the method.

621.3.013.78 578

**The Magnetic Field Strength in the Corners of Screened Enclosures (Corner Effect)**—H. Kaden. (*Arch. elekt. Übertragung*, vol. 10, pp. 275–282; July, 1956.) The screening effect is reduced in the corners when the external magnetic field is perpendicular to the screen. The explanatory theory is developed and the increase of field strength in the corners is found to be inversely proportional to the 4/3rd power of the distance from the corner. Curves show the direction and increase of the field in the corners. This effect is absent when the external field is parallel to the screen

621.314.634 579

**The Operation of Selenium Rectifiers at Audio Frequencies**—I. G. Nekrashevich. (*Zh. tekh. Fiz.*, vol. 26, pp. 560–567; March, 1956). A detailed report is presented on an experimental investigation. For sinusoidal voltages over a frequency range of 50 cps–20 kc, the average rectified current is practically independent of frequency, provided the working area does not exceed 1 cm.$^2$.

621.319.339 580

**Investigations on Self-Excitation in Band-Type [Van de Graaff] Generators**—W. Herchenbach and H. Sigel. (*Z. Angew. Phys.*, vol. 8, pp. 355–360; July, 1956.) The suitability of various insulating materials is investigated; a new symmetrical circuit arrangement is described.

## TELEVISION AND PHOTOTELEGRAPHY

621.397.5 581

**Pay-As-You-See Television**—(*Engineering, Lond.*, vol. 182, p. 434; October 5, 1956.) Outline of a system developed in the U.S.A., in which six groups of signals are transmitted, viz., the coded picture and sound signals, picture and sound programme-announcing signals, a decoding signal, and a signal controlling the financial transaction. Technique for preventing unauthorized reception is indicated.

621.397.5:535.623 582

**The Amplitude of the Chrominance Carrier in the N.T.S.C. System**—H. Grosskopf. (*Nachrichtentech. Z.*, vol. 9 pp. 289–292; July, 1956.) Calculations and consideration of the color triangle show that for the transmission of normal television pictures the amplitude of the chrominance carrier is likely to be considerably smaller than that required for a bar pattern of pure colors.

621.397.6 583

**Improving the Linearity of Vertical Scanning in Television Equipment**—E. Giua. (*Piccole Note Ist. super. Poste e Telecommunicazioni*, vol. 5, pp. 366–373; May/June, 1956.) A simple cathode-follower circuit is described which adds a parabolic component to the linear sawtooth voltage waveform, thus linearizing the current waveform in the deflection coils in magnetic deflection systems. The arrangement can also be used to improve es deflection.

621.397.6 584

**Slow-Sweep TV for Closed-Circuit Use**—H. E. Ennes. (*Electronics*, vol. 29, pp. 140–143; November, 1956.) A system for transmitting still pictures uses a line scan at mains frequency and a frame frequency of 2 to 7 per sec, in conjunction with a long-persistence-phosphor cr-tube screen. The bandwidth of the picture signal can be accommodated on a program-type telephone line. The camera tube is of vidicon type.

621.397.61.001.4 585

**Proof of Performance for TV Broadcasting**—J. R. Sexton. (*Electronics*, vol. 29, pp. 150–154; November, 1956.) The system of tests imposed by the U.S. FCC before issuing or renewing television broadcasting licenses is described.

621.397.611.2 586

**Ebicon Sees in Dark**—(*Electronics*, vol. 29, pp. 206–210; November, 1956.) Brief description of a camera tube with a Cs-Sb photocathode and a semiconductor target within which avalanche effects are produced by the passage of the high-velocity photoelectrons. The tube may be associated with an image intensifier.

621.397.62:535.623 587

**Three-Phase Detector for Colour-Television Receivers**—A. A. Goldberg. (*Electronics*, vol. 29, pp. 157–159; October, 1956.)

621.397.62:535.623:621.385.832 588

**Flat Tube for Colour TV**—(*Wireless World*, vol. 62, pp. 570–572; December, 1956.) Outline of a lecture to the Television Society by Gabor. The three-beam cr tube is in the shape of a flat glass box which is typically about $4\frac{1}{2}$ inches deep and has a 21 inch-diagonal fluorescent screen. The tube is divided into two halves in depth by a metal plate which carries the whole electron-optical system and serves also as a magnetic screen. The electrons start parallel to the plate from three independently modulated cathodes, pass through an es line-deflection system and then through an em collimator which bends the beam around the edge of the plate. The final bending to cause the beam to strike the fluorescent screen, and the frame scan, are achieved by means of a traveling potential wave. No frame-deflection circuit is required. The shadow mask used for color control is about 0.025 inch from and fixed directly to the fluorescent screen. A cut-away view of the tube and a sketch illustrating the principle of operation of the frame-scanning array are given.

621.397.62:621.397.8 589

**A Possible Method of Improving the Definition of a Television Picture**—V. F. Samoilov and V. M. Rodionov. (*Radiotekhnika, Moscow*, vol. 11, pp. 44–48; April, 1956.) A circuit and method are described for improving the contrast of weak details. The circuit 1) differentiates the signal, 2) delays the signal derivative for a period equal to the duration of one element of the picture, 3) compares the signs of the derivatives of the delayed and direct signals, and 4) in the event that the signs are opposite, increases the amplitude of the signal.

621.397.5 590

**Colour Television Standards [Book Review]**—D. G. Fink (Ed.). McGraw-Hill, London, 1955, 520 pp., 64s (*Nature, Lond.*, vol. 178, pp. 764–765; October 13, 1956.) A selection of papers and records of the U. S. National Television System Committee; a considerable amount of the material has been published previously (see in particular PROC. IRE, vol. 42; January, 1954). While useful primarily as a reference work for engineers, the book also includes a nontechnical description of the N.T.S.C. system.

## TRANSMISSION

621.396.61:621.373.4 591

**An Analysis of Pulse-Synchronized Oscillators**—Salmet. (See 386.)

621.396.61.029.62 592

**An Experimental Assessment of the Linearity of a V.H.F. Transmitter**—D. E. Hampton.

(*Proc. IEE*, Part B, vol. 103, pp. 752–756: November, 1956.) "An experimental procedure is described for testing the assumption that a generator behaves as a linear source. The source admittance is obtained from measurements made when the generator is operating normally, and the problem considered is that of matching this source admittance to the characteristic admittance of the feeder connecting it to a wide-band antenna."

## TUBES AND THERMIONICS

**621.314.632** 593
**Note on Turnover in Germanium Contacts**—P. T. Landsberg. (*Proc. Phys. Soc.*, vol. 69, pp. 763–765; July 1, 1956.) Discussion of the *I/V* characteristics of Ge diodes; Burgess's arguments (994 of 1956) are examined and some alternatives are advanced.

**621.314.634** 594
**An Electron-Diffraction Investigation of the Layer Adjacent to a Cadmium Electrode in a Selenium Rectifier**—V. A. Dorin and D. N. Nasledov. (*Zh. Tekh. Fiz*)., vol. 26, pp. 284–292; February, 1956.)

**621.314.7** 595
**Internal Oscillations in Transistors**—D. Geist. (*Z. Angew. Phys.*, vol. 8, pp. 337–339; July, 1956.) Internally excited oscillations observed in point-contact transistors [see *e.g.*, 2540 and 3392 of 1954 (Hollmann)] and in hook transistors [878 of 1953 (Ebers)] cannot be explained on the basis of stability theory; further experiments are required to clarify the phenomenon.

**621.314.7** 596
**Transit-Time Transistor**—G. Weinreich. (*J. Appl. Phys.*, vol. 27, pp. 1025–1027; September, 1956.) Analysis based on that developed by Mason (374 of 1955) is used to show that an appropriately designed transistor can be operated as a three-terminal amplifier in special frequency bands far above $\alpha$ cutoff. These bands lie one on each side of that corresponding to the negative-resistance condition observed when the device is operated as a diffusion-delay diode [3388 of 1954 (Shockley)]. The existence of the bands is generally independent of the magnitude of the base resistance, but their width decreases with increasing base resistance.

**621.314.7** 597
**A Note on the Extended Theory of the Junction Transistor**—T. Misawa. (*J. Phys. Soc. Japan*, vol. 11, pp. 728–739; July, 1956.) The examination of dc or lf performance at high injection levels may be based on boundary conditions in terms of either es potential or quasi-Fermi levels. At higher frequencies the latter approach is preferable. The author's approach is contrasted with that of Rittner (3390 of 1954).

**621.314.7:546.289** 598
**Current Gain in Formed Point-Contact *n*-Type Germanium Transistors**—D. Haneman. (*Proc. Phys. Soc.*, vol. 69, pp. 712–720; July 1, 1956.) "Experiments on Ge point contact transistors in which collector points of pure donor and acceptor elements are formed show that in general donors appear to be essential to obtain current gain improvements. Successful forming was also achieved with pure gold points. It is assumed that the atoms of the material of the collector diffuse into the germanium during forming. This leads to a change in the amount of lowering of the surface barrier due to the holes injected at the emitter, and accounts for the increased current gain."

**621.314.7:546.289** 599
**Operation of a Crystal Amplifier under Conditions of Carrier Extractions**—V. I. Stafeev V. M. Tuchkevich, and N. S. Yakovchuk. (*Zh. Tekh. Fiz*)., vol. 26, pp. 15–21; January, 1956.) Experiments were conducted with *p-n-p* Ge transistors, based on the extraction of minority carriers as proposed by Angello and Ebert (758 of 1955). The stability of such transistors is good at high temperature; the power amplification is not greatly dependent on temperature and is higher than in amplifiers based on the injection of minority carriers. It is suggested that the principle of extraction can be used in the case of semiconductors with a small energy gap and a high concentration of minority carriers even at room temperature. A combination of extraction and injection enables the dissipated power to be maintained constant, for a given signal level, within a wide range of temperatures.

**621.314.7:546.289** 600
**Relation between Surface and Volume Recombinations in Germanium Triodes with Alloyed *n-p*-Junctions**—A. O. Rzhanov. (*Zh. Tekh. Fiz.*, vol. 26, pp. 239–240; January, 1956.) It has been suggested in the literature that the effect of the volume recombination of the excess charge carriers on the characteristics is negligibly small in comparison with that of the surface recombination. Experiments confirming this are briefly described.

**621.314.7.001.4(083.74)** 601
**IRE Standards on Solid-State Devices: Methods of Testing Transistors, 1956.** (PROC. IRE, vol. 44, pp. 1542–1561; November, 1956.) Standard 56 IRE 28. S2.

**621.383.2** 602
**Photocells and Photomultipliers with Magnesium Photocathodes for detecting Ultraviolet Radiation**—O. P. Dorf, N. G. Kokina, T. M. Lifshits, and D. A. Shklover. (*Radiotekhnika i Elecktronika*, vol. 1, pp. 106–113; January, 1956.) An experimental investigation is reported. The absolute sensitivity and spectral characteristic are not materially affected by the presence of amounts up to a few parts per thousand of Al, Ca, Mn, Zn, Cu, Ni, and Fe. The sensitivity is of the order of 0.32–0.38 $\mu A/\mu w$ at a wavelength of 253.7 m$\mu$; the photomultiplier can detect amounts of power down to $10^{-15}$ w.

**621.383.2** 603
**Antimony-Lithium Photocathode.**—Yu. A. Nemilov and V. E. Privalova. (*Zh. Tekh. Fiz.*, vol. 26, pp. 61–63; January, 1956.) Antimony-lithium photocathodes were prepared having a basic sensitivity of 35–40 $\mu$A/lm for a light-source temperature of 2850°. The spectral sensitivity characteristic indicates that the red boundary of the photoeffect is near 6200 Å. The average value of the thermoelectric work function is 1.5$\pm$0.1 ev. The average value of the activation energy is 1.0$\pm$0.1 ev. The maximum value of the secondary emission coefficient is 15. Heating of the cathode to 100° C decreases the photosensitivity by 50 per cent and increases the work function to 2 ev. The photoconduction processes exhibit appreciable time lag.

**621.383.2** 604
**Investigation of the Phenomenon of Fatigue in Silver-Oxide-Caesium Photocathodes**—P. G. Borzyak, V. F. Bibik, and G. S. Kramarenko. (*Radiotekhnika i Elektronika*, vol. 1, pp. 358–369; March, 1956.) Results of an experimental investigation indicate that the fatigue is due not only to changes in the adsorbed film on the cathode and changes in the work function but also to volume processes in the $Cs_2O$ which take place under the influence of the illumination.

**621.383.2** 605
**The Efficiency of Antimony-Caesium Film-Type Photocathodes**—P. G. Borzyak, B. I. Dyatlovitskaya, and T. N. Chernysheva. (*Radiotekhnika i Elektronika*, vol. 1, pp. 370–376; March, 1956.)

**621.383.4** 606
**The Photo-effect in Lead Sulphide and Related Materials: Part 2**—R. Stein and B. Reuter. (*Z. Naturf.*, vol. 10a, pp. 894–896; November, 1955.) Theoretical interpretation of the observed effects described previously (3249 of 1956).

**621.383.42** 607
**The Influence on the Properties of Selenium Photocells of Near-Infrared Radiation at Low Temperatures**—G. Blet. (*C.R. Acad. Sci., Paris*, vol. 243, pp. 798–800; September 10, 1956.) An account is given of resensitization effects observed when Se photocells whose sensitivity has been reduced by cooling are subjected to infrared radiation of appropriate wavelength.

**621.383.5** 608
**High Time Constants in Se Photocells**—G. Blet. (*J. Phys. Radium*, vol. 17, pp. 430–439; May, 1956.) Observations have been made of various phenomena with time constants $>1$ min, including fatigue effects following strong and weak illumination. A qualitative explanation of the observations is provided by a theory based on the density of free electrons and holes in the barrier layer

**621.385:537.2** 609
**Potential Distribution between Parallel Planes in Electron Tubes**—P. L. Copeland, L. M. Sachs and E. C. Czech. (*J. Appl. Phys.*, vol. 27, pp. 816–819; July, 1956.) Formulas and graphs are presented for calculating potential distributions due to space charge between parallel plane electrodes whose dimensions are large compared with their separation.

**621.385:537.533:621.375.2** 610
**Instability in Hollow and Strip Electron Beams**—C. C. Cutler. (*J. Appl. Phys.*, vol. 27, pp. 1028–1029; September, 1956.) Observations have been made of the breaking up of beams in magnetic fields; the disturbance can be initiated and controlled by applying a space-periodic deflecting voltage. Use of the effect to produce dc amplification is discussed.

**621.385:537.533.087.5** 611
**A Photographic Method of Studying the Spread of Trochoidal Electron Beams**—N. Strandell. (*Kungl. Tek. Högsk. Handl., Stockholm*, no. 106, 13 pp.; 1956.) The vertical and lateral spread of the beam in the trochotron is studied by a photographic method which is applicable at pressures as low as $10^{-5}$ mm Hg.

**621.385:537.533.8** 612
**Influence of Secondary Electron Emission of Insulators on the Stability of the Parameters of Electronic Valves**—N. V. Cherepnin. (*Radiotekhnika i Elektronika*, vol. 1, pp. 38–50; January, 1956.) Experimental results are presented for several types of tube. Steps recommended for effecting improvement include the use of the shortest possible oxide coating on the cathode, increasing the surface resistance, and screening of the supports, use of several types of getters and decreasing the operating voltages.

**621.385:621.317.39.082** 613
**Calculation of Fundamental Parameters for Electronic Valves used for the Measurement of Acceleration**—L. A. Goncharski. (*Radiotekhnika, Moscow*, vol. 11, pp. 49–58; April, 1956.) See also 2076 of 1955.

**621.385.029.6** 614
**The Relativistic Magnetron and the Effective Mass Anisotropy**—L. Gold. (*J. Electronics*, vol. 2, pp. 17–32; July, 1956.) Development of the analysis presented previously (3203 and 3401 of 1954). Effective mass anisotropy arises

as a consequence of the relativistic condition, the magnitude of the anisotropy being governed by the direction and magnitude of the particle injection velocity and the ratio of the electric and magnetic fields. The cutoff relation for the cylindrical magnetron is deduced.

621.385.029.6 615

**Some Properties of Magnetrons Using Spatial-Harmonic Operation**—R. G. Robertshaw and W. E. Willshaw. (*Proc. IEE*, Part C, vol. 103, pp. 297–306; September, 1956.) The design of multicircuit magnetrons for low-power operation at frequencies of the order of 10 kmc with voltages <1kv is difficult because of the close tolerances imposed. A design having advantages from this point of view uses an anode with only a few gaps, *e.g.*, two or four, the electron stream interacting with the Fourier space harmonics of the traveling field. A useful tuning range is achieved by adjustment of an external cavity. Some performance details are given for experimental tubes.

621.385.029.6 616

**On the Possibility of Extending the Concepts of Similarity to Multiresonator Magnetrons with Different Numbers of Resonators**—I. E. Rogovin. (*Radiotekhnika i Elektronika*, vol. 1, pp. 51–70; January, 1956.)

621.385.029.6 617

**Oscillations of the Space-Charge Cloud in a Cylindrical Magnetron**—V. P. Tychinski and Yu. T. Derkach. (*Radiotekhnika i Elektronika*, vol. 1, pp. 233–244, 344–357; February and March, 1956.) Report of a theoretical and experimental investigation. The conclusions drawn from an approximate calculation of the natural oscillation frequencies of the electron cloud in the magnetron, based on consideration of a small perturbation of the static state, are: 1) the oscillation frequencies are principally determined by the characteristics of the interaction space and are approximately subject to the Hartree threshold-voltage formula, and 2) the dispersion of waves in the electron stream leads to limiting of the excited spectrum and deviation from Hartree's formula. The experimental results confirm the theoretical conclusions.

621.385.029.6 618

**Periodic Focusing of Beams from Partially Shielded Cathodes**—K. J. Harker. (IRE Trans, vol. ED-2, pp. 13–19; October, 1955, Abstract, Proc. IRE, vol. 44, p. 274; February, 1956.) Analysis covering both shallow and deep beam scalloping is presented. The relation between the magnetic-field coefficient and the space-charge coefficient is shown in graphs.

621.385.029.6 619

**Space-Charge Waves in Velocity-Modulated Electron Beams**—P. V. Bliokh and Ya. B. Fainberg. (*Zh. Tekh. Fiz.*, vol. 26, pp. 530–535; March, 1956.) Analysis shows that, for a definite law of velocity modulation, waves with exponentially increasing amplitude can be generated. By taking into account, in the hydrodynamic approximation, the thermal motion of electrons it is possible to determine the frequency dependence of the amplification factor and also the upper frequency limit of amplification.

621.385.029.6 620

**An 8-mm Klystron Power Oscillator**—R. L. Bell and M. Hillier. (Proc. IRE, vol. 44, pp. 1155–1159; September, 1956.) A floating-drift-tube single-cavity klystron suitable for use in a low-noise cw radar transmitter is described. The beam current is 0.1 A and the beam voltage 3.5 kv; the output power is 12 w. Aluminosilicate glass was used in the construction, to permit baking out at 700° C. A high-convergence electron-optical system permits use of a sprayed-oxide cathode with 1000-h life.

621.385.029.6 621

**A Five-Cavity X-Band Klystron Amplifier**—G. O. Chalk, B. W. Manley, and V. J. Norris. (*J. Electronics*, vol. 2, pp. 50–64; July, 1956.) Design details and performance figures are presented for a klystron giving an output of several watts with a gain of 70 db. The gain per stage can be predicted with fair accuracy, and the over-all performance can be determined by considering each stage as independent.

621.385.029.6 622

**Generation of Electromagnetic Oscillations by a Travelling-Wave Valve with an External Sectionalized Helix**—V. S. Mikhalevski and D. N. Venerovski. (*Zh. Tekh. Fiz.*, vol. 26, pp. 526–529; March, 1956.) Experiments indicate that by dividing the helix into a number of sections with suitably chosen parameters it is possible to ensure the generation of an oscillation whose frequency remains constant when the anode voltage and the intensity of the magnetic field vary within fairly wide limits.

621.385.029.6 623

**Experimental Investigation of Noise Reduction in Travelling-Wave Tubes**—N. B. Agdur. (*Chalmers tek. Högsk. Handl.*, No. 139, pp. 12; 1954.) The noise power output of the tube is studied as a function of the distance between the anode and the helix input. This measurement gives a determination of the length of the space-charge waves in the drift region, the ratio between the minimum and maximum noise factors, and the distance from the anode to the first noise minimum.

621.385.029.6 624

**The Nature of Power Saturation in Traveling-Wave Tubes**—C. C. Cutler. (*Bell Syst. Tech. J.*, vol. 35, pp. 841–876; July, 1956.) Report of experiments made using a large-scale model of a traveling-wave tube with a helix 10 ft long and 1½ in. in diameter, operated at a frequency of 100 mc with a beam voltage of 400 v. Measurements were made of the efficiency and power output, and of the spent beam velocity and current as a function of rf phase and amplitude. The best value of efficiency, about 38 per cent, is obtained with a value 0.14 for the gain parameter *C*.

621.385.029.6 625

**Nonlinear Wave Propagation in Traveling-Wave Amplifiers**—A. Kiel and P. Parzen. (IRE Trans., vol. ED-2, pp. 26–34; October, 1955.) "A method is given for computing the efficiency of traveling-wave amplifiers with high gain and low *C*, including the effects of space charge and attenuation. The ballistics of the electrons is governed by the Boltzmann transport equation which, together with the circuit equation, is solved in a power series expansion of the input voltage. Only the first two terms of this series are computed and various nonlinear results are given in the form of curves. It appears that for small *C* and small convergence parameters overtaking affects the nonlinear operation slightly."

621.385.029.6 626

**Recent Developments in O-Type Carcinotron Tubes**—P. Palluel. (*Ann. Radioélect.*, vol. 11, pp. 145–164; April, 1956.) See 1904 of 1956 (Palluel and Goldberger).

621.385.029.6:537.533 627

**The Electronic Theory of Tape-Helix Traveling-Wave Structures**—M. Scotto and P. Parzen. (IRE Trans., vol. ED-2, pp. 19–25; October, 1955. Abstract, Proc. IRE, vol. 44 p. 274; February, 1956.)

621.385.029.6:[621.372.2+621.372.8 628

**Electron Waves in Retarding Systems**—Vainshtein (See 338.)

621.385.029.6:621.372.2 629

**Investigation of an Interdigital Delay Line**—F. Paschke. (*Arch. elekt. Übertragung*, vol. 1 pp. 195–206; May, 1956.) The investigation is directed mainly to the possibilities of obtaining a wide tuning range in backward-wave tubes. Analysis is presented for the all-pass line; the first backward-wave harmonic can be largely suppressed by suitable design. Modification of the all-pass line to give a band-pass characteristic is discussed. An oscillator with an f.m. characteristic similar to that of a reflex klystron can be produced. Highpower wide-band amplifiers can also be designed.

621.385.029.6:621.372.2 630

**The Dimensions of Helices in Travelling-Wave Valves**—W. Klein. (*Arch. Elekt. Übertragung*, vol. 10, pp. 261–265; June, 1956.) The importance is examined of the proportions and accuracy of the helices where high-power amplification is to be achieved over a given frequency range without risk of self-excitation and undesired fm. Curves show the relation between tube characteristics and helix parameters. See also 1823 of 1955 (Klein and Friz).

621.385.029.6:621.372.413 631

**A Method of tuning Resonant Cavities**—W. M. Haywood. (*Electronic Eng.*, vol. 28, pp. 395–397; September, 1956.) Tuning arrangements for cavities associated with klystrons are discussed. The effect of enclosing an inductive tuning rod in a fixed glass dome projecting into the cavity, instead of in a bellows, is investigated. The main disadvantage with the glass used was a reduction in cavity shunt impedance of 5–20 per cent; this could be materially reduced by the use of glass with low dielectric loss. The method would not be suitable at high dc power levels, and mechanical difficulties might prevent its use for $\lambda<3$ cm.

621.385.029.6:621.372.413
:621.376.32[:621.318.134 632

**Magnetic Tuning of Resonant Cavities and Wide-Band Frequency Modulation of Klystrons**—G. R. Jones, J. C. Cacheris, and C. A. Morrison. (Proc. IRE, vol. 44, pp. 1421–1438; October, 1956.) A detailed account of work previously reported briefly [3441 of 1955 (Cacheris *et al.*)]. Theoretical and experimental results are presented. Linear frequency deviations up to 240 mc are obtained with only 13 per cent amplitude variation.

621.385.029.6:621.385.2:621.396.822 633

**Monte Carlo Calculation of Noise Near the Potential Minimum of a High-Frequency Diode**—P. K. Tien and J. Moshman. (*J. Appl. Phys.*, vol. 27, pp. 1067–1078; September, 1956). Details are given of a method suitable for carrying out the calculation with a high-speed electronic computer, starting with the generation of random numbers to simulate the electrons emitted by the cathode. Figures were obtained for a total of 3000 unit time intervals. The reduction of noise current due to space-charge smoothing is estimated. Application of the results to microwave beam tubes leads to a minimum-noise-figure/frequency curve exhibiting a sharp minimum at about 2.5 kmc and a flat maximum at about 4 kmc and an approach to the full shot-noise value of 6.3 db at higher frequencies.

621.385.029.6:621.396.822 634

**Klystron Oscillator Noise Theory**—R. L. Bell. (*Brit. J. Appl. Phys.*, vol. 7, pp. 262–266; July, 1956.) The principal sources of undesired am and fm in a floating-drift-tube klystron are shot and partition noise in the beam at the oscillation frequency and its harmonics, except at lowest modulation frequencies where flicker noise predominates in the am. Formulas are presented for the background noise temperature generated by the am and for the fm spectral density.

621.385.029.6:621.396.822 635
**Spurious Modulation in Q-Band Magnetrons**—T. M. Goss and P. A. Lindsay. (PROC. IRE, vol. 44, pp. 1474–1475; October, 1956.) Further evidence is presented confirming the occurrence of the effects reported by Cutler (1256 of 1956).

621.385.029.6:621.396.822:621.372.5 636
**The Equivalent Noise Quadripole of Transit-Time Valves**—H. Bauer and H. Rothe. (*Arch. Elekt. Übertragung*, vol. 10, pp. 283–298; July, 1956.) Application of previous work (380 above) to various types of transit-time tubes. The network parameters are used to analyze the sources of noise, as illustrated by the example of a defective low-noise traveling-wave tube. Fundamental laws of transit-time tubes applicable to the small-signals case are examined.

621.385.032.21:537.533 637
**Field Emission and Field-Emission Cathodes**—D. V. Zernov and M. I. Elinson. (*Radiotekhnika i Elektronika*, vol. 1, pp. 5–22; January, 1956.) A review is presented of theoretical and experimental work. 54 references, including 8 to Russian literature.

621.385.032.213 638
**The Physical Properties of a Porous Metallic-Film Thermionic Cathode: Part 1**—N. D. Morgulis. (*Zh. Tekh. Fiz.*, vol. 26, pp. 536–548; March, 1956.) Various aspects of the operation of a cathode consisting of a w base coated with a Ba monolayer in a state of dynamic equilibrium are considered. Experimental data are given on the formation of Ba in the cathode material, its subsequent diffusion to the surface of the cathode, evaporation from this surface, etc., and on the operation of these cathodes in electronic devices. A brief report is also presented on an electron-diffraction investigation into the structure of the cathode.

621.385.032.213 639
**Bariated Tungsten Cathode: Part 1—General Studies**—Y. Koike and T. Shibata. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, vol. 7, pp. 67–74; September, 1955.) The effect of Si, Zr and WC, used as reducing agents, on the emission of capillary cathodes formed of sintered mixtures of barium silicates and tungsten is described; good results are obtained with Si and Zr.

621.385.032.216 640
**Measurement of the Electrical Conductivity of the Oxide-Coated Cathode**—T. B. Tomlinson and R. E. J. King. (*Brit. J. Appl. Phys.*, vol. 7, pp. 268–269; July, 1956.) Two similar Ni annuli with the oxide sandwiched between them formed the cathode, which was heated by 1-mc eddy currents. Conductivity and emission were measured at temperatures from 300° to >1000°K.

621.385.032.216 641
**Investigation of an Oxide Cathode on a Core of Tungsten-Activated Nickel**—Yu. G. Ptushinski. (*Zh. Tekh. Fiz.*, vol. 26, pp. 232–234; January, 1956.) A detailed report on an experimental investigation of the activating mechanism of *W* in oxide cathodes.

621.385.032.216 642
**The Evaporation of Barium from 'L' Cathodes**—I. Brodie and R. O. Jenkins. (*J. Electronics*, vol. 2, pp. 33–49; July, 1956.) The possibility of reducing the evaporation rate by altering the porosity of the tungsten disk or by varying the Ba compound used to supply the activating Ba has been investigated experimentally. Differing theories regarding the mechanisms of limitation of Ba evaporation are discussed and their range of validity indicated. See also 2918 of 1956.

621.385.032.216 643
**Thermionic Emission from Oxide Cathodes as a Function of Core-Metal-Impurities**—H. Bender. (*Le Vide*, vol. 11, pp. 112–128; May /June, 1956. In French and English.) A review including tables showing the effect of various core impurities on emission.

621.385.032.216 644
**Internally Coated Cathodes**—P. O. Hawkins and J. S. Thorp. (*Nature, Lond.*, vol. 178, pp. 380–381; August 18, 1956.) Measurements were made of the emission from hollow cathodes for different values of temperature electric field and magnetic immersion field. The results confirm the view advanced previously [1260 of 1956 (Bright and Thorp)] that the emission occurs mainly at the edges of the hole. Currents much in excess of the space-charge-limited values may be obtained because of the predominance of the edge effect.

621.385.032.216:621.385.029.6 645
**The Mechanism of Pulse Temperature Rise on the Surface of Thermionic Cathodes**—R. Dehn. (*Brit. J. Appl. Phys.*, vol. 7, pp. 210–214; June, 1956.) Spontaneous fluctuations observed in the cathode surface temperature of pulsed magnetrons are probably caused by resistive heating due to the passage of current through a highly resistive surface layer; the phenomenon probably occurs in other thermionic emitters, particularly of the oxide-coated type and must affect both operation and life.

621.385.032.216:621.396.822 646
**Notes on a Source of Intermittent Noise in Oxide-Cathode Receiving Valves**—M. R. Child. (*Proc. IEE*, Part B, vol. 103, pp. 667–668; September, 1956.) "One form of 'spike' noise in high-grade tubes is shown to be due to intermittent leakage between electrodes caused by the deposition of barium or magnesium, or both on the supporting micas. The magnitude of the effect is shown to increase with the anode voltage between 100 and 300 volts. No effect can be detected below 60 volts in the case of passive cores. The effect may be successfully avoided by the use of mica shields."

621.385.032.24 647
**Study of the Surface of the Grid in a Vacuum Tube**—M. Wada and A. Sato. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, vol. 6, pp. 137–156; 1955.) In an experimental investigation of grid emission due to contamination by cathode material, it was found that at the same working temperature a control grid of gold wire is deactivated, presumably by the diffusion of gold atoms through the contaminated layer, while a molybdenum grid is activated. The results are discussed in relation to factors affecting work function.

621.385.3/.5 648
**Determination of the Thermal Behaviour of the Control Grid of a Valve from its Emission Current Measured by a Pulse Method**—V. Ya. Kunin and M. O. Ratsun. (*Radiotekhnika i Elektronika*, vol. 1, pp. 377–380; March, 1956.) The results obtained in a typical tube by the pulse method described agree with those calculated from an equation given by Wagener (*Z. Tech. Phys.*, vol. 18, pp. 270–280; 1937. Abstract 4112 of 1937.)

621.385.83 649
**Some Problems in the Theory of Focusing of Electron Beams**—P. P. Kas'yankov. (*C.R. Acad. Sci. U.R.S.S.*, vol. 108, pp. 813–816; June 11, 1956. In Russian.)

621.385.832 650
**Limits of Electron-Beam Focusing**—U. Pellegrini. (*Piccole Note Ist. super. Poste e Telecommunicazioni*, vol. 5, pp. 346–360; May /June, 1956.) Factors limiting the reduction of the spot size in cr tubes are examined, and a simple equation for the spot radius is derived, taking account of the spherical aberration of the electron lens. A nomogram is given for evaluating the minimum attainable spread due to space-change effects under any beam conditions. The possibility of compensating spherical aberration by means of space-charge effects is mentioned [See also 1363 of 1956 (Dolder and Klemperer)].

621.385.832:621.397.62:535.623 651
**Flat Tube for Colour TV**—(See 582.)

621.385.832.032.7.002.2 652
**The Choice of a Mould for the Manufacture of Glass Cones for Cathode-Ray Tubes with a Rectangular Screen by the Centrifugal Method**—V. Ya. Savel'ev. (*Zh. Tekh. Fiz.*, vol. 26, pp. 640–645; March, 1956.) Difficulties arise when the centrifugal method is applied to the manufacture of tubes with a rectangular screen, since the liquid glass tends to concentrate at the edges of the pyramid to which the screen is to be sealed. A calculation is made of the shape of a mould which would ensure that glass would reach all points of the wide part of the tube simultaneously. Some photographs of cones for 21-in. tubes produced in this manner are shown.

621.387 653
**Thyratron Behaviour at Low Anode Voltages**—(*Mullard Tech. Commun.*, vol. 2, pp. 242–247; July, 1956.) A discussion of effects including delayed onset of anode conduction and delayed attainment of full conduction, resulting from use of anode voltages as low as or lower than twice the grid voltage.

621.387 654
**Firing Delay and Build-Up Time of the Discharge in the Thyratron**—H. Appel and E. Fünfer. (*Z. Angew. Phys.*, vol. 8, pp. 322–327; July, 1956.) Measurements were made of the relation between the amplitude of the firing pulse and the delay and build-up times. For tetrodes, the delays are of the order of $10^{-7}$s; for triodes, of the order of $10^{-8}$s; they are nearly independent of grid bias. Over the range investigated, the build-up time varies with the pulse amplitude in triodes, but not in tetrodes.

621.387 655
**Confirmation of the Laws developed by Rogowski and Fucks for Firing in Commercial Cold-Cathode Discharge Tubes**—W. Kluge and A. Schulz. (*Z. Angew. Phys.* vol. 8, pp. 328–331; July, 1956.)

621.387:681.142 656
**Gas-Diode Voltage Characteristics**—(*Elect. Rev., Lond.*, vol. 158, pp. 1029–1030; June 15, 1956.) An inexpensive method has been developed at the National Bureau of Standards for equalizing and stabilizing the voltage characteristics of low-cost indicator lamps, such as neon tubes, so as to render them suitable for use as computer elements. The process involves the application of pulses simultaneously to a large number of tubes.

## MISCELLANEOUS

621.3.002.2 657
**Quality Control in Electronics**—M N. Torrey (PROC. IRE vol. 44, pp. 1521–1530; November, 1956.) A review, with 65 references, discussing the objects and techniques of quality control in the manufacture of electronic equipment.