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THE COVER—The vertical and circular arrows, denoting electric and magnetic forces, were borrowed from the official IRE insignia to symbolize the most important IRE event of the year, the National Convention. The complete program, including paper abstracts, starts on page 622, and the list of exhibitors begins on page 103A.

World Radio History

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Scanning the Issue

The Last Days and Ghost of Sputnik I (pp. 610–612)—John Kraus, in a letter to the editor, gives the world its first clear picture of how and when Sputnik met its end. He reveals for the first time that, rather than suddenly disappearing on January 4 as Russia reported, it gradually broke apart into at least eight pieces over a two-week period beginning in late December, with the final fragment coming down shortly after January 9. The unique tracking method, described in another letter, involved picking up WWV signals from Washington after they had been reflected by Sputnik's ionized trails, much like the meteor burst communication techniques described in the December issue of PROCEEDINGS. In yet another letter Harry Wells reports tracking Sputnik's 40 mc signals three different times only to find to his surprise it was actually on the opposite side of the earth. Apparently he was tracking Sputnik's ghost image which formed at a point on this side of the earth diametrically opposite Sputnik.

Electronics: What's Coming After the Missile Age? (Baker, p. 534)—The electronics industry, already a 12 billion dollar a year colossus, has witnessed four major surges in its relatively young life: the radio era, radar era, television era, and now the missile era. Each surge has been governed not only by technological advances but also by the economic climate, military requirements and other factors peculiar to each era. What will cause the next big surge: industrial electronics, says this noted leader of our industry, a broad field which will encompass all electronic fields other than entertainment services and consumer and military goods, and which will include even the field now served by the electrical industry.

Thermoelectric Effects (Jaumot, p. 538)—In this month's invited review paper a noted authority takes up a subject which, because of recent theoretical and experimental advances, is rapidly coming to the fore as one of the most promising frontiers in engineering and science. Indeed, current efforts to develop solid-state materials that combine certain thermal and electrical phenomena could well lead to an industrial revolution as important to our economy as that caused by the transistor. These phenomena concern the reversible interchange of heat and electrical potential energy, and can be made to occur at a junction of two dissimilar conductors (Peltier effect) or in a conductor in which a temperature gradient is present (Thomson effect). Beside electronic refrigerators and air conditioners, a host of other thermoelectric devices have been suggested, among them such glamorous applications as providing power for earth satellites and more mundane uses such as powering lawn mowers. An especially interesting proposal has been the thermoelectric transistor, a batteryless amplifier that would operate entirely independent of conventional sources of power. All readers will welcome this excellent description of the principles and applications of thermoelectricity and discussion of the present state of this important art.

A Communication Technique for Multipath Channels (Price and Green, p. 555)—An important new technique has been developed for reducing the adverse effects of signal fading due to multipath transmission. The system, called Rake, is able to detect separately, and then recombine, each of the many components of the transmitted signal arriving with different delays over various propagation paths. An outstanding feature of the system is that it performs a continuous measurement of the characteristics of the multipath transmission and then uses this additional information to help recombine the components and interpret the signal correctly. This work shows promise of materially improving the performance of existing teletype transmission systems.

Low-Noise Tunable Preamplifiers for Microwave Receivers (Currie and Forster, p. 570)—By means of a specially designed gun the noise figure of a backward-wave amplifier has been reduced to as low as 3.7 db, almost half the theoretically predicted limit. This is the lowest noise figure ever reported for any microwave tube and shows that even lower noise figures are possible, not only for backward-wave amplifiers, but for other types of microwave tubes as well. Perhaps even more important, it makes possible the use of backward-wave amplifiers in microwave receivers, thus creating an important new class of receiving tube which, in addition to the extremely low noise factor, is highly selective and rapidly tunable over a wide frequency range. This tube may lead to the application of the tuned-radio-frequency concept to microwave receivers.

Atmospheric Noise Interference to Short-Wave Broadcasting (Aiya, p. 580)—By comparing time measurements of typical lightning discharges with the response of the human ear to sound impulses, the author develops an interesting method of assessing the radio interference effects of a thunderstorm on a listener of a short-wave broadcast. The results provide criteria for measuring and evaluating this common source of noise, and may eventually lead to improvements in the design of noise limiters in receivers.

Theory of Junction Diode and Junction Transistor Noise (van der Ziel and Becking, p. 589)—During the past two years five papers have appeared in the PROCEEDINGS in which the theory of noise in junction diodes and transistors has been carried forward to the point where the noise behavior of these devices is now reasonably well understood. Still lacking, however, was a proof that was both fully rigorous and not limited in its applicability. This proof is provided by the paper in this issue. It presents a picture of diode and transistor noise which is easy to visualize, simple mathematically, and illuminating to any engineer really wishing to understand this subject.

Ferrite Microwave Detector (Jaffe, *et al.*, p. 594)—The rash of new microwave devices made of ferromagnetic materials is currently undergoing its second major surge in seven years. To the earlier developments, such as isolators and phase shifters, there are now being added in rapid order many new and important devices that depend on previously neglected and more sophisticated effects related to ferromagnetic resonance, among them being frequency doublers, frequency mixers and paramagnetic amplifiers. Another new development, a microwave detector, is now described. While this particular effect has been observed before in ferrites and still requires considerably more development, this paper is noteworthy because it reports on the first practical device which has been built in the U. S. by which microwave detection can be accomplished by ferrites.

The Effects of Short Duration Neutron Radiation on Semiconductor Devices (Behrens and Shaull, p. 601)—Some tests have been made which reveal what effect nuclear radiation has on the electrical characteristics of transistors, semiconductor diodes and solid-electrolyte batteries. The tests are of practical significance because of the increasing use of semiconductor devices in equipment used in conjunction with nuclear reactors. The results will be of considerable interest in that they are among the first data of this type to appear in print.

Optimum Filters with Monotonic Response (Papoulis, p. 606)—A valuable new class of filter has been conceived that combines the most desirable features of two major classes of LC filters, Butterworth and Tchebycheff. The result is a type of filter that is easily realizable, whose amplitude characteristic has no ripple in the pass band, and which has a fast rate of cutoff.

IRE National Convention Program (p. 622)—This feature includes abstracts of some 275 papers to be presented on March 24–27. The 850 exhibitors are listed in the advertising section on page 103A.

Scanning the Transactions ***appears on page 669.***

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Poles and Zeros

International Scene. With the approval of the IRE Board of Directors on January 7, we welcome another outside-the-USA section, that in the Republic of Colombia, headquartering in Bogota. The Editor, having visited that pleasant country in 1954 and having received a warm welcome from many of Colombia's engineers at that time, adds his personal greetings.

This event further emphasizes the international viewpoint which has always been maintained by the Board of Directors, a viewpoint well-justified by the 5500 non-USA members of the IRE. This Institute of Radio Engineers is not a private club for its members, but is an organization dedicated to giving technical service to the electronic profession as a whole. While publications comprise one means by which such service can be provided to the members, a great additional benefit can be derived by dissemination and discussion of information directly with the members, wherever they happen to be. The Section is the geographically distributed means of providing this opportunity.

The Colombian Section now joins Buenos Aires, Rio de Janeiro, Tel Aviv, Cairo, Tokyo, Honolulu, and twelve Canadian Sections in taking electronic knowledge and advancement outside continental USA.

Backwards Space. The letter from Dr. Harry Wells, on page 610 of this issue, discussing his results of tracking Sputnik by radio interferometer raises a question. He found his 40 mc observations several times giving him a fix somewhere overhead when actually Sputnik was diametrically opposite on the other side of the world, thus indicating very interesting ionosphere properties which at those times produced a virtual radio image of the satellite half around the world.

Fully realizing the difference in propagation properties of 40 mc and the radio-astronomy frequencies, it is still possible to wonder if the radio telescopes are always looking at some remote object in space or on its image? If space is curved, do we also have to reckon with it being backwards?

Scanning Lines. The evidently warm reception given our Managing Editor's "Scanning the Issue" page two years ago, has led to suggestions that a similar section for "Scanning the Transactions" would also serve a useful purpose, and the page was introduced in last month's issue of the PROCEEDINGS. This page will survey the most interesting papers as they appear each month in the PG TRANSACTIONS, the IRE STUDENT QUARTERLY, and the National and WESCON CONVENTION RECORDS. It will not attempt to cover all papers so published, since they are already listed in each issue of the PROCEEDINGS. The new page is designed to provide a sketch of what is going on in other areas, and is not necessarily addressed to those who subsequently will order a TRANSACTIONS and read the article.

The page has already served one useful purpose—it made a member of the Board of Directors aware that for years the PROCEEDINGS had been providing a list of current TRANSACTIONS papers and, in most cases, their abstracts. Thus we once more should realize that a Director is really just another guy with too little time today to get his job done!

While on the general subject of time to read the avalanche of literature in our field, it will probably not add to our complacence if we record that in 1957 the IRE published 11,892 pages of editorial material for its members. Such is the explosive field of electronics.

The Convention. When this is read many of us will have completed plans for attendance at our annual New York Convention. Dismayed as a few may be by the stupendous over-all size, we think all should keep in mind the tremendous service given to the profession by this one meeting. Where else can you meet as many leaders in the field (although be unable to locate Joe Doaks or Professor Goodfellow at the Monday evening social), where else can you see the complete products of an expanding industry displayed in one building (although admitting that a slight delay may occur in getting eight feet across an aisle), and where else can you obtain a view ahead to the new things which electronics still has in store (although some of the papers may be called abstruse or even esoteric)?

The IRE did not make The Convention big—the field of electronics is big and this is its convention. With the present breadth of the field, and its future in space exploration, The Convention has to be big to represent the products and applications of the field. One should probably not quarrel with a growing profession which provides us with much more than adequate intellectual challenge. The time to become dissatisfied is when the technological challenge disappears and our job becomes only that of making big things bigger. As long as The Convention remains dynamic and full of youth, so long will our field remain dynamic, interesting and challenging.

More Students. How many of our members having high school or grade school science-oriented progeny have spent some time with their offsprings' textbooks? We suggest such study as a pastime for these long winter evenings, and as a means of finding out what really gives in the schools. As for ourselves, we have found glaring errors of scientific fact and a very considerable amount of poor English, which would certainly have been edited out of these PROCEEDINGS. Our high school senior mentioned that his physics laboratory book was difficult to understand. We agree. If they had removed the words, we might have been able to find the facts.—J.D.R.

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W. R. G. Baker

WINNER OF THE 1958 FOUNDERS AWARD

Walter R. G. Baker was born November 30, 1892 in Lockport, N. Y. He received his bachelor's degree in 1916 and his master's degree in 1918, both from Union College. Later, in 1935, that same institution presented him with an honorary Doctor of Science degree, and in 1951 Syracuse University honored him with a Doctor of Engineering degree.

Dr. Baker's career with the General Electric Company began in 1917 in its Research Laboratory, and culminated in the present Electronics Park at Syracuse, N. Y. His last position with the firm, before his retirement last year, was as vice-president and consultant to C. W. LaPierre, executive vice-president of the Electronic, Atomic and Defense Systems Group. Dr. Baker is now vice-president for Research of Syracuse University, Syracuse, N. Y.

Dr. Baker was chairman of the first and second National Television System Committees, and headed the Radio Technical Planning Board. He is, at present, president of the Electronic Industries Association, which in 1953 voted him its Medal of Honor.

Dr. Baker became an Associate of the IRE in 1919 and a Fellow in 1928. He has been an IRE Director for the past eleven years, President in 1947, and Treasurer since 1951. He has served on many IRE and technical committees, and he is the present chairman of the Professional Groups Committee. He also served as IRE Representative on the ASME Glossary Review Board from 1948 to 1952.

Dr. Baker 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. Dr. Baker, the fourth winner of this award, receives it "for outstanding contributions to the radio engineering profession through wise and courageous leadership in the planning and administration of technical developments which have greatly increased the impact of electronics on the public welfare."

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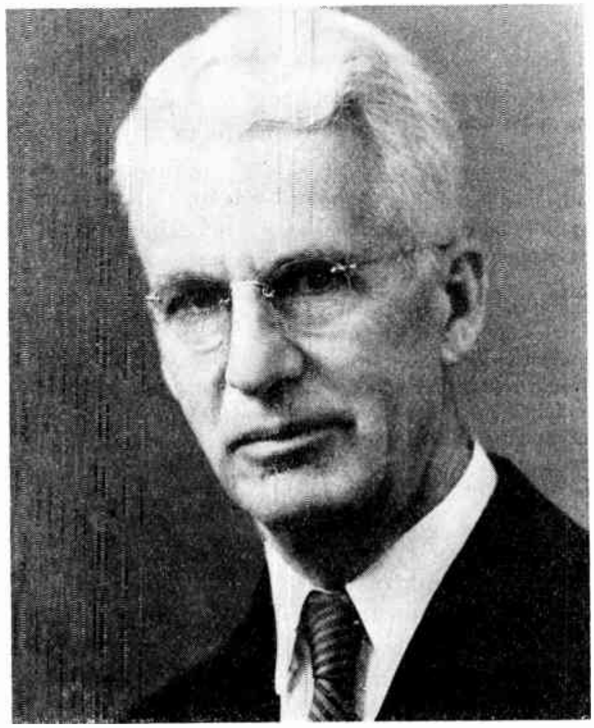

Albert W. Hull

WINNER OF THE 1958 MEDAL OF HONOR

Albert W. Hull was born in Southington, Conn., on April 19, 1880. He received the A.B. degree from Yale University in 1905; in 1909 Yale also awarded him a Ph.D. degree.

At Worcester Polytechnic Institute he was an instructor in physics from 1909 to 1912, and assistant professor from 1912 to 1914.

He began his association with the General Electric Research Laboratory in 1914 as a research physicist. In 1928 he became assistant director of the laboratory, a position from which he directed all radio research of the laboratory during World War II. Following the war he was manager of the General Physics department of the laboratory until his retirement in 1950. Since 1950 he has been a consultant to the Research Laboratory.

Dr. Hull has been credited with creating a greater number of new types of electron tubes than any other man. Among his many important inventions are: the dynatron, the first vacuum tube having negative resistance; the magnetron, precursor of the tube used in World War II radar; the screen-grid tube, the first practical means for efficient amplification of high radio frequencies; the thyratron, now widely used in electronic industrial control equipment; and metal-glass seals which made possible metal receiving tubes.

His work in X-ray crystal analysis won him the Howard N. Potts Medal of the Franklin Institute in 1923. In 1930 he received the Morris Liebmann Memorial Prize of the IRE for his contributions to electron tubes.

Dr. Hull is a fellow and past president of the American Physical Society, and a member of the American Association for the Advancement of Science, National Academy of Sciences, Phi Beta Kappa, and Sigma Xi.

Dr. Hull will receive the Medal of Honor, IRE's highest technical award, at the banquet of the annual IRE National Convention this month for "outstanding scientific achievement and pioneering inventions and development in the field of electron tubes."

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Electronics: What's Coming After the Missile Age?*

W. R. G. BAKER†, FELLOW, IRE

Summary—**The electronics industry has experienced four major periods of development: the radio era, the radar era, the television era, and the missile era. The radio and television eras, both being consumer-goods eras, were strongly influenced by the economic climate and by certian contractual arrangements peculiar to the electronics industry, whereas the radar and missile eras were dominated by factors related to the urgent development of new military weapons.**

Following a discussion of the characteristics of these eras, the author looks into the future to the possibility of the fifth surge in the industry, which he calls the industrial electronics era—a broad term which is intended to include all electronic fields other than consumer goods, military goods, and entertainment services, and including the field now covered by the electrical industry. The paper concludes with a discussion of the major traits and the absence of any one predominating trait in this new era of the future.

Introduction

As ENGINEERS, we are inclined to think of the growth of electronics primarily in terms of a succession of technical discoveries and developments. We think of electronics first as a field of technical endeavor, and quite naturally so.

However, electronics is also a business, a vast twelve billion dollar business. As we shall see in the following discussion, the growth of electronics, whether considered as a business or an engineering art, is strongly influenced by economic and other factors, as well as by purely technical achievements.

Our discussion of the growth of the electronics industry will be centered on the graph in Fig. 1, opposite, which shows the industry volume as a function of time, divided into four major eras.

It indicates the end of the radio era as of 1939 with an annual volume of about one billion dollars. The radar era peaks in 1943 at about five billion dollars of annual business, but by 1946 the annual volume had decreased to just under two billion. The television era is indicated as starting in 1946 and peaks at just under six billion in 1950. Then the missile era is shown as starting in about 1950 with a peak annual volume of just under twelve billion in 1956.

It should be noted that these eras are not too distinctive. The name of the era indicates only the predominant electronics business in that particular period of time. Hence, the total annual volume at any particular point represents the entire business of the electronics industry, which may, of course, be broken down into its component parts.

In addition, in each era there is an overlap representing products of the preceding period and, to some extent, products of the following era. It is interesting to note that in the entire span of forty-one years there have been four major surges in the electronics industry. Two of these have been caused by demand for consumer products, and two have been the results of war or defense military activity.

In the following, we will consider the general characteristics on an industry-wide basis of the two types of eras shown and then guess as to the possibilities of what might produce the fifth era.

Consumer-Goods Eras

Lumping the radio and television ages together, we will notice certain characteristics common to both these consumer-goods eras. These characteristics were partly influenced by the economic climate and partly by certain contractual arrangements rather peculiar to the electronics industry.

In the first place, both radio and television rendered a new entertainment service to the American public, one that they could afford and buy either outright, or with the assistance of liberal consumer credit. Moreover, in both of these periods there was a full play of the free enterprise climate with severe competition present during each era.

Both of these new services produced a great product demand. Since in the early stages of each age the profit margins were rather attractive, there resulted an influx of new companies. Because of the product demand the marketing problem was not too serious in the initial stages, but as saturation began to make itself felt, the selling function became increasingly important.

Another point of similarity was that because each of these consumer eras represented a new technology, the engineer was perhaps of major importance. The next important factor was a manufacturing organization capable of assembling rather complex electronic equipment.

Each consumer era developed a singular economic cycle. This cycle is not peculiar to the electronics industry. It has appeared in the automobile industry and many others. The general characteristics of the economic cycle are an increase in the product supply until it exceeds demand, with the result that the producers reduce the price to the loss point just to stay in business. Coupled with the effect of saturation, this means extreme competition for a shrinking market. In the consumer area, the market may become saturated; that is, the market at a specific price level has been essentially satisfied. Hence, further market expansion is dependent on replacement business and new home formations.

* Original manuscript received by the IRE, October 9, 1957; revised manuscript received, January 2, 1958.
† Vice-President for Research, Syracuse University, Syracuse, N. Y.; formerly, Vice-President, General Electric Co., Syracuse, N. Y.

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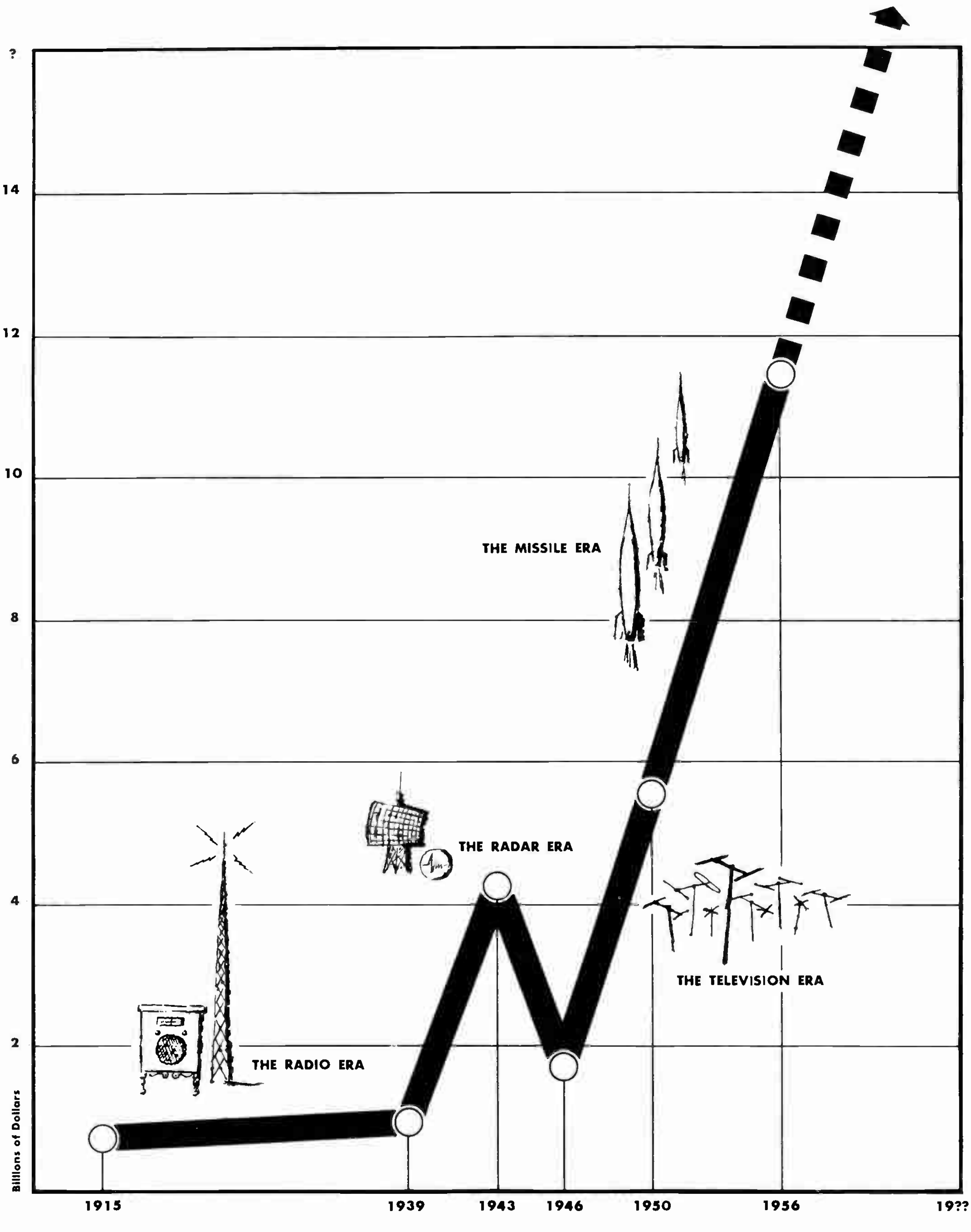

Fig. 1—Growth of the electronics industry.

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In some instances, saturation may also be caused by failure to do a creative job; that is, failure to create new products, failure to develop products for different income levels, or failure to determine how the consumer can be motivated to buy a somewhat modified product. In any event, the end result of this economic cycle is a period of severe attrition and finally a balance between supply and demand through companies which have a profitable business.

One of the results of the economic cycle is a form of standardization. The rate of technological advance slows down and the marketing aspect begins to assume a predominant position. Style becomes an important factor in an attempt to offset the deceleration of the technology, and a more serious effort is made to determine consumer motivation.

The foregoing are not all of the characteristics of the consumer-goods eras but they are representative of the major trends.

Before turning to the characteristics of the eras dominated by military products, let us look a little more closely at the television era. This period is somewhat of a repetition of the radio era. But certain companies had distinctly benefited by the work they had undertaken during the radar age and many had an advantageous position because of facilities put in place during the war.

Three major factors made the television era different from the radio era: first, the continuing influence of military business. Secondly, the economic cycle accelerated more rapidly than during the radio era. This was probably due to the great increase in the technological effort during the war and its accompanying increase in manufacturing facilities. Finally, the tremendous acceptance by the public of television as a new form of entertainment.

The factor that was overlooked by many at the beginning of the television era was the influence of a continuing military load due to the cold war which was accelerated by the conflict in Korea. Somehow, many small companies—and others not so small—lived through the severe dip that the electronics business took in 1946. But after the Korean War, the cold war forced the television era to taper off into the missile era.

Military Eras

Now let us turn to the two eras which were dominated by military products and look at their major characteristics. In the radar and missile ages we find a number of important factors. The design and production of war material was a controlling factor. New weapons and weapons systems required major advances of the technology of both end products and components. As a result, the industry expanded rapidly and many new companies entered the electronics industry with or without government money.

The radar era undoubtedly saved many companies which were having financial difficulties during the closing period of the radio age. It should also be noted that the industry had started work on monochrome television in the period just prior to the radar era and a few sets had been built. As a result, enough was known about television so that during the radar era this technology was either directly or indirectly improved. As for those individuals and companies that had not appreciated previously the great potential of the electronics industry, they were now really alerted.

During the latter part of the radar era, planning started for the television age. Based on past experience, it was expected that the economy would revert to a peacetime level. Although a transition of this sort is not without its problems, it was aided by the fact that certain technological advances in the radar period were directly applicable to television. This was particularly true of basic components such as the picture tube.

The end of the radar era found two types of companies in the industry: those which had been an important factor in the consumer goods industry and had improved their position during the war, and those which had entered the industry through military work. Many small companies that were dependent upon government contracts for their support faced a difficult transition to civilian production at the end of the radar era. Their plight is indicated by the sharp drop in industry volume during this time.

The missile era, which is still in its early stages, is really more than just missiles. But the indications are that it will truly be a missile age in the not too distant future. In its early stages, this era includes not only major advances in aircraft but also further advances in more complex weapons and weapons systems.

One of the early manifestations of the missile era has been a gradual drawing together of the aircraft industry, including propulsion, with the electronics industry. In some instances this has caused the aircraft industry to go into the electronic business. The reverse is not true, except that some large companies are engaged in propulsion as well as electronics.

The missile age has also provided an outlet and hence greater security for the smaller electronics companies that were formed during the war or even after the radar era. One of the interesting by-products of the radar era was the conversion, or perhaps we could call it transformation, of top scientists and engineers into the owners and operators of small companies. Meanwhile, the large companies with strong engineering departments and extensive facilities are especially equipped to assume a major position in this era.

Another characteristic of the missile era has been the increased impact that missiles have had on the weapons system concept.

It is problematical whether the missile era can provide business for those primarily engaged in consumer goods. Perhaps the inability of this era to provide business for these firms will accelerate the attrition that started for them toward the end of the television era. To some extent this trend is now in effect.

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Certainly, the missile era will require a new look at the components section of the electronics industry. Perhaps those companies deeply engaged in missiles will have to take on the major components. If this is true, the growth of the components industry in terms of companies may be retarded. The situation may even result in severe attrition in the components field.

What about the missile era? How long can it last and what is the basic philosophy of this era? Certainly, if we are concerned with only the next five or perhaps ten years, maybe we have no worry. But if we are concerned with a longer period, then we should consider the possibility of the fifth surge in the industry. A straw in the wind, and perhaps there is no straw or no wind, is the recent action of Great Britain.

Britain for certain reasons, perhaps largely economic, arrived at conclusions that the defense of Great Britain is possible only as the collective defense of the free world; that there is no real defense against attack with nuclear weapons; and that the influence of Britain in the world was dependent upon the health of her internal economy and her success in export trade.

Britain has perhaps given up the idea of total preparedness and total security, if such preparedness and security are measured in terms of a large Army, Air Force, and Navy. The doctrine of massive retaliation which has previously been enunciated by both Winston Churchill and Secretary of State Dulles has a great impact on military business in general, and the missile era in particular. Apparently, the British feel that a large ground force is unnecessary in nuclear warfare. Certainly no one can say that ground troops will never be used. The control of small peripheral wars will undoubtedly require troops and the more conventional weapons rather than missiles. But these troops might be provided by the United Nations.

Presumably the day of the battleship in its traditional role is over. The carrier seems to still maintain its position but whether or not it will be eventually replaced by a missile ship is not apparent at this time. It appears that basic changes will come in the treatment of airpower.

What does all this mean? Is it possible that defensive and offensive action will center on about fifteen or twenty different types of missiles launched from stationary and moving platforms? If so, the missile era will truly have arrived.

The Future

Now, what about the future of the consumer goods area? What new consumer goods services are on the horizon?

Excluding modifications of present products by such components as transistors and other technological advances, we have at least two possibilities: color television and subscription television.

With respect to color, there is little doubt that some day it will be a very important service. It may not be in the same demand area as measured by monochrome since it is not in the same class as a new service. Its future will probably be determined by such factors as its technical development, its performance, the economics of the product line, and the program material.

Subscription television is a most interesting modification of an existing service. It seems to the author that its success is dependent upon just one factor. Can subscription television provide program material of such merit that the American public will pay for the program? We will not say that subscription television cannot be made a success. It most certainly will be a major undertaking and will encounter great opposition.

A factor of increasing importance in the fourth and fifth eras will be components. The requirements of these periods will be entirely different than for the entertainment eras.

It is probable that new materials, techniques, and processes may result in an entirely different line of components. Perhaps there will be just as radical a change as is represented by semiconductors. Such a change may offer those companies with adequate research and engineering facilities a new opportunity in the component area.

Unless some new consumer service appears, there probably is not enough demand for the present garden variety of components to sustain all of the existing component companies. Also, there is little incentive for the larger companies to enter the more usual component production. Unless a company can make a real contribution in the component area, it probably would have a difficult competitive position with respect to the companies already entrenched.

The Fifth Era

It seems quite safe to assume that some portion of the electronics volume in the fifth era will be made up of those electronic businesses that were included in the preceding four eras. But will there be some predominating business in the fifth era as there was in the preceding ages?

The author's guess is that the fifth era will have no one predominating trait but will comprise many important product lines. For that reason the economics of this era may be more healthy and vital than the preceding periods. Also, it appears that there will exist a group of product lines which will dominate the era. Based on these product lines the next age might well be called the era of industrial electronics.

We would like to define the scope of industrial electronics since, as the term is used here, it is far more comprehensive than usually interpreted.

By industrial electronics, we mean all electronic equipment or applications other than consumer goods, military goods, and entertainment services. This field would include communications, control, computers—of all types and for all applications, industrial applications of television, the field covered by the electrical

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industry, medical and irradiation applications including radioactive isotopes, and all types of instrumentation.

It is interesting to consider the possible major new components of the electronics industry in the fifth era. We can see even now that the computer, data processing, and related applications will be important. The real possibilities of medical electronics and irradiation are still relatively unexplored, as is the field of ultrasonics.

Because of its history, it seems that many people traditionally look upon electronics as a means of accomplishing new objectives that cannot be attained by any other scientific or engineering means. Electronics tends to be overlooked as a means of doing something better or cheaper or faster, and is considered rather as a way of doing something that is beyond the ability of any other technology. For this reason its infiltration into already highly developed industries has been exceedingly slow.

We have in mind especially the electrical industry. It is here that a number of large electronics companies have almost unlimited possibilities in the fifth era. They have the research, engineering, and manufacturing facilities. But, even more important, they have the marketing facilities, the customer relations, and complete knowledge of the equipment and systems to which electronics should be applied.

This new task is not any easy one. It will require education and selling work by engineers and marketing people in the entire electrical industry area. It will require a continuous survey of not only the technology of the electronics industry but also the electrical industry. It will require the maximum in creative and imaginative ability. It will cost money and it is a real considered risk. But the prize will be well worth the game.

In the last analysis the electronics industry has reached the position it holds today because, above all, it has been forward-looking. It is the presence, or absence, of this most important of all ingredients that will in the end determine the fate and fortunes of the industrial electronics era, just as surely as it has influenced the eras which preceded it.

Thermoelectric Effects*

> 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*

FRANK E. JAUMOT, JR.†

Summary—**This paper is a review of thermoelectric effects in solids, with emphasis on the practical application of these effects. The basic principles of thermoelectricity are reviewed, the present status of the problem and recent achievements are outlined in terms of specific practical applications, and the present status of the more detailed theoretical treatments is discussed in a nonmathematical fashion. The more useful equations describing the important parameters are tabulated in Appendixes I and II.**

I. Introduction

Attainment of a combination of certain, not unreasonable thermal and electrical properties in solid-state materials could lead to an industrial revolution at least as important to our economy as the one now being experienced as a consequence of the discovery of the transistor. This fact has been recognized since before the turn of the century, but it has been only recently that the theoretical and experimental advances have been such as to permit materials to be "tailored" (to some degree) to have certain properties and thus put the actual accomplishment into the practical realm. This approach to the desired aim has had three consequences: 1) There has been a vigorous revival of theoretical work on thermoelectric properties; 2) there has been a considerable increase in the use of thermoelectric measurements as an aid in deducing material properties; and 3) a great number of papers describing thermoelectric effects in metals have appeared. The potential commercial importance of this field has precluded the publication of much of the data obtained on the more interesting semimetal and semiconducting materials. Consequently, the published material does

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not accurately reflect either the quantity or the degree of success (from a practical applications viewpoint) of the work in this field.

The present paper will attempt to review the basic principles of thermoelectric effects, describe the present status of the problem and achievements, and review briefly the more important practical applications of thermoelectric effects. No attempt will be made to review extensively the tremendous body of theoretical and experimental work, other than by listing key references; rather, the present status and the direction in which the work should, and probably will, progress will be outlined. The approach will be entirely qualitative, even descriptive, and the interested reader is urged to consult the references indicated.

II. Background and Basic Principles

Definitions and Concepts

Any phenomenon involving an interchange of heat and electrical potential energy may be called a thermoelectric effect. If the interchange can occur in one sense only, as from electrical energy to heat, but not so as to develop electrical potential energy at the expense of heat, the phenomenon is irreversible. It is reversible if the phenomenon can occur in either sense.

The most familiar irreversible thermoelectric effect is the *Joule effect*, which defines resistance as the property of a conductor that determines conversion of electrical potential energy into heat. The so-called Joule heating is given by I^2R, where I is the electric current and R is the total resistance of the circuit. This phenomenon occurs in one sense only, reversal of the current does not affect it, and by it electrical potential energy cannot be obtained from heat. In practical applications of thermoelectric effects, Joule heating always works against us and with the exception of superconductors, it occurs in all circuits.

The thermoelectric effects more specifically implied by the common use of the term are the reversible phenomena which occur at junctions of dissimilar conductors and throughout regions of conductors in which finite temperature gradients are present. (For a rigorous treatment, irreversible thermodynamics must be used.[1] However, unless otherwise stated, only the reversible portions of the effects will be considered in the general discussion.)

The *Peltier effect* may be defined as the reversible transformation of electrical potential energy and heat at a junction of dissimilar conductors. That is, given a junction of materials A and B, if current flows from A to B there will be a rise in potential across the junction and heat will be absorbed. The rate at which Peltier heat is transferred is given by ΠI, where Π is the Peltier coefficient having dimensions of energy per unit quantity of electricity. The Peltier coefficient depends on the materials of the junction and the temperature of the junction in question; it is independent of other temperatures in the circuit.

The *Thomson effect* may be defined as the reversible transformation of electrical potential energy and heat due to a finite temperature gradient in any single conductor. The rate at which Thomson heat is transferred into a small region of a conductor carrying a current I and supporting a temperature difference dT is equal to $NIdT$, where N is called the Thomson coefficient. The Thomson coefficient depends on the material of the conductor and on the mean temperature of the small region under consideration.

Since the Peltier and Thomson effects are reversible, any junction of dissimilar conductors or any portion of a conductor in which there is a temperature gradient constitutes a source of electromotive force. In a complete circuit, the sum of such emf's is the *Seebeck emf* of the circuit. This sum will be referred to as the thermoelectric force, Θ.

The thermoelectric power, S, which perhaps is the most important quantity involved, is the name given to the rate of change of the thermoelectric force with temperature. That is,

$$S = \frac{\partial \Theta}{\partial T}.$$

There exist well-known thermodynamic relations between the thermoelectric effects. Using subscripts 1 and 2 to denote two different conductors, these may be written as: (see Appendix I for derivation)

$$\Pi_{12} = T\frac{d\Theta_{12}}{dT} = TS_{12} \tag{1}$$

$$N_1 - N_2 = T\frac{d^2\Theta_{12}}{dT^2}. \tag{2}$$

From the derivation, S_{12} can be recognized as the entropy difference per unit charge between phases (conductors) 1 and 2. One may write, $S_{12}=S_1-S_2$, where, now, S_1 and S_2 are called the absolute thermoelectric powers of the phases 1 and 2. Since these are essentially absolute entropies, they are determinable, in principle, to within an additive constant which may be fixed by application of the third law of thermodynamics.

One can use (2) to define the absolute thermoelectric force per degree as

$$\frac{d\Theta}{dT} = \int_0^T \frac{N}{T}\,dT.$$

[1] For the present status of the irreversible thermodynamics of thermoelectricity see, C. A. Domenicali, "Irreversible thermodynamics of thermoelectricity," *Rev. Mod. Phys.*, vol. 26, p. 237, 1954.

——, "Irreversible thermodynamics of thermoelectric effects in inhomogeneous anisotropic media," *Phys. Rev.*, vol. 92, p. 877; 1953.

H. B. Callen, "The application of Onsager's reciprocal relations to thermoelectric, thermomagnetic, and galvanomagnetic effects," *Phys. Rev.*, vol. 73, p. 1349; 1948.

S. R. DeGroot, "Thermodynamics of Irreversible Processes," Interscience Publishers, Inc., New York, N. Y.; 1951.

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In order to obtain $d\Theta/dT$, it is necessary to measure at least one material down to the absolute zero. When this has been done, $d\Theta/dT$ can be found for other materials by measurement of the relative thermoelectric powers or by measurement of the Thomson coefficients and integrating (2). The obvious method of extrapolating the relation between N and T to the absolute zero is useless since N behaves in a very complicated manner at low temperatures. Instead, one takes advantage of the properties of a metal in the superconducting state. It is well known that a superconductor has vanishing electrical resistivity. What is not as well known, is that its thermoelectric power and Thomson coefficient are likewise zero. Hence, by measuring the thermoelectric force of a material against a superconductor one can obtain the absolute thermoelectric force for temperatures below the transition temperatures of the superconductor, and measurement of the Thomson coefficient enables one to obtain $d\Theta/dT$ at higher temperatures.

So far, we have considered only the simplest phenomenological concepts. Any serious pursuit of a thermodynamic treatment should include the irreversible thermodynamics of the situation,[1] which means, of course, increasing the complexity as well as the rigor. On the other hand, the important details are to be obtained from the electronic behavior of the materials involved. Thus, a more profitable approach will come through a detailed electron theory approach, which is quite complex, and no attempt will be made to discuss it here (a qualitative discussion of the general theoretical treatment required is given in Section V). However, it may be worthwhile to consider a general concept somewhat out of context. One can relate the thermopower to the electrochemical potential as

$$S \propto \frac{1}{e}\,\frac{\Delta\mu'}{\Delta T}$$

where e = the charge of the carrier, μ' = the electrochemical potential, and T = the temperature. Now, the electrochemical potential,

$$\mu' = \mu + e\phi$$

includes both the chemical potential (or Fermi level) μ, and the electrical potential, ϕ. The two contributions to the thermopower are often referred to as the static $(\partial\mu/\partial T)$ and the dynamic $(\Delta\phi/\Delta T)$ contributions. While we do know something about both terms in μ', what we know is, as yet, of only moderate usefulness.

In a homogeneous phase, the chemical potential depends only on the temperature and does not depend on whether or not there is a temperature gradient, nor does it depend on the temperature at surrounding points. Thus, $\mu = \mu(T)$. On the other hand, the quantity $(e\Delta\phi/\Delta T)$ at a point depends on the detailed environment at that point. The nature of this (electrical) environment is determined by the distribution of electrons which, in turn, is determined to a great extent by the detailed way in which electrons are scattered by lattice vibrations, impurities, or other imperfections. In metals

$$\frac{\partial\mu}{\partial T} \ll \left| e\,\frac{\Delta\phi}{\Delta T} \right|$$

and one has only to consider scattering effects. In near-insulators

$$\frac{\partial\mu}{\partial T} \gg \left| e\,\frac{\Delta\phi}{\Delta T} \right|,$$

and only the chemical potential is operative; this may also be the case in some semiconductors. In most semiconductors and in semimetals both terms are operative and most certainly interact with each other.

The Figure of Merit

Since the practical application of thermoelectric effects is, to a large extent, a materials problem, it is most desirable to have a figure of merit which characterizes the potential of various solids. It turns out that one has a dimensionless product of materials constants which fulfills this function admirably. Inasmuch as thermoelectric effects can be used for heating, cooling, or generation of electricity, the approach to the derivation of this figure of merit can be made in several ways. However, apart from the probable desirability of different temperature dependencies of the materials constants for different applications, the same analytic expression for the figure of merit obtains from any approach. Here, a refrigerator will be considered.

Consider a closed circuit containing a generator and two branches of dissimilar material. Since we want a figure of merit for a single material, assume one of the two materials is in the superconducting state ($\rho = S = N = 0$) and has such a favorable ratio of length to cross section that any heat conducted along this branch is negligible. The generalization to two materials in the normal state can be made easily when we have learned how the intrinsic properties of a single substance determines the performance of a thermoelectric device.

Let the density of the current in the normal branch be J, and let the two junctions be at temperatures T_0 and T_1, where $T_1 > T_0$. At the junction where the current enters the normal branch at T_0, heat will be absorbed at the rate JST_0 watts per unit area, where S is the absolute thermoelectric power of the normal branch at the temperature T_0. Simultaneously, heat conducted along the normal branch brings $\kappa(dT/dx)$ watts per unit area into the junction, where κ is the thermal conductivity and (dT/dx) is the temperature gradient in the neighborhood of the junction. For simplicity, we will assume this latter quantity to be constant and equal to

$$\frac{\Delta T}{L} = \frac{T_1 - T_0}{L},$$

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where L is the length of the normal branch. The net effective rate of refrigeration per unit area is then,

$$\frac{dQ}{dt} = JST_0 - \kappa\frac{\Delta T}{L}. \tag{3}$$

The power, W, required to drive the current around the circuit is, per unit area, the product of the current density and the sum of the thermoelectric emf's developed and the potential drop caused by Joule losses, hence,

$$\frac{dW}{dt} = J(\Theta + J\bar{\rho}L) \tag{4}$$

where $\bar{\rho}$ is the average resistivity of the normal branch and the other quantities have been defined previously. It is convenient to expand the thermoelectric potential difference in the form

$$\Theta = S\Delta T\left(1 + \frac{N\Delta T}{2ST_0}\right)$$

where N is the Thomson coefficient. The second term may be neglected for small $\Delta T/T_0$.

The coefficient of performance of the refrigerator, C, is the ratio of the quantities (3) and (4) or, after straightforward division,

$$C = \frac{\dot{Q}}{\dot{W}} = C_0\frac{1 - \dfrac{\kappa\Delta T}{JLST_0}}{1 + \dfrac{J\bar{\rho}L}{S\Delta T}} \tag{5}$$

where C_0 is the coefficient of performance of a Carnot refrigerator (equal to $T_0/\Delta T$).

The product JL is the single externally disposable parameter in this expression. Maximizing the coefficient of performance with respect to the parameter gives for the required value

$$JL = \frac{\kappa\Delta T}{T_0S}\left(1 + \sqrt{ + \frac{S^2T_0}{\kappa\rho}}\right). \tag{6}$$

The combination, $S^2T_0/\kappa\rho$, is the dimensionless product of material constants mentioned above; we shall refer to it as the figure of merit, ϑ. Large values of ϑ are desirable since the optimum coefficient of performance is

$$C_{opt} = C_0\left\{\frac{2 + \vartheta - 2\sqrt{1+\vartheta}}{\vartheta}\right\}. \tag{7}$$

Obviously, the above treatment is overly simplified. However, more refined calculations do not change the relative importance of the quantity, ϑ, nor do they result in significant changes in the optimum value of JL. They do result in more accurate values of the optimum coefficient of performance (which in any event is determinable with accuracy only by measurement, in the presence of external factors).

For completeness, it might be mentioned that aside from irreversibility considerations, more accurate treatments require an examination of the temperature distribution within a current-carrying conductor.[2] In the simple one-dimensional case, the rather lengthy calculations involved lead to results which have a simple physical meaning. The Joule heat splits in half, part going to the hot junction, part to the cold junction. Likewise, the Thomson heat is absorbed symmetrically, half from the hot junction, half from the cold. The resistivity is a simple average of its values at the hot and cold junctions.

The extension to two dissimilar materials in the normal state is straightforward. The ratio of the total rate of refrigeration to the corresponding power is

$$C = \frac{\dot{Q}}{\dot{W}} = \frac{\dot{Q}_1A_1 + \dot{Q}_2A_2}{\dot{W}_1A_1 + \dot{W}_2A_2} \tag{8}$$

where A is the area of the branch in question and the subscripts identify the individual branches. The expression is a maximum when $\dot{Q}_1/\dot{W}_1 = \dot{Q}_2/\dot{W}_2$ or when

$$\frac{J_1L_1}{J_2L_2} = \sqrt{\frac{\kappa_1\rho_2}{\kappa_2\rho_1}}. \tag{9}$$

If, for convenience of design, L_1 and L_2 should be equal, then the condition that $I = J_1A_1 = J_2A_2$, where A is the cross-sectional area, determines that

$$\frac{A_2}{A_1} = \sqrt{\frac{\kappa_1\rho_2}{\kappa_2\rho_1}}. \tag{10}$$

Similarly, simple substitution of the proper parameters give, for the figure of merit of a junction of two dissimilar materials,

$$\vartheta_{12} = \frac{(S_1 + S_2)^2T_0}{(\sqrt{\kappa_1\rho_1} + \sqrt{\kappa_2\rho_2})^2}. \tag{11}$$

The expression for C_{opt} is given in Appendix II.

Russian workers (see practically any of the Russian references listed) use a convenient parameter to describe the merit of a material, defined by

$$z = \frac{S^2}{\kappa\rho} = \frac{\vartheta}{T_0} \tag{12}$$

with dimensions of reciprocal temperature.

Anticipating our later discussion, for practical applications, one needs a value of ϑ in excess of unity, or a z of approximately 4×10^{-3} deg^{-1}.

The figure of merit, ϑ, can, of course, be expressed in the language of the electron theory of metals. This has

[2] See, for example, C. A. Domenicali, "Stationary temperature distribution in an electrically heated conductor," *J. Appl. Phys.*, vol. 25, p. 1310; 1954.

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been done by many workers[3,4] (see Appendix II) and, in fact, perhaps the best expression one has for ϑ is the general expression in terms of general transport integrals. On the other hand, said integrals are virtually impossible to evaluate without assumptions which (the writer believes) are to be viewed with considerable suspicion.

Since general expressions which cannot be evaluated are understandably frustrating to the engineer who wants to construct a device, we will content ourselves here with a brief discussion of the important concepts.

The properties of the material (S, ρ, and κ) entering into the expression for ϑ are not independent of each other; in fact, they are all functions of the concentration of free-current carriers, n (see Fig. 1).

The electrical conductivity, $\sigma=1/\rho$, is roughly proportional to n. On the other hand, S tends to zero when n tends to infinity, and tends to infinity when n tends to zero. The thermal conductivity is the sum of two components, the lattice (κ_{ph}) and the electron (κ_{el}) thermal conductivities; the latter is, to a first approximation, proportional to the carrier concentration.

Using the simplest form of the electron theory, the product $S^2\sigma$ (or S^2/ρ) is a maximum at concentrations, n, of the order of 10^{19} cm^{-3}. This is approximtaely 1000 times smaller than n for metals, and even if it is in error by an order of magnitude (which is more than a possibility) one would still conclude that semimetals or semiconductors are the most likely candidates for promising materials. Further, at this concentration of carriers, the electronic contribution to the thermal conductivity is small, which is favorable.[5] These more practical aspects will be discussed in more detail later; here we simply wish to point out that by a suitable selection of electron (or hole) concentration, it is possible to increase the efficiency of thermojunctions by tens of times.

The electron theory also tells us that phonon (lattice wave) scattering and electron scattering are governed by different laws. Thus, it should be possible to increase ϑ by decreasing thermal conductivity without appreciably affecting electrical conductivity (see below).

We already have mentioned that ϑ has various temperature dependencies; this comes about through the carrier mobility and the phonon thermal conductivity. The mobility, u, is defined by the relation $\sigma=nue$, where e is the charge of the carrier. The phonon thermal conductivity has a temperature dependence of $1/T$ or is independent of temperature depending on whether one

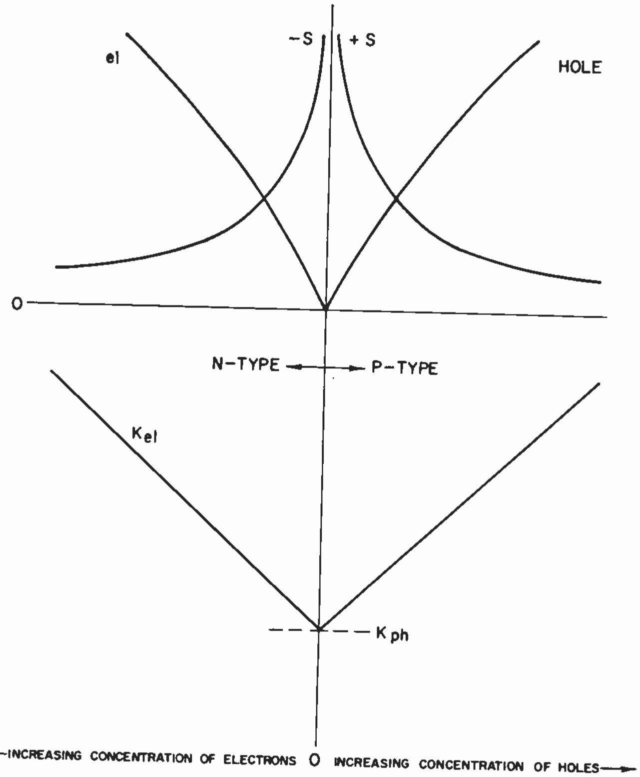

Fig. 1—Approximate dependence of figure of merit parameters on the current carrier concentration.

is considering true phonon scatter or scattering on lattice defects. The mobility has various temperature dependencies. At normal temperatures one generally finds empirically that intermetallic compounds have mobilities which vary (approximately) as either T^{-3} or $T^{-3/2}$; at low temperatures, when scattering on impurity ions predominates, u varies roughly as $T^{3/2}$. Using $z(z \propto T^{3/2}u\kappa_{ph}^{-1})$ rather than ϑ, for convenience, one finds that the various combinations lead to temperature dependencies for this figure of merit of roughly T^4, T^3, T^1, T^0, $T^{-1/2}$, and $T^{-3/2}$. In practice, only $z \propto T$, $z \propto$ constant, and $z \propto T^{-1/2}$ have been observed, to the best of our knowledge. However, for generators, one would like to have $z \propto T^3$, and for refrigerators $z \propto T^{-3/2}$.

Physically, these desired temperature dependencies mean that: 1) for generators, one would like to find a material in which charge carrier scattering is predominantly by impurity ions, and phonon scattering is predominantly by lattice defects; this is a large order; 2) for refrigerators, one would like a material in which the mobility has a T^{-3} temperature dependence, but the phonon scatter is primarily by lattice defects. This is not too improbable, inasmuch as most of the interesting intermetallics do exhibit $u \propto T^{-3}$; however, their phonon thermal conductivity goes as T^{-1}, by and large.

III. The Materials Problem

Objectives

The validity of the figure of merit as an indication of the usefulness of a material in practical applications is

[3] H. J. Goldsmid and R. W. Douglas, "The use of semiconductors in thermoelectric refrigeration," *Brit. J. Appl. Phys.*, vol. 5, p. 386; 1954.

[4] L. S. Stil'bans, E. K. Iordanishvili, and T. S. Stavitskaya, "Thermoelectric cooling," *Izvest. Akad Nauk S.S.S.R., Ser. Fiz.*, vol. 20, p. 81; 1956.

[5] This is strictly true only in the context here. In general, one wants the smallest possible total thermal conductivity, but would like the ratio of electron to lattice component to be as large as possible.

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well established. Thus, a statement of the material objective is obvious. One wants a high thermopower, high-electrical conductivity, and low-thermal conductivity. Inasmuch as the electrical and thermal conductivities are intimately related, this objective becomes a material with maximum ratio of electrical to thermal conductivity and a high thermopower. For the reasons discussed above, these objectives rule out metals (as we usually speak of them) and insulators. Now let us briefly consider the situation existing with respect to each of the parameters involved.

While it is true that one should speak of optimum, rather than maximum, thermopower as an objective, it does not follow that the often stated "approximately 200 μv/degree"[6] is a true optimum value. This figure comes from overly simplified theoretical treatments. Since all of the material parameters contained in the figure of merit are influenced by any factor which affects the electron distribution function and its behavior (*e.g.*, the energy gap between filled and conduction bands; electron-phonon, phonon-phonon, and even electron-electron interactions, lattice defects, etc.), it is only the sum total of these influences and the relative importance of each to the individual parameters which will decide the optimum value of the thermopower.

In the ratio of electrical to thermal conductivities one has really three parameters to worry about because the thermal conductivity is a sum of electron and lattice conductivities. In pure materials and perfectly homogeneous alloys the minimum thermal conductivity is represented by the lattice contribution. (Since this is obtainable only with zero charge carriers, which would mean zero electrical conductivity, one has to have some electron contribution.) Then, in order to have a maximum ratio of σ to κ, one needs relatively few carriers and very high mobilities, plus a low-lattice conductivity. High mobilities are most easily obtained in intermetallic compounds of medium atomic weight elements, indium antimonide being a striking example. On the other hand, minimum lattice thermal conductivities are found in conjunction with low Debye temperatures and highly anharmonic vibrations (usually observed as large thermal expansion coefficients). These conditions are best satisfied by compounds and alloys of the heavy elements, particularly those in the middle groups of the periodic system.

Given a system which has, initially, a favorable ratio of electrical and thermal conductivity there are several things one can do to improve it. First of all, S^2 varies more rapidly with carrier concentration, n, than do the other parameters. Thus, we can "dope" the material to obtain optimum carrier concentrations providing, of course, n was too low initially. Empirically, compensation (*i.e.*, the reduction of effective carriers by additions of impurities leading to the opposite type of conduction) is not, in general, effective; thus, if we want a positive and negative arm of the same material we should ideally start with intrinsic material having very low charge carrier densities. One will find, usually, that doping this intrinsic material has not only increased S but also has increased the ratio σ/κ to some extent. It is then possible to improve the latter ratio by reduction of the lattice thermal conductivity.[7] This may be achieved by introducing into the lattice another substance (either element or compound) which crystallizes in a similar lattice and has approximately the same lattice constant; such a system should exhibit fairly extensive solid solubility. The distortion of the basic lattice by the added impurity is then relatively small and is limited to crystal regions in direct contact with impurity atoms. Such distortions are reasonably effective in scattering thermal oscillations (whose wavelength at ordinary temperatures is of the order of the lattice constant). As a result, lattice thermal conductivity is reduced appreciably, but the current carrier mobility is not affected significantly because lattice periodicity is not greatly affected and, thus the electron waves with their longer wavelengths are not effectively scattered.

It would appear, then, that we have a good empirical guide to the desirable materials. The difficulty is that there are so many possibilities and nearly every possibility involves three or more elements. This leads to so great a number of combinations and variations of combinations that the probability of obtaining a good material by cut and try is quite remote. Thus, more definitive guides would be most useful. At the moment, it would appear that a much better theoretical understanding, coupled with experimental results of basic materials research, offers the best means of obtaining these guides.

Present Status

The desirability of elements of high atomic weight leads one to believe that intermetallic compounds or alloys of Pb, Hg, Bi, Tl, and possibly Sb with Te, Se, and S would make attractive possibilities. All the former elements in combination with Te have been tried (but not in all proportions); only relatively few selenides and sulfides have been investigated to any extent. A tremendous number of other alloys, elemental semiconductors, and compounds have been studied, but, in general, either the thermal conductivities have been too high or the thermopowers too low.

To date, the best materials on which data have been published are Bi_2Te_3[3] and PbTe.[4] The materials reported had ϑ's of 0.173 and 0.32, respectively, at $T_0=300°K$; as refrigerators, the maximum temperature differences obtained were 26°C and 48°C, in each case with a cold junction temperature of −14°C.

[6] A. F. Ioffe, S. V. Airapetyants, A. V. Ioffe, N. V. Kolomoets, and L. S. Stil'bans, "On improving the efficiency of semiconductor thermoelements," *Dokl. Akad. Nauk. S.S.S.R.*, vol. 106, p. 981; 1956.

[7] The author has heard this discussed by many workers in the field. However, it appears to have been published first by Ioffe, *et al.*, *op. cit.*

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Although data have not been published, junctions of p-type Bi_2Te_3 and Bi (similar to Goldsmid and Douglas')[3] have been prepared which produce maximum temperature differences of 40°C. Junctions of p- and n-type Bi_2Te_3 have frequently produced maximum temperature differences of 60°C and the writer is aware of special junctions capable of producing a temperature difference of 75 to 80°C. "Standard" (otherwise undefined) thermoelements producing maximum temperature differences of 55–60°C and individual elements up to 70–80°C have been reported by Russian workers.[8]

For generation of electricity, the optimum efficiency may be expressed as

$$\eta_{opt} = \frac{T_1 - T_0}{T_1} \frac{M - 1}{M + \dfrac{T_0}{T_1}} \tag{13}$$

where

$$M = \sqrt{1 + \frac{\vartheta}{2} \frac{T_1 + T_0}{T_0}}. \tag{14}$$

For $T_1 = 600°C$ and $T_0 = 300°C$, the above p-type $Bi_2Te_3 - Bi$ and n- and p-type Bi_2Te_3 junctions give $\eta_{opt} = 5.5$ per cent and 7.4 per cent, respectively; the others give correspondingly higher values of η.

Some workers,[8–10] find "cascading" (the use of multiple junctions as illustrated schematically in Fig. 2) attractive. For generators this means higher voltages are developed, or, alternately, a reduction in the severity of the mechanical distortion problem. However, with the availability of semiconductor converters, the attainment of desirable higher voltages directly from the thermoelements should not be necessary. Thus, cascading offers its greatest promise in refrigeration. In theory one should be able to achieve any desired temperature difference by using a sufficient number of stages. However, the load capacity of each stage is decreased drastically so that it has not yet proved practical to use more than two stages. (The gain in the coefficient of performance in conversion from a two-stage to a three-stage pile is very small.) On the other hand, the maximum temperature difference in a practical cooling situation (of which the writer is aware) is that reported by Iordanishvili and Stil'bans[9] who achieved a maximum temperature difference of 73° (+26, −47°C) with a three-stage pile. Inevitably, the economics must be considered, and the cost per unit of cooling is always higher when couples are cascaded. This is primarily because the advantage in efficiency offered by two stages is not so great when the actual ΔT is smaller than $(\Delta T)_{max}$. Also, cascading involves the use of more material and fabrication labor.

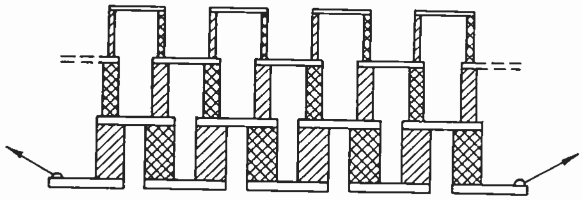

Fig. 2—Schematic diagram of a three-stage thermoelectric cascade.

Paralleling the couples, as opposed to cascading, does not result in a reduction in any phase of the performance and, naturally, is required to increase the capacity (either cooling or generating) of any thermoelectric device.

In any practical junction, there is an optimum current which gives, say, the maximum temperature difference. This current depends on so many extraneous factors, as well as on the physical properties of the materials and the geometry of the junction, that it usually must be obtained empirically. However, it is easy to see from the simple principles given in Section II that such an optimum exists. The Peltier cooling increases with the first power of the current while the Joule heating goes as the square of the current. Thus, one wants that current at which the increase in Joule heating from a further increment of current just balances the increase in the Peltier cooling. Typical curves of ΔT vs I are given in Fig. 3.

At the present time, as can be seen from Fig. 3, thermoelectric devices require very large currents (and very low voltages). Further, for optimum operation, the current must be virtually ripple-free direct current. This means transformers, rectifiers, and filters. It also makes the use of thermoelectric junctions as generators in conjunction with cooling junctions most attractive, since thermoelectric generators supply exactly the type of power required.

Material Fabrication

The fabrication of the materials for the junction requires considerable care. First, one wants homogeneous single crystals, at least for a determination of the potential of the material. Each laboratory has its preferred method of growing single crystals, but for the intermetallic compounds of the heavier elements, the Bridgeman technique appears to be satisfactory. It has two advantages in that it requires a minimum of monitoring and the temperature at which solidification occurs can be controlled quite accurately. In general, one grows the crystals in a vacuum or, if one of the components is significantly more volatile than the other, in an over-pressure of helium or argon (most materials seem to entrap less argon than helium). However, one always has to investigate the crystal for inhomogeneity

[8] E. K. Iordanishvili and L. S. Stil'bans, "Thermoelectric microrefrigerators," *Zhur. Tekh. Fiz.*, vol. 26, p. 945; 1956.
[9] N. E. Lindenblad, U.S. Patent No. 2,734,344.
[10] B. J. O'Brien, C. S. Wallace, and K. Landecker, "Cascading of Peltier couples for thermoelectric cooling," *J. Appl. Phys.*, vol. 27, p. 820; 1956.

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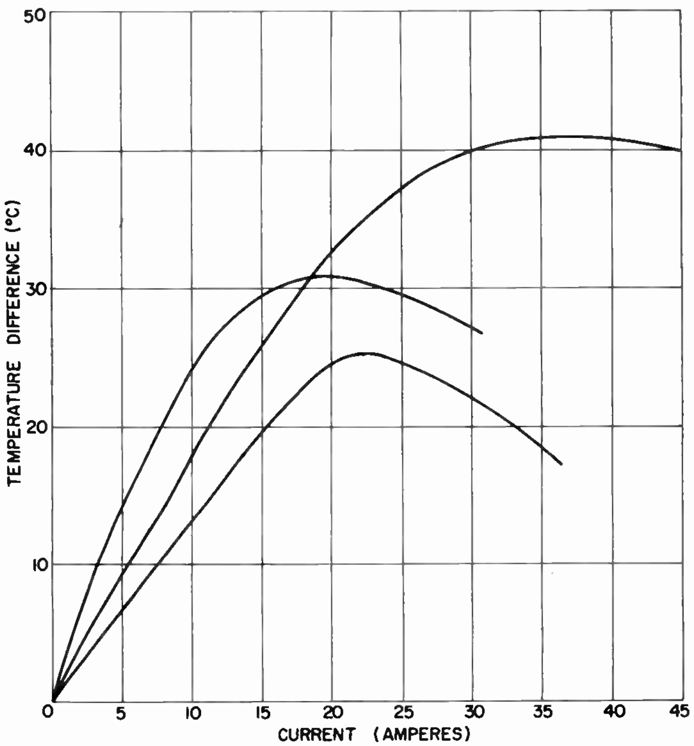

Fig. 3—Typical curves of temperature difference vs current (different junctions). The hot junction temperatures are within about 10°C of ambient.

both radially and axially. It is not too uncommon to find that a crystal exhibits a reasonably large and negative thermopower at one end and a large and positive thermopower at the other.

This latter phenomenon is the immediate result of thermal fluctuations and/or changes in composition of the liquid (due to any one of a number of reasons; *e.g.*, volatilization of one component). However, it is often connected to the two more basic factors that the compound's stoichiometric composition does not have the maximum melting point in the composition range involved, and, in any event, the compound has a finite composition width. These factors increase the difficulty of the doping processes mentioned above, since one would like to be certain of his starting material. For example, if the material present in excess is an electron acceptor, one may not want to dope with a compound having an excess of electron donor material if this were only to lead to compensation and consequently lower electrical conductivity and higher thermal conductivity. (It is emphasized that this statement is not to be construed as saying that we never want to achieve compensation.)

Having grown a good crystal, it must be cut or otherwise formed into the desired shape and size. For this process, each man is on his own; for the most brittle materials, acid cutting may have to be used.

Orientation of the crystal lattice with respect to the direction of heat flow can be most important. To consider a specific example, Bi_2Te_3 has a rhombohedral structure with very marked cleavage planes. While the electrical conductivity and thermoelectric power are virtually independent of lattice orientation (in general, the thermopower is isotropic unless the mean free path of the electron depends on angle as well as energy; *i.e.*, the scattering is anisotropic), the thermal conductivity is a factor of more than three smaller perpendicular to the cleavage planes than parallel to them.[11] This certainly is a situation to be used to advantage.

Once a reasonably good material is obtained it is worthwhile investigating the effects of various annealing procedures and even the oxidation of similar materials, both after forming and during growth of the crystal.

There is a final consideration which has been little investigated. Given a material with a high figure of merit, the possibility of increasing its usefulness by reducing its thermal conductivity by mechanical means should be investigated. That is, if it can be used in powdered, or preferably, in thin sheet form in such a way that the interparticle contacts are intimate enough to permit good electrical conductivity but still retard heat flow, one could improve the figure of merit beyond the basic limit predicted from the material parameters.

Formation of the Junction

In order to use thermoelectric devices in practical hardware, one has to absorb heat at the junction. To do this one wants a relatively large area of material exhibiting good thermal conductivity. Coupled with these factors are the practical aspects of the junction materials. They are relatively expensive, difficult to fabricate, and generally brittle. Thus one wants the actual "legs" of the junction to be as small as possible. Since, for optimum operation, one has to maintain a maximum temperature difference across the legs of the junction, which means a steep temperature gradient, the material of the legs cannot be of too large diameter. Similarly, they cannot be long and thin because of the excessive Joule heating in the relatively high-resistance material. For the better materials to date, this leads to junction legs of two inches or less in length and one-half inch or less in diameter subject, of course, to the condition of (10).

The net result of these considerations is that one end of each of the two legs of the junction is affixed to a large area of a material such as copper or silver. The other end is then provided with some method of cooling (usually fins or water jacket). Two simple methods are illustrated in Fig. 4; an interesting annular construction has been described by Shilliday.[12] The method of joining the junction materials to the copper (for example) is important and provides an area of development in itself.

[11] H. J. Goldsmid, "The thermal conductivity of Bismuth Telluride," *Proc. Phys. Coc. (London)*, *B*, vol. 69, p. 203; 1956.

[12] T. S. Shilliday, "Performance of composite Peltier junctions of Bi_2Te_3," *J. Appl. Phys.*, vol. 28, p. 1035; 1957.

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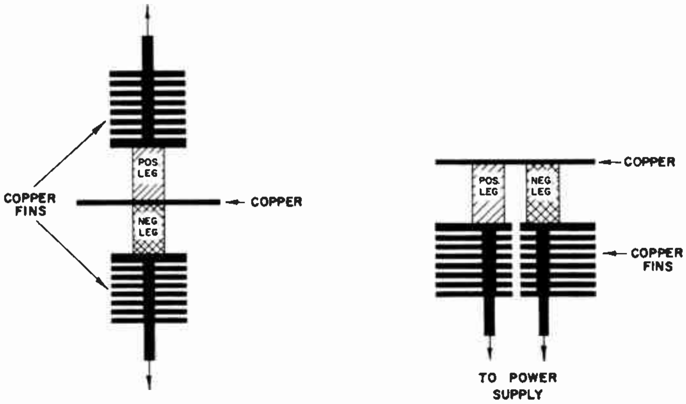

Fig. 4—Two possible methods of forming a junction in such a way as to provide efficient heat absorption and radiation.

Contact resistance must be kept to an absolute minimum since the resulting Joule heating subtracts directly from the operating efficiency. The contact must be intimate and cover the complete area so that small "reverse" junctions are not formed and so the entire area can be used for the desired purpose.

To date, the two most popular methods of forming the junctions have been soldering and diffusion bonding. The latter holds a great deal of promise but requires considerable effort. It requires long annealing times at temperatures near the melting point of the material having the lowest melting point. The results achieved so far have been erratic, presumably due to excessive diffusion distances in the presence of lattice defects and a consequent "smearing" of the junction; also, a good mechanical and electrical bond is difficult to obtain with many materials.

All things considered, soldering has proved the most efficacious method, although it is not easy to obtain optimum results. The solder should be of high conductivity, extremely thin, and must not soften at any temperature in the working range; it also should have a thermal expansion coefficient intermediate between that of the junction material and that of the conductor. Thus, the solder problem also calls for development work and, so far, each experimenter appears to find a different solder best for his purpose. (However, for the lower temperature refrigerating junctions, solders with indium or bismuth as a base have proved quite useful.)

Obviously, all surfaces must be as planar as possible and most workers have found that an etched surface gives better results than a polished or ground surface.

There is some controversy as to any additional treatment after the junction has been formed. Some workers have found that "aging" is desirable; others, that it is undesirable. This aging can take the form of passing excessive currents, either continuous or cyclic, through the junction, or it can be simply a high-temperature anneal. Although the aging may only be a case of improving an unsatisfactory solder job, the writer has been told of situations where current aging has improved the over-all figure of merit of a junction by more than a factor of two. He also has personally experienced ruination of good junctions by high-temperature annealing.

IV. Practical Applications

At the moment, thermoelectric devices are not economically practical, in this country, for the usual consumer appliances. It is estimated that devices with a figure of merit of unity or slightly larger would be feasible for nearly all types of refrigeration and air conditioning. A figure of merit of roughly the same magnitude would make thermoelectric generation practical for consumer use in specialized cases; *e.g.*, aircraft, radio, and television applications. A figure of merit several times larger would be required to make commercial power production competitive. However, there are many applications which could use thermoelectric devices to good advantage at the present time. These can generally be categorized as situations where a large amount of waste heat is present, or where unique performance is more important than cost.

Perhaps the most truly practical application is the thermostating of electronic devices. Since passing a current through the junction in one direction produces cooling and in the other produces heating, all that is required in order to have a controlled temperature chamber is a bimetallic strip operating as a switch to change the direction of current flow. Using present junctions a temperature range of 40°C on either side of ambient is possible, if the load is not too large. Certainly 20 to 25°C on either side of ambient is easily obtained. Also, the response time of a good junction is very rapid; for example, a Bi_2Te_3-Bi couple achieves a temperature change of over 80 per cent of the maximum possible change in twenty seconds or less. Although couples can be cascaded for greater temperature differences (see Section III), the inherent difficulties in fabricating multiple units make this feasible only for small volumes.

Another very real possibility for use of the thermoelectric devices now is the rocket-type airplane, either manned or unmanned. Such planes must have a static generating device, if at all possible, and they have large amounts of waste heat available. At the moment, the amount of power required (10 kw or more) may cause problems of weight; on the other hand, with semiconductor converters the desired voltages are no problem, and thermoelectric junctions are inherently high-current devices. To cite a specific example, a Bi_2Te_3-Bi couple will produce a current of about one-half ampere at several millivolts, *at a temperature difference of 20°C.* Normally, one would expect to use temperature differences of several hundred degrees centigrade if the mechanical distortion produced by such an extreme temperature gradient is not excessive (see Appendix II). Also, it is emphasized that Bi_2Te_3-Bi couples are not the best available; there are couples available which are approximately twice as efficient.

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There are many other more limited applications. Perhaps the most glamorous of these is the use of thermoelectric generators to provide power for space satellites.[13] The use of a reflector aimed constantly at the sun would keep one junction hot, while the end away from the sun would be cold. Such junctions would automatically fill the requirements of a static generator with long life, and able to withstand the launching acceleration.

Use of thermoelectric junctions to measure air humidity eliminates practically all the difficulties encountered in present dew point hygrometers. One version of a humidity device has been described in detail by Kolomoets, *et al.*[14] Thermoelectric effects can also be used for improved zone melting and for the production of ultrasonic vibrations, according to Ioffe.[15]

One cannot dismiss application of thermoelements to consumer appliances too readily. It has been asserted that Russia is producing commercial refrigerators. Recently, a refrigerator was described (including a photograph) which had a 40-litre chamber capacity, kept the chamber at 0°C or below, and used from 55 to 70 watts of power.[16] The statement is made that "Judging by the data published in foreign journals, the performance of the experimental model of the 'electronic' refrigerator developed in the United States and advertised by the Radio Corporation of America is much inferior to that of the *commercial* (italics added) Soviet thermoelectric refrigerator." (On the other hand, RCA's refrigerator included an ice cube freezing fixture.)

Also, RCA recently received considerable publicity as a result of air conditioning a small room using thermoelements only (no moving parts).

It would appear that the first consumer applications would be speciality devices. A particularly attractive one would be portable refrigerators for campers, fishermen, and motorists with small children. The cost of thermoelectric refrigerators, unlike compressor types, is directly proportional to their size. Unfortunately, the retarding factor at present appears to be that the volume of sales involved is not sufficiently attractive to cause the large manufacturers to divert their development effort to the problem, and the smaller manufacturer finds the initial research costs too high.

Certainly, a thermoelement powered electric lawnmower would be a boon to all antinoise groups and suburbia in general.

There are many, many other applications which could be mentioned. For example, we have neglected heating possibilities almost entirely. In moderate climates, where the outside temperature does not go below 30°F, thermoelectric heating with a coefficient of performance of 1.5 could be achieved now; if the minimum temperature were 40°F, the advantage over normal electric heaters would be 2 to 1. For smaller temperature differences the advantages of thermoelectric heating are even more striking; *e.g.*, with the best present junctions, the electric power required to raise the temperature of water by 10° and 5° is, respectively, five and eight times less than that required by conventional electric heaters. The use of thermoelectric devices as heat pumps permitting the utilization of the heat (which is otherwise not required) in the surrounding atmosphere is a fascinating subject for contemplation. However, it must be obvious to the reader that given junctions with a figure of merit of the order of unity, the applications possibilities are restricted only by the limit of the imagination of development engineers.[17] The writer has no doubt that within 5 to 10 years such junctions will be relatively common.

V. Theory

We will now attempt to give a brief, nonmathematical description of the theoretical problem. Although nonmathematical, considerable familiarity with electron theory and quantum mechanics, in general, is assumed. There will be no attempt to make this section tutorial; rather we simply want to point out the direction which we believe really useful theoretical work along electron theory lines must take.

A brief statement of the aim of the desired theory would be a solution of Boltzmann's transport equation after the transport equation itself and its components have been derived, with maximum possible rigor, from first principles. To accomplish this aim requires a detailed knowledge of the behavior of the electrons in the lattice, the lattice vibrations, and their various interactions. Portions of these subjects have been treated by a number of investigators (see references and Bibliography) and, naturally, all treatments have required approximations. The validity and usefulness of the approximations are open to question, so that one important phase of a theoretical treatment should be a critical evaluation of the various assumptions.

One of the many possible procedures for attacking the problem will now be outlined, after which the more important factors will be discussed briefly as separate topics.

The Boltzmann equation in its most general form can be readily transformed into an operator equation by introduction of an appropriate integral operator. A general variational principle which makes no assumptions about the nature of the energy surface or the distribu-

[13] First suggested to the writer by Werner von Braun.

[14] N. V. Kolomoets, M. S. Starnzas, L. S. Stil'bans, and N. P. Fateev, "Measurement of air humidity using semiconductor thermoelements," *Zhur. Tech. Fiz.*, vol. 26, p. 686; 1956.

[15] A. F. Ioffe, "Two new applications of the Peltier effect," *Zhur. Tekh. Fiz.*, vol. 26, p. 478; 1956.

[16] A. F. Ioffe, L. S. Stil'bans, E. Iordanishvili, and N. Fedorovich, "Thermoelectric refrigerator," *Kholodil'naya Tech.*, vol. 33, p. 62; 1956.

[17] An interesting indication of this is the thermoelectric transistor proposed by Kelen and Svedberg which could be a batteryless amplifier entirely independent of conventional sources of electrical energy. *Appl. Sci. Res.*, vol. 6B, p. 369; 1957.

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tion function could be used to give a set of linear equations. Instead of an expansion of the distribution function, f, in the usual manner, a generalized Fermi function or Maxwell-Boltzmann distribution, together with a general Fermi surface would give the most realistic treatment. Obviously, if an expansion of f is not used and the Fermi surface is not assumed spherical, the form of the interaction (integral) operator would not be simple, nor could it be expressed simply as an expansion in the energy. Rather, the expansion of the integral operator would have to depend on the direction as well as the magnitude of the energy surface.

The procedure outlined, using a variational principle, would permit one to obtain the transport quantities in a general form, as ratios of determinants. (It would be necessary to introduce further assumptions only to obtain numerical results.) It should be possible, with the aid of adequate computational facilities, to evaluate the determinants in general without losing the significance of the results by an unwarranted expansion. Alternately, we could treat the integrals as parameters and express the transport quantities in terms of these. The latter method, however, has the major disadvantage that the various integrals are interrelated and cannot be treated as independent parameters; besides, there are so many of them. In any case, the evaluation of the transport quantities would be left until the end of the treatment where it becomes clearer what assumptions are necessary in order to make them tractable.

The Boltzmann Equation

The Boltzmann equation can be written

$$\left(\frac{\partial f}{\partial t}\right)_{\text{collisions}} = -\left(\frac{\partial f}{\partial t}\right)_{\text{external forces}} \tag{15}$$

or

$$\left(\frac{\partial f}{\partial t}\right)_{\text{coll}} = -\frac{2\pi e}{h}\left(\vec{\mathcal{E}} + \frac{1}{c}\vec{v} \times \vec{\mathcal{H}}\right)\cdot \nabla_k f + \vec{v}\cdot\nabla_r f \tag{16}$$

where f = the carrier distribution function, e = the charge on the electron, h = Planck's constant, $\mathcal{E}$ = the electric field, c = the velocity of light, v = the velocity of the electron, $\mathcal{H}$ = the magnetic field, and k denotes the wave vector.

The left-hand side of the equation (and more generally, the kernel in the integral equations involved in the operational form of the equation) depends on the interaction of the electrons with the lattice vibrations, the lattice distribution function, and probably on electron-electron interactions.

The reason for considering a variational principle as a solution is that the Boltzmann equation is solved to a fair approximation by expanding the distribution function in an infinite series in powers of the energy; this suggests a close connection to the Ritz variational method. Such a treatment has already been developed by Enskog[18] for classical statistics and is easily generalized for Fermi statistics. Also, the operator form of the equation retains the symmetric form of the equations necessary for the Kelvin relations.

Rather than deal separately with the familiar electron-lattice and electron-electron interactions, we will include a few brief statements here.

It has been customary to neglect the interaction between the electrons through their coulomb field. Admittedly they probably are not too important, but they do mix different energies without randomizing velocities and thus produce effects which make some simple treatments impossible.[19] This indicates they cannot be completely neglected. Fortunately, Landau[20] has shown that these interactions can be taken into account quite easily. Also, as mentioned below, their effect may be taken into account by a single unified treatment.

In order to calculate the probability of transitions between energy states as a result of electron-lattice interactions, one has to set up a wave function and calculate the matrix elements which describe the transition probabilities. The most popular method seems to be to calculate the transition probabilities by ordinary time dependent perturbation theory, after the matrix elements are obtained, by one of the following means: 1) the method of deformable potential,[21] 2) the method of distorted crystal,[22] and 3) the method of motion of the ions.[23] The objection to these methods (aside from the obvious simplifications of a spherical energy surface, the smallness of the change in the magnitude of the wave vector, etc.) lies in the perturbation method itself. Thus, at present, it would appear that the possibility of a totally different approach offered by the field theoretical[23–27] and collective electron[28,29] descriptions should

[18] D. Enskog, Ph.D. Dissertation, Upsala College, East Orange, N. J., 1917.

[19] C. Herring, "Transport and deformation-potential theory for many-valley semiconductors with anisotropic scattering," *Phys. Rev.*, vol. 101, p. 944; 1956.

[20] L. Landau, "Kinetic equation for the Coulomb effect," *Physik. Z. Sowjetunion*, vol. 10, p. 154; 1936.

[21] A. Sommerfeld and H. A. Bethe, "Handbuch der Physik," vol. XXIV/II, Springer-Verlag, Berlin, Germany, ch. 2; 1933. F. Bloch, *Z. Physik.*, vol. 52, p. 555; 1938.

[22] E. L. Peterson and L. W. Nordheim, "Resistance of univalent metals," *Phys. Rev.*, vol. 51, p. 355; 1937.

[23] J. Bardeen, "Conductivity of univalent metals," *Phys. Rev.*, vol. 52, p. 688; 1937.

[24] T. Toya, *Busseiron Kenkyu*, vol. 59, p. 179; 1952.

[25] D. Bohm and T. Stover, "Application of collective treatment of electron and ion vibrations to theories of conductivity and superconductivity," *Phys. Rev.*, vol. 84, p. 836; 1952. Given in more detail in T. Stover's unpublished Ph.D. Dissertation, Princeton University, 1952.

[26] S. Nakajima, *Proc. Internatl. Conf., Theo. Phys.*, Tokyo, Japan, p. 916; 1953.

[27] H. Froehlich, "Interaction of electrons with lattice vibrations," *Proc. Roy. Soc. A*, vol. 215, p. 291; 1952.

[28] D. Bohm and D. Pines, "A collective description of electron interactions, I. Magnetic interactions," *Phys. Rev.*, vol. 82, p. 625; 1951. "II. Collective versus individual particle aspects of the interactions," vol. 85, p. 338; 1952. "III. Coulomb interactions in a degenerate electron gas," vol. 92, p. 609; 1953. "IV. Electron interaction in metals," vol. 92, p. 626; 1953.

[29] J. Bardeen and D. Pines, "Electron-phonon interaction in metals," *Phys. Rev.*, vol. 99, p. 1140; 1955.

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be exploited. A choice between these two appears to favor the latter which takes Umklapprocesses into account and also takes Coulomb interactions between the electrons into account to an extent which may be sufficient, without explicitly introducing them into the theory. We emphasize again, however, that the nature of the interaction may remain unspecified in the formal treatment, becoming necessary only when integrals have to be evaluated for numerical results.

The Lattice Distribution Function

The usual electron theory treatment does not consider the effect on the electrical and thermal properties of the departure of the lattice-distribution function from its equilibrium value,[30] although it has been recognized for some time[21] that a nonequilibrium distribution function will lead to a lattice heat current. Makinson[31] has obtained an expression for the lattice distribution function for metals; this was extended to the case of semiconductors by Haar and Neaves.[32]

When including these effects, one must remember that the total time rate of change of the distribution function due to the scattering of the lattice waves is due to: 1) interaction with conduction electrons, 2) irregularities in the lattice of atomic dimensions, 3) microscopic lattice defects (grain boundaries, external boundaries, etc.), and 4) anharmonic terms in the equations of motion of the lattice.

Also, in the simple model on which most conduction theory is based, it is assumed that transverse waves do not interact with the electrons. It is probably more accurate to assume that they interact to about the same extent as the longitudinal modes.

Processes 1) and 4) have been taken into account in an expression for conductivity derived by Bethe,[21] and 4) has been treated in detail by Peierls[33] and Blackman.[34]

The Distribution Function

As its name implies, this is simply a function which describes the distribution of the electrons in the various energy states, or more elegantly, gives the probability of occupation of an energy state. The distribution of most interest here is given, in form, by the Fermi function,

$$f_0 = [\exp \beta_0(\epsilon - \mu_0) + 1]^{-1},$$

for the equilibrium situation. Here $\beta_0 = 1/kT$, ϵ = the energy of the electron, μ_0 = the Fermi energy, k = Boltzmann's constant, and T = the absolute temperature.

Generally, the nonequilibrium function, f_s, is expressed simply as an expansion (often in terms of Legendre polynomials). However, even the more complex expansions end with the result that the influence of an external electric field is to translate the Fermi sphere without any deformation. This cannot be wholly correct, but for finite temperatures the change does occur gradually, within a thickness of approximately kT.

It is not difficult to define a surface $\mathcal{S}$ which demands only that $f_s = \frac{1}{2}$ on $\mathcal{S}$ and, in fact, as long as μ (not μ_0) is a slowly varying function of the polar coordinates ϑ, ϕ, $\mathcal{S}$ is more or less a spherical surface. In order to take into account the angular and temperature dependence, Mühlschlegel[35] suggests using

$$f_s = [\exp \beta(\epsilon - \mu) + 1]^{-1},$$

where now β and μ are functions of ϑ, φ, T, ϵ, and ΔT, and must be determined from the Boltzmann equation.

For simplicity in calculation, the number of electrons per unit volume with a given wave number can be expressed as a sum of two terms. The first can be chosen so as to determine the shape of the surface, leaving the second influencing only the shape of the transition layer.

Thermomagnetic Effects

For the most part, we have, so far, ignored the effects of a magnetic field. This is probably justified since experimentation shows that only very large magnetic fields have a measurable effect on the thermoelectric properties of the more interesting materials. However, in a general study of transport phenomena they must be considered.

When a magnetic field is present along with electric and thermal currents, the resulting phenomena are extremely complex even in isotropic media. Here, we can classify the effects in terms of the relative directions of the electric and magnetic field.

First, we have the transverse effects in a transverse magnetic field where the primary current and the electric field and/or temperature gradient are perpendicular to one another and the magnetic field. By substituting the proper magnitudes of the individual components of the various field and current vectors into the general solution for this case, one defines (among many other possibilities) the Hall coefficient, the Ettinghouse coefficient, the Ettinghouse-Nernst coefficient, and the Righi-Leduc coefficient.

Second, we have longitudinal effects in a transverse magnetic field, where the primary current and electric field are parallel to each other but perpendicular to the magnetic field.

Third, we have longitudinal effects in a longitudinal magnetic field where all the fields are parallel.

The latter two groups (longitudinal effects) include the electrical and thermal magneto-resistance and the

[30] A. H. Wilson, "Second-order electrical effects in metals," *Proc. Cambridge Phil. Soc.*, vol. 33, p. 371; 1937.
[31] R. E. B. Makinson, "The thermal conductivity of metals," *Proc. Cambridge Phil. Soc.*, vol. 34, p. 474; 1938.
[32] D. ter Harr and A. Neaves, "On the thermoelectric power of metals," *Proc. Roy. Soc. A*, vol. 228, p. 568; 1955.
[33] R. Peierls, "Kinetic theory of heat conductivity in crystals," *Ann. Physik.*, vol. 3, p. 1055; 1929.
Ann. Inst., H. Poincaré, vol. 5, p. 177; 1933.
[34] H. Blackman, "Heat conductivity of simple cubical crystals," *Phil. Mag.*, vol. 19, p. 989; 1935.
[35] B. Mühlschlegel, "Contributions to the conductivity of metals at low temperatures," *Ann. Physik*, vol. 17, p. 199; 1956.

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influence of a magnetic field on the thermoelectric power. The latter, of course, is the effect of interest here, but its complexity is such that the writer is willing to abide by experimental results and neglect it.

The Time of Relaxation

This is perhaps the most frustrating quantity in electron theory. If we could define a general relaxation time, τ, we could replace the right-hand side of (15) or (16) with

$$\frac{f_s - f_0}{\tau}.$$

It is, of course, always possible to define a relaxation time in terms of the electrical conductivity or the thermal conductivity but, in general, these relations give different values for τ. Thus, a universal definition of τ, applying to all conduction phenomena, is not possible. Consequently, it is necessary to solve the full integral equations for the distribution function directly, and to deduce the conduction parameters from the solution. After this has been done, one may still define a relaxation time only for electrical conduction or thermal conduction, etc., and the term loses its usefulness.

The difficulty with this parameter is easier to comprehend if we consider the mean free path $l[l \propto (1/\tau)]$. At moderate and high temperatures, l is independent of the energy, ϵ, if the electrons are scattered principally by the acoustical modes of the lattice (phonon-electron scattering). However, optical modes may scatter electrons, particularly in compounds, and in this case l is proportional to ϵ. If ionized impurities are the dominant mechanism, l is proportional to ϵ^2. If more than one mechanism operates, the appropriate combinations must be used.

Further, one must remember that a phonon as well as an electron mean free path exists. The meager experimental data available indicates that, at moderate to high temperatures, both have the same temperature dependence. If this be universal, one can write $l_{ph} = al_{el}$, and the equations will involve mostly the ratio, a. This does not ease the situation too much, however, since a varies over a factor of four or five.

The most immediate interest in the definition of a mean free path lies in its connection with the computation of the dependence of the figure of merit on the Fermi energy (or, more accurately, on the degeneracy parameter, μ/kT). If one takes reasonable values of a, the velocity of sound, and the effective electron mass (which quantities may vary by factors of 5, 3, and 10 or more, respectively) and further assumes that at any given temperature only one scattering mechanism predominates (so that $l = l_s\epsilon^s$), one can estimate the desired dependency. It is found that for $s=0$, ϑ has a strong maximum in the neighborhood of zero degeneracy. The value of ϑ_{max} increases and its position moves to more negative values of μ/kT as the effectiveness of the lattice thermal conductivity decreases. For other energy dependencies (other values of s) one finds a similar situation; that is, ϑ_{max} increases and moves to more negative values of μ/kT, the higher the value of s. From this, a brave man might conclude that the theoretical treatment agrees with the more empirical discussion above, in predicting the desirability of minimum lattice thermal conductivity and domination of impurity scattering. The opinion here is that a great deal more theoretical work is needed.

The Assumptions

The Boltzmann equation is a complicated integro-differential equation and, quite understandably, various assumptions have been introduced in previous treatments to facilitate solution. Unfortunately, not only are these assumptions not evaluated critically, often they are not even stated explicitly. We list here the more important ones and discuss their merits briefly.

1) Squares and products of the electric field and temperature gradients may be neglected. This is probably justified on the grounds that no quadratic effects are observed experimentally.

2) A relaxation time or a mean free path may be defined. This has been discussed above.

3) The quantity $[(f_s - f_0)/f]$ is much smaller than one, in the presence of external fields and/or temperature gradients. This simply is not proven. However, this assumption combined with assumption 2) suffice to prove that we may use the Lorentz expression[36] for the distribution function, and this is one of the main objections to present theory as discussed above.

4) The influence of the lattice may be neglected, apart from the fact that it provides a mechanism by which a finite relaxation time may be realized. This is discussed in several places above.

5) One may use the one electron approximation; that is, treat the carriers as if they were independent. Russian authors have tried to eliminate this assumption but, so far, their theories do not lend themselves to numerical predictions (see Anselm[37] and Wonssowski[38] for a discussion of these theories). However, it appears that this assumption does not appreciably affect the final results (on the other hand, see Bontsch-Brujewitsch[39] and Pines[40]). One might expect the one electron approximation to lead to good approximations since, in metals, the Pauli principle will restrict interactions between conduction electrons; while in semiconductors the carrier density is small enough to make two carrier processes negligible.

[36] H. A. Lorentz, *Arch. Neerl. Sci.*, vol. 10, p. 336; 1905.
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[37] A. I. Anselm, "Some problems of the electron theory of crystals," *J. Tech. Phys.*, U.S.S.R., vol. 21, p. 489; 1951.
[38] S. W. Wonssowski, *Fortschr. Phys.*, vol. 1, p. 239; 1954.
[39] W. L. Bontsch-Brujewitsch, *Fortschr. Phys.*, vol. 3, p. 408; 1955.
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6) The Coulomb interactions due to electron-electron collisions may be neglected. This was discussed above; the necessity for this assumption may be removed by use of a collective electron approach.

7) The equilibrium distribution function, f_s, depends on the carrier wave vector k, only through the energy, ϵ. This was discussed above.

8) The energy surfaces are spherical. This was discussed above. It is surprising that no attempts seem to have been made to introduce the vast results from band theory into a systematic discussion of transport phenomena, apart from the use of an effective mass[41–43] and the two-band approximation.[44–45] Recent experimental work[46,47] has shown the necessity for considering warped energy surfaces, and the so-called many-valley model for semiconductors has been relatively extensively discussed.[48–50]

Although we still believe the assumptions should be made as late in the calculations as possible, it appears impossible to remove all of them. Certainly the very method of attack depends on the assumptions made. With this in mind, we can only hope that every attempt will be made to keep the approximations consistent. That it is not easy to accomplish this was illustrated by Sondheimer[51] who showed that the work on thermoelectric power in metals by Haar and Neaves[32] suffered from inconsistency in the approximations used.

Appendix I

Derivation of the Kelvin Relations

The following derivation is the simplest possible. It is based on the "quasi-thermodynamic" treatment which is open to serious objection, but which gives the relations correctly.

Consider the circuit of Fig. 5, with two conducting

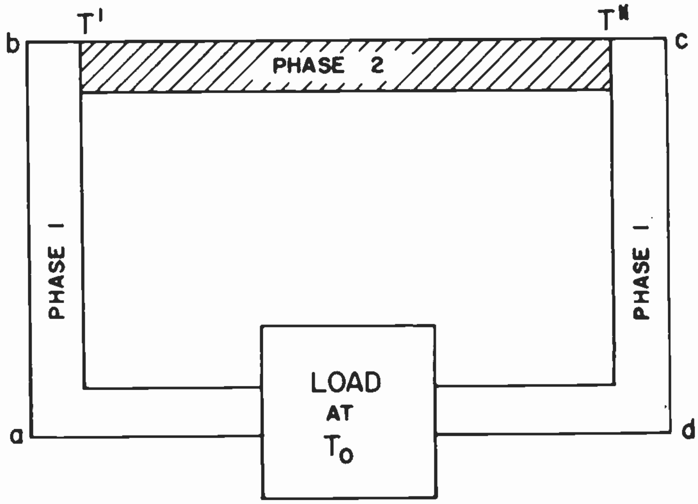

Fig. 5—Schematic thermoelectric circuit.

phases connected in series with a load at constant temperature T_0. The load is a reversible source of emf adjusted to be always almost in equilibrium with any emf's in the rest of the circuit. It is the rest of the circuit which forms the thermodynamic system of interest here.

We now let a charge δq pass around the circuit in the direction a, b, c, d and return to its original condition. Then, the first law of thermodynamics may be applied in the form

$$\delta U = 0 = \delta Q - \delta W \tag{17}$$

where δQ is the heat absorbed by the charge and δW is the work done by the charge on the load. This latter term is simply $\Theta\delta q$, where Θ is any emf developed in the circuit. To obtain δQ, we follow the charge around the circuit. In phase 1 (a to b) the temperature of the charge is raised from T_0 to T'. Unless the charge carriers have zero specific heat, this increase in temperature requires an absorption of heat,

$$\delta q \int_{T_0}^{T'} N_1 dT$$

where N is the specific heat per unit charge or the Thomson coefficient of charge carriers in phase 1.

The charge must now transfer from phase 1 to phase 2 and if the charge carriers in these phases have different entropies, a latent heat appears which is given by

$$\Pi_{12}'\delta q$$

where Π_{12}' is the latent heat per unit charge between phases 1 and 2 at temperature T' (otherwise the Peltier coefficient). Completing the circuit gives two more integrals over the specific heat and another latent heat at the junction between phases 2 and 1 at temperature T'' (obviously $\Pi_{21}'' = -\Pi_{12}''$). Collecting these terms and factoring out the common δq in (17) we have

$$\Theta = \Pi_{12}' - \Pi_{21}'' + \int_{T'}^{T''} (N_2 - N_1) dT. \tag{18}$$

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[51] E. H. Sondheimer, "The Kelvin relations in thermo-electricity," *Proc. Roy. Soc. A*, vol. 234, p. 391; 1956.

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To this point, there is no reason to question the validity of (18) or the means by which it was obtained. However, the next step, which is the application of the second law, cannot be justified within the framework of classical thermodynamics. We postulate that the irreversible production of entropy resulting from the heat flow in the two conductors may be ignored and the second law applied to the reversible part of the process. (The same results can be obtained using a rigorous treatment, but not quite so simply.)

Applying the second law in the form

$$\delta S = 0 = \frac{\Pi_{12}'}{T'} + \frac{\Pi_{12}''}{T''} + \int_{T'}^{T''} \frac{N_2 - N_1}{T} dT \qquad (19)$$

gives us our second equation. From (18) and (19) the relations between Θ, Π, and N may now be obtained. First, we differentiate both equations with respect to T'' and set T' equal to T to obtain

$$\frac{d\Theta_{12}}{dT} = N_2 - N_1 + \frac{d\Pi_{12}}{dT} \qquad (20)$$

$$0 = \frac{N_2 - N_1}{T} + \frac{d}{dT}\left(\frac{\Pi_{21}}{T}\right). \qquad (21)$$

Eliminating $N_2 - N_1$ from these equations we obtain the first Kelvin relation,

$$T \frac{d\Theta_{12}}{dT} = \Pi_{21}, \qquad (22)$$

and by substituting this value for Π_{21} in (21) we obtain the second relation

$$T \frac{d^2\Theta_{12}}{dT^2} = N_1 - N_2. \qquad (23)$$

Eqs. (22) and (23) are (1) and (2) of the text (remember that $S = d\Theta/dT$).

Appendix II

Useful Formulas for Practical Application of Thermoelectric Effects

The following formulas are all based on approximate treatments, but yield numerical answers sufficiently in agreement with experiment to be of considerable value. Their derivations can be found in many places, but perhaps may most easily be obtained in collected form in the Russian literature.

Efficiency of a Thermoelectric Generator

1) If r is the internal resistance of the thermoelement and R the external load resistance absorbing the useful power and we write $R/r = m$, then the efficiency η is given by

$$\eta = \frac{T_1 - T_0}{T_1} \frac{\dfrac{m}{m+1}}{1 + \dfrac{\kappa_r}{S^2}\dfrac{m+1}{T_1} - \dfrac{1}{2}(T_1 - T_0)\dfrac{1}{m+1}} \qquad (24)$$

where κ is the thermal conductivity of the entire thermoelement, and T_0 is the temperature of the cold junction.

2) The optimum value of the ratio m is

$$\left(\frac{R}{r}\right)_{opt} = M - \sqrt{1 + \frac{1}{2}\vartheta\left(\frac{T_1 + T_0}{T_0}\right)}. \qquad (25)$$

3) The maximum efficiency is

$$\eta_{opt} = \frac{T_1 - T_0}{T_1} \frac{M - 1}{M + T_0/T_1}. \qquad (26)$$

Note that [provided the individual legs of the thermoelement obey (10)] neither geometrical dimensions nor details of construction enter into the expressions.

Equations Relating to Cooling

1) The maximum coefficient of performance for junctions of dissimilar materials is given by

$$C_{opt} = \frac{T_0}{T_1 - T_0} \frac{\sqrt{1+M} - \dfrac{T_0}{T_1}}{\sqrt{1+M} + 1} \qquad (27)$$

where now

$$M = \frac{T_1 + T_0}{2T_0}\vartheta_{12} \qquad (28)$$

and ϑ_{12} is given by (11) of the text.

2) The maximum temperature difference obtainable is

$$\Delta T_{max} = \frac{1}{2}\vartheta T_0. \qquad (29)$$

Equations for Mechanical Consideration

1) For optimum operation, either as a generator or refrigerator, the element components should have dimensions adhering to the relation

$$\frac{A_2 L_1}{A_1 L_2} = \sqrt{\frac{\kappa_1 \rho_2}{\kappa_2 \rho_1}} \qquad (30)$$

where A is the cross-sectional area, L the length, κ the thermal conductivity, and ρ the resistivity of the "leg" indicated by the appropriate subscript. Generally design convenience will dictate that $L_1 = L_2$.

2) The radius of bending R_b of a thermoelectric device having a coefficient of linear expansion β is equal to

$$R_b = \frac{1}{\beta(T_1 - T_0)}. \qquad (31)$$

Relative Effect of Lattice and Electron Thermal Conductivity on the Figure of Merit

If the electronic and phonon thermal conductivities are κ_{el} and κ_{ph}, respectively, the figure of merit can be written

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$$\vartheta = \frac{S^2\sigma T_0}{\kappa_{el}} \frac{\frac{\kappa_{el}}{\kappa_{ph}}}{1 + \frac{\kappa_{el}}{\kappa_{ph}}}. \tag{32}$$

Approximate Expressions for the Important Parameters Obtained from Application (First-Order Approximation) of the Electron Theory

We define the transport integral of index r in the standard way as:

$$K_r = -\frac{16\pi m(kT)^r}{3h^2}\int_0^\infty l_{el}\left(\frac{\partial f_0}{\partial \epsilon}\right)\epsilon^r d\epsilon. \tag{33}$$

Here m = the electron mass, k = Boltzmann's constant, h = Planck's constant, T = the absolute temperature, l_{el} = the electron mean free path (note the implicit assumption), f_0 = the equilibrium distribution function (Fermi-Dirac), and ϵ = the energy.

1) Under isothermal conditions the electrical conductivity, σ, is

$$\sigma = e^2 \mathcal{E} K_1 \tag{34}$$

where e = the charge on the electron and $\mathcal{E}$ = the electric field.

2) The electron contribution to the thermal conductivity (with $\sigma = 0$) is

$$\kappa_{el} = \frac{K_1K_3 - K_2^2}{K_1 T}. \tag{35}$$

3) The absolute thermoelectric power (with $\sigma = 0$) is

$$S = -\frac{k}{e}\frac{(K_2 - \xi K_1)}{K_1 T} \tag{36}$$

where ξ = the degeneracy parameter (sometimes known as the Fermi energy, $\xi = \mu/kT$, see text) and is related to the electron concentration, n, through

$$n = 4\pi\left(\frac{2mkT}{h^2}\right)^{3/2}\int_0^\infty f_0\epsilon^{1/2}d\epsilon = 4\pi\left(\frac{2mkT}{h^2}\right)^{3/2} F_{1/2}(\xi)$$

Here we have defined the Fermi-Dirac functions,

$$F_s(\xi) = \int_0^\infty f_0\epsilon^s d\epsilon \tag{37}$$

which are well tabulated.[52]

4) The figure of merit, provided the contribution of the lattice to the thermal conductivity can be neglected (*i.e.*, for metals), is

$$\vartheta = \frac{(K_2 - kT\xi K_1)^2}{(K_1K_3 - K_2^2)}. \tag{38}$$

5) When the lattice thermal conductivity, κ_{ph}, cannot be neglected, the figure of merit, to a first approximation, is

[52] P. Rhodes, *Proc. Roy. Soc. A*, vol. 204, p. 396; 1950.

$$\vartheta = \frac{(K_2 - kT\xi K_1)^2}{(K_1K_3 - K_2^2)} \frac{\frac{\kappa_{el}}{\kappa_{ph}}}{1 + \frac{\kappa_{el}}{\kappa_{ph}}} \tag{39}$$

which is another form of (31).

6) If we assume that $l_{ph} = al_{el}$ where a = a numerical factor, and $l_{el} = l_e\epsilon^s$, we may write the figure of merit in terms of the tabulated Fermi-Dirac functions, $F_s(\xi)$ defined above. Then,

$$\vartheta = \frac{B[(2+s)F_{1+s} - (1+s)\mu F_s]^2}{(1+s)F_s + B[(1+s)(3+s)F_sF_{2+s} - (2+s)^2F_{1+s}^2]} \tag{40}$$

where

$$B = \frac{16\pi m(kT)^2}{3Nah^3v}$$

and N = the density of atoms and v = the velocity of sound in the crystal. Note that this still requires a single energy dependence of l; that is, a unique value of s.

The Lattice Thermal Conductivity

To a first approximation, we may write

$$\kappa_{ph} = \tfrac{1}{3}C_v v l_{ph}, \tag{41}$$

where C_v = the specific heat per unit volume. Using the high temperature value of $C_v = 3Nk$ and $l_{ph} = al_{el}$ we have

$$\kappa_{ph} = aNkvl_{el}. \tag{42}$$

The variation of a and v from material to material is discussed in the text.

VI. Acknowledgment

The author is deeply grateful for the many interesting discussions with numerous colleagues interested in this field. Unfortunately, the proprietary nature of most of this work precludes giving their names. Similarly, the writer is grateful to the reader for suffering the lack of detailed data on the latest and most interesting developments which was necessitated for the same reason.

Dr. G. E. Tauber of Western Reserve University, deserves special mention for his many lengthy and animated discussions of the theoretical aspects; he is responsible for most of the ideas expressed in Section V.

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A Communication Technique for Multipath Channels*

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Summary—Application of principles of statistical communication theory has led to a new communication system, called Rake, designed expressly to work against the combination of random multipath and additive noise disturbances. By coding the Mark-Space sequence of symbols to be transmitted into a wide-band signal, it becomes possible at the receiver to isolate those portions of the transmitted signal arriving with different delays, using correlation detection techniques. Before being recombined by addition, these separated signals are continuously and automatically processed so as to 1) apply to each an optimum weighting coefficient, derived from a measurement of the ionosphere response, and 2) introduce in each an appropriate delay such that they are all brought back into time coincidence.

After a brief introduction, a functional description of the system is presented. There follows a review of the communication theory studies, which indicate that such systems have certain optimal properties. Details of design of an experimental prototype Rake system are followed by the results of limited field tests of this prototype. Conclusions and recommendations for future work are given.

I. Introduction

MULTIPATH is a troublesome condition in many communication channels; the signal proceeds to the receiver along not one, but many paths, so that the receiver hears many echoes having, in general, different and randomly varying delays and amplitudes. The most familiar example occurs in high-frequency communication via the ionosphere [1, 2].

Multipath has been a problem from the early days of short-wave radio communication. Its influence on a communication system is usually described in terms of two effects—selective fading, and intersymbol interference—of which one or the other may be of predominant importance in the particular communication system being discussed [9].

Selective fading has to do with the relative *rf phases* of the signals delivered to the receiving antenna via the various paths. At any one frequency, the total received signal is a vector sum of individually delayed signals, their relative phase angles depending on the frequency and the echo amplitudes and delays. Therefore, since the echo amplitudes and delays are time varying, one observes large variations of the received signal strength at a single frequency as a function of time, or of the strength at a given time as a function of frequency; the latter is termed "selective fading."

Intersymbol interference is associated simply with the *time delay* between first and last significantly large echoes. If the modulation is rapid enough, the echoes appearing in this modulation will result in a jumbling or smearing of the intelligence, regardless of the form of signal used.

Previous system design for effectively combating multipath disturbances might be considered "passive"; that is, the effort has mainly been directed toward minimizing the undesired effects without having actual knowledge of the multipath characteristic. The use of single-sideband transmission (with or without exalted-carrier reception) [3], synchronous AM reception [4], and various forms of diversity reception [5] as anti-selective fading measures represent this type of approach. Fading has also been attacked by using coding [6].

The intersymbol interference aspect of multipath was long ago recognized to place a limit on the rate at which digital information could be communicated with time-division schemes. The use of multiple subcarriers ("frequency division") each having long symbol-waveforms to carry a fraction of the total information rate over multipath has been standard for many years, and has recently received additional impetus from new techniques which yield considerably greater efficiency of frequency spectrum utilization [7–8]. A method of extending frequency-division approach to transmission of analog information has also been proposed [10].

Two other techniques, of quite a different nature from those mentioned previously, are directed toward the actual or effective suppression of all but one dominant path. The first employs a complex, steerable antenna array [50], while the second uses frequency modulation [12]. The latter method appears to work under only certain conditions which, unfortunately, are seldom met in practice [11].

In contrast to the philosophy of the systems just enumerated, the system which is the subject of this paper performs a continuous, detailed measurement of the multipath characteristic. This knowledge is then actively exploited to combat the multipath effectively. Simply stated, selective fading is opposed by detecting the echo signals individually, using a correlation method, and adding them algebraically (with the same sign) rather than vectorially, and intersymbol interference is dealt with by reinserting different delays into the various detected echoes so that they fall into step again. For reasons that will be seen shortly, this approach has been dubbed the "Rake" system.

This system evolved from and is largely justified on the basis of the application of methods of statistical communication theory to the problem of communication through multipath disturbances. Soon after the idea was thus established, as often happens, a heuristic

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physical interpretation of the system became apparent. We shall, in the interest of clarity, present the latter straightforward, functional explanation of the Rake system first, in Section II. Without detracting from its significance, we defer until Section III the more rigorous derivation, which indicates that such a system, in addition to making intuitive sense, has certain optimal properties. In Section IV, the details of the construction of an experimental Rake system are given, and in Section V some results obtained in tests of the system over a transcontinental circuit are presented. In the concluding section the advantages and present drawbacks of the Rake system in relation to conventional systems are discussed, and suggestions are made for future improvement and extension.

II. Functional Description of the Rake System

A. General Principles of Design

Suppose we have decided to build a radioteletype system that will detect separately, then add up, each of the multiplicity of delayed signals arriving as a result of multipath. The transmission will then have to be wide-band,[1] for otherwise its time waveform cannot possess sufficient detail to permit the waveform at one instant of time to be distinguished unambiguously from that at another. Naturally, the wider the bandwidth, the finer will be the time resolution.

In addition to the requirement of providing multipath resolution, the transmission must of course be capable of carrying information. In a teletype system, this is accomplished by transmitting at will two distinguishable waveforms, one representing the Mark baud (or signalling element),[2] and the other Space. A commonly-used system is frequency-shift keying (fsk), where two sine waves of slightly different frequency are employed [13]. By analogy, we shall employ two different wide-band waveforms for the Rake transmission. Then by proper treatment of the received signal we can isolate a narrow region of delay and select for demodulation only the signal lying within it.

The principle of the Rake technique can be explained by starting with a simple fsk system shown in Fig. 1, which will shortly be reinterpreted to permit the use of a wide-band transmission and the attainment of the desired delay-isolation. The Mark and Space local oscillators shown in the receiver may be viewed as *reference signals*, with which the received signal is compared.

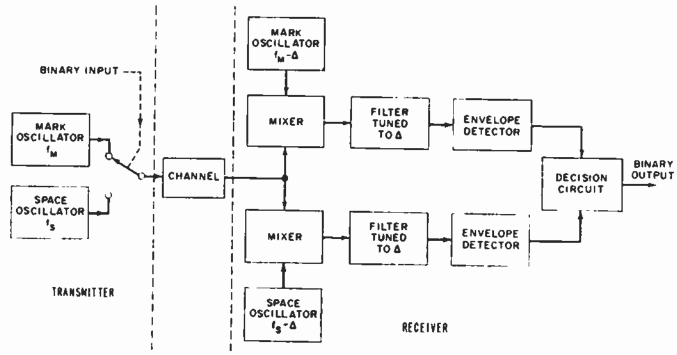

Fig. 1—Simple fsk system.

The comparison takes the form of determining (in the decision circuit) whether it is the envelope of the difference-frequency tone from the Mark filter or the Space filter that is the larger. The decision then represents one binary element of the output teletype sequence. Mixing (multiplying) the incoming and reference signals, and passing the difference-frequency result through a narrow-band filter is equivalent to the operation of *cross correlating* the two [16, 17] (that is, multiplying and then integrating).[3] Since the decision is based on which symbol yields the larger correlation, it is apparent that for best performance the frequency-shift $|f_M - f_S|$ should be sufficient to yield uncorrelated Mark and Space waveforms [18].

A plot of the output of the filter as a function of relative delay between the reference signal and a copy of it arriving from the transmitter is the autocorrelation function of the reference (the Fourier transform of its power density spectrum) [17]. In the case of a reference having a continuous spectrum of nominal bandwidth W, the autocorrelation function has a single peak of width about $1/W$ seconds, centered at the origin, and disappearing toward zero elsewhere. Portions of the transmitted signal arriving with various delays relative to the appropriate reference appear in the output of the corresponding integrating filter as sine waves of amplitudes given by the values of the autocorrelation function of the reference at the respective delays. With fsk, W is small, the central correlation peak is very wide, and path contributions will appear in the filter outputs for a wide variety of delays. We have already discussed the destructive interference (selective-fading) that can result when contributions add with random phases.

When the receiver of Fig. 1 is modified by the substitution of wide-band transmitted and reference waveforms, the correlation function narrows very greatly, so that the correlator will make use of only those echoes which arrive within $1/W$ of synchronization with the references. Thus we need only make W sufficiently

[1] The inherent advantage of using a wide-band transmission in a multipath environment was first suggested by R. M. Fano in 1952 (private communication), who reasoned that it averaged out the selective fading. Subsequent channel-capacity analyses carried out on the fading justified this notion from the information-theory point of view [14]. An informal note by Fano [15] proposes another method of employing wide-band signals against multipath. Kharkevich [52] has independently suggested the use of wide-band signals as a measure against selective fading, but his receiver employs completely incoherent detection and this is comparatively inefficient.

[2] The word "baud," taken from teletype usage, will be employed frequently in this paper instead of "signalling element," or "bit."

[3] A more detailed discussion of the relationship of this mix-and-filter scheme of correlation ("difference-frequency correlation" or "band-pass correlation") to the true mathematical operation of multiplying and integrating is given in [17].

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wide to separate out the various echoes (this also corresponds to making W wide enough to "average out" the selective fading). Finally, we can be assured of making use of all the signals delivered to the receiver by the entire multipath structure, regardless of whether it consists of discrete paths or is a continuum, if we use a series of such correlators, each synchronized at successive delay increments of roughly $1/W$. Enough of them must be employed to span a region of delay sufficiently wide to encompass all echoes that are likely to appear. The name "*Rake*" seems an appropriate designation for such a scheme. Making each correlator synchronize at its assigned value of delay can be done by inserting the right amount of delay in either the reference or received signals. The latter has been chosen for several important reasons to be dealt with later.

With the above ideas in mind, it is now possible to set down almost completely an elementary block diagram of such a Rake receiver. This is done in Fig. 2. The reference sources, emitting the same Mark and Space waveforms that the transmitter uses (except for a frequency displacement of Δ_1), feed a series of correlators arranged along a delay line whose input is the received signal. Each correlator consists of a multiplier and an integrator, but since the correlator outputs should be added together, in order to make full use of each echo, we may either combine after separate integrations, or, what is equivalent, use a common integrator. The latter is obviously simpler, so, as the figure shows, the output of each Mark multiplier is added to that of the others on a "bus," and the sum is then passed into a common integrator. The same is done with the Space multipliers. A decision on whether Mark or Space was sent from the transmitter is made according to which of the two sums of correlations is the larger.

In order to be sure that the difference-frequency signals appearing in the buses all add constructively, their phases must be brought into common agreement. Furthermore, those multipliers responding to large paths should have their contributions to the buses accentuated, while those not synchronizing with any significant path, and thus being affected mainly by the channel noise, should be greatly suppressed, in order to increase the snr and consequently decrease the probability of making a decision error. More precisely, we may invoke the well-known result [19] that the maximum snr of a weighted sum, of which each term is the combination of a signal and an additive, independent noise of fixed power, is achieved when the amplitude weighting is done in proportion to the signal strength (in voltage). Thus the weighting coefficients a_i in Fig. 2 should be proportional to as good a measurement as can be made of path strengths at the corresponding delays. (In Section III-A, the same result will be obtained from a more general point of view.) The phase corrections are indicated by ϕ.

Provided that W is large enough to isolate a number of independently-fading echoes, the deleterious effects of selective fading are largely eliminated by such a

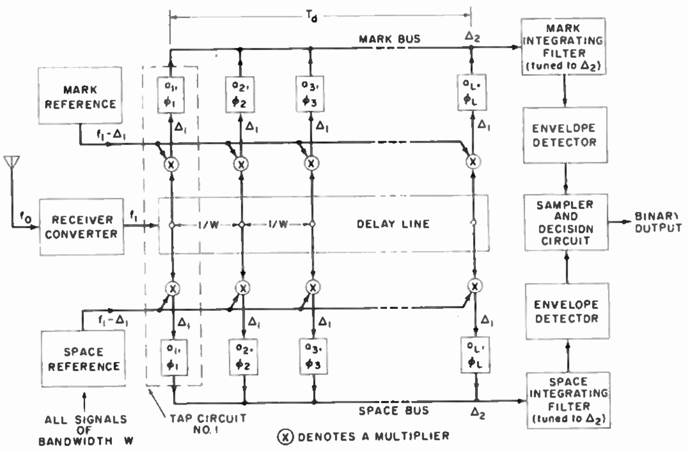

Fig. 2—Simplified block diagram of Rake receiver. The f's and Δ's refer to center frequencies. The a's and ϕ's are weightings and phase corrections, respectively based on path structure measurements.

scheme, since the path contributions are added algebraically, not vectorially. It happens that intersymbol interference is also eliminated by this Rake scheme, although a few words of explanation will be required to show that this is so. Consider the situation when the transmitter, which has been sending Mark, begins transmitting Space. Any particular echo delivers this Mark-to-Space transition to the receiver input at a different delay from the other echoes. The transition supplied by this echo, however, will be detected only at that delay line tap which corresponds to the delay of the echo, since at all other taps the echo contribution and the reference signals are uncorrelated. With the timing of the reference signals set properly so that they correlate with the last arriving echo, successively earlier echoes will correlate at points on the delay line correspondingly further delayed from its input. Thus the total propagation time from transmitter to receiver bus outputs is the time from transmitter to receiver input plus the line length to the appropriate tap, and this sum is *the same for all paths*. Hence only a single Mark-to-Space transition appears at the receiver output and intersymbol interference is eliminated.

B. Measurement, Weighting, and Phase Correction

The required weighting and phase correction is performed by the *measurement function* of the Rake receiver, which determines the path strengths and phases (including in the latter possible small variations in the positioning of the taps on the delay line). This measurement is accomplished again by the use of correlation,[4] since, as already shown, the use of a sufficiently wideband signal enables the various paths to be isolated. The integrating filters in this case, however, have a very long integration time, in order for the measurement to be as noise-free as possible. The integration performed

[4] This method is practically identical to that proposed by J. B. Wiesner and Y. W. Lee [20, 21].

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in the Mark and Space integrating filters is, by definition, performed over the duration of a signalling element or baud. But the time constant of each of the *measurement filters*, as they are called, should be as long as the rate of change of the multipath structure will permit. In order to assure uniform measurement regardless of the proportion of Marks and Spaces in the transmission, the measurement is provided by the sum of the Mark and Space correlations.

The method by which the correlation measurements are performed and applied is shown in Fig. 3, where a typical *tap circuit*, indicated by the dashed enclosure of Fig. 2, is depicted in greater detail. The same pair of multipliers (A and B) is used to obtain both the baud-correlations and the measurement-correlation (that is, the tones to be fed to the integrating filters and the measurement filter, respectively). The integration of the latter is performed by narrow-band filtering of the combined Mark-Space multiplier output. The amplitude of the sine wave out of this measuring filter (C) is proportional to the strength of the path to which the tap circuit responds, while the phase reflects the combination of the path and tap phasing. In the final multiplier tubes (D and E) this sine wave is mixed with the original outputs of A and B, thus performing the desired multiplication of these latter signals by the proper weighting function.[5]

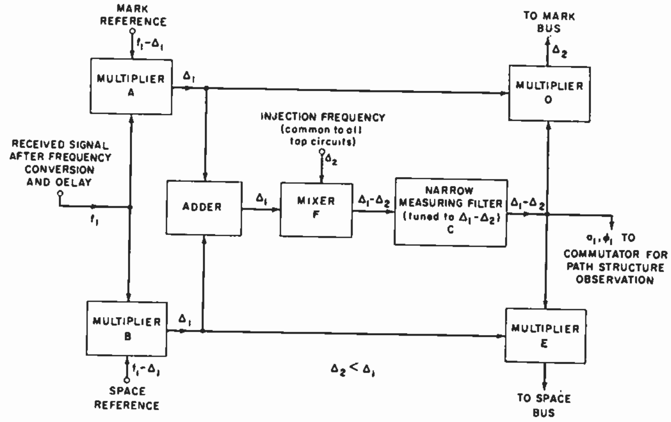

Fig. 3—Block diagram of tap circuit. (Dashed portion of Fig. 2.)

The way in which the phase correction is applied warrants more detailed discussion. To begin with, we require that the Mark and Space references at the receiver be in the same phase relationship to each other as at the transmitter, in order that the Mark-Space modulation produce no undesirable phase discontinuities through the measuring filter. When this is done the tones of frequency Δ_1 out of the first two multipliers, A and B, are phase-coherent with each other (although only one is on at a time). Let us call this phase angle θ_1. If θ_2 is the phase of the injection frequency Δ_2, then the phase of the tone of frequency $(\Delta_1-\Delta_2)$ reaching the second multiplier is $(\theta_1-\theta_2+\theta_0)$. θ_0 represents any additional phase shift added during filtering, and the tap circuits are aligned to have identical θ_0. In the second multiplier the original phase θ_1, whatever it was, cancels itself out, and only $(\theta_2-\theta_0)$ remains. But θ_2 is the constant phase of an oscillator acting as a common injection for all taps circuits. Therefore, all such outputs at frequency Δ_2 add with the same phase angle. (Recently we have learned that a practically identical scheme of phase alignment was invented by Earp [53] for predetection combining of diversity receivers.)

C. Reference Signals

As we have seen, the Rake system requires Mark and Space reference signals of wide bandwidth W in order for the receiver to be able to resolve the multipath and isolate and constructively utilize the contributions of the various arriving paths. But given this requirement of wide-bandedness, what particular sort of waveforms should these signals have, out of the great number of possibilities? The choice is narrowed considerably when the following additional factors are considered:

1) It must be possible to store the Mark and Space signals separately at transmitter and receiver.
2) It must be possible to keep the timing of the reference signals at the receiver from drifting appreciably with respect to that at the transmitter, and to align the time bases to allow for the lag given by the time of flight along the longest path.
3) If repetitive wide-band signals are used, they must have a repeat time at least as great as the multipath duration. If the repeat time is less, paths of separation equal to a repeat time or its multiples will be indistinguishable. This condition is later interpreted in terms of sampling theory.[6]
4) Each signal, Mark and Space, should have an autocorrelation function as near zero as possible within the first repeat time away from the central peak (which is approximately $1/W$ wide).
5) The spectral components outside the assigned frequency band should be small enough not to cause interference to other services. Also, as will be seen in the next section, the spectrum should be reasonably flat inside the band.
6) Because high-frequency transmitters tend to be limited by peak power rather than average power, it is desired that the transmitted signal envelope be reasonably uniform, in order that a given transmitter may supply as much energy as possible to each baud.

[5] The frequency translations of Δ_1 and Δ_2 allow mixers rather than ideal multipliers to be used for A, B, D, and E. In addition to simplifying the circuitry, the translation Δ_1 has the effect of broadening the tolerance on reference waveform synchronization [17] in elements A and B.

[6] We shall assume $S(\omega)$ to be continuous, whereas the shift-register transmission is actually a line spectrum. There is no real difficulty here, however, for by the sampling theorem [33] so long as the (complex) frequency samples are taken no farther apart than $1/T_M$, where T_M is the multipath spread, the multipath impulse response can be completely determined. The appropriate finite-interval, orthogonal interpolation functions are of the (sin nt/sin t) variety. See p. 13 of [14].

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7) For good discrimination in the presence of noise, Mark and Space signals should have a zero-shift cross correlation as near zero as possible. (The integration time here is the baud length T.)
8) Satisfactory performance of the Rake receiver measurement function requires that the signals have a long-term cross-correlation function (that is, integrated over the "ring time" of the measuring filters) that is zero for all τ shifts.

The study of *maximal-length* (or null-sequency) *binary shift register sequences* [22–24, 51] has revealed one way of constructing waveforms having the desired properties.[7] The complete scheme for generating Mark and Space signals, which is duplicated at transmitter and receiver, is shown in Fig. 4.

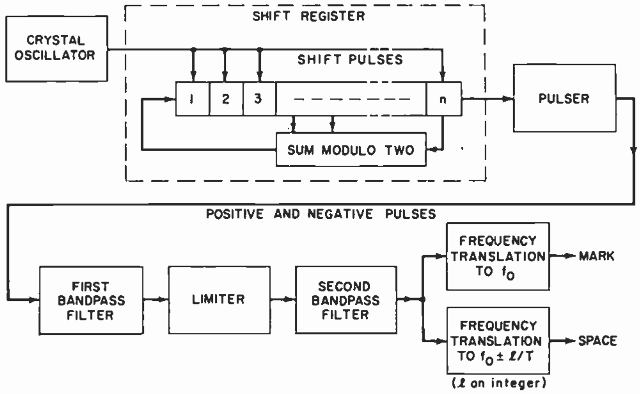

Fig. 4—A method of generating suitable reference signals.

The repetitive sequence is generated by a self-driven binary shift register in which the modulo-two sum of the digits appearing in the output and certain of the intermediate stages is fed back to the input. For an n-stage shift register, the maximal-length binary sequence has a period $m=2^n-1$, and is generated through suitable choice of which intermediate stages are fed back to the input. One such shift register can easily be set up at the transmitter, and another at the receiver (requirement 1). Each can be driven by pulses derived from crystal oscillators (requirement 2) with a period m made as long as necessary by choosing a sufficiently large number of stages n (requirement 3).

Maximal-length shift register sequences were chosen in preference to other classes of binary sequences because of a peculiar property that is useful in satisfying requirement 4: if we let positive and negative impulses represent the binary variable, the shift-register output is, of course, a sequence of such pulses repeating every m pulse. The autocorrelation function of this maximal-length sequence is a sequence of impulses, each of area (-1) except at the origin and integral multiples of the period m, where there are impulses of area m. This autocorrelation function is, for large m, very nearly that of a repeated impulse, and hence the power density spectrum exhibits a nearly uniform comb of spectral lines.[8]

To limit the frequency spectrum (requirement 5) the shift register output sequence is passed through a rectangular, band-pass filter of width W cycles per second. The envelope of the resulting signal fluctuates considerably, contrary to requirement 6, but by following the filter with a sharply limiting device this fluctuation can be suppressed. Such a nonlinear operation of course spreads the spectrum somewhat (resulting, for example, in a spilling of 10 per cent of the power outside the original band when the output pulse rate of the shift register is large enough compared to W to yield approximately Gaussian noise at the filter output [25]). Accordingly, a second filter, identical to the first, follows the limiter to restore the nearly rectangular power spectrum without producing as severe envelope fluctuations as are present in the output of the first filter.

To satisfy requirement 7, we seek orthogonal, wideband Mark and Space waveforms having about the same energy. If one reference signal, say Mark, has only slight envelope fluctuation (as described in the preceding paragraph), a frequency shift of only (l/T) is adequate to produce another nearly-orthogonal waveform that we can use for Space, even though the Mark and Space spectra may overlap considerably. Here l is any nonzero integer and T is the baud length.[9] This frequency shift is the final operation shown in Fig. 4. The long-term zero cross-correlation function stipulated by requirement 8 can be attained only if the long-term power spectra of the Mark and Space signals are disjoint. This will be assured so long as the frequency shift is not equal to or near a multiple of the reciprocal of the shift-register repeat time.

D. Integration and Decision

The integrations upon which each Mark-Space decision are based are required to be performed over the corresponding baud interval of length T. A filter of very high Q (time constant several times T), tuned to the bus frequency Δ_2, will faithfully perform the integration of the tap circuit outputs. It must be "quenched" or "dumped," however, after each decision to prevent ringing from the previous baud from carrying over into the succeeding baud intervals [8, 26].

Just previous to quenching, samples are taken of the

[7] The use of short pulses would fit naturally into the Rake concept, if it were not for the limitation of requirement 6.

[8] The effect of the small negative impulses between the large positive ones in the autocorrelation function is to suppress almost completely every mth tooth of the frequency comb, beginning at zero frequency.

[9] *Proof:* Let the Mark baud-waveform be represented as $R(t)\cos[\omega_0 t+\phi(t)]$; the frequency-shifted (Space) waveform is then $R(t)\cos[(\omega_0+2\pi l/T)t+\phi(t)]$. Neglecting terms near $2\omega_0 t$, the cross-correlation envelope at the time of sampling is

$$\max_{\theta} \int_0^T R^2(t)\cos(2\pi l t/T+\theta)dt \approx 0$$

for $R(t)$ approximately constant, and tends more toward zero as l increases. See also [8].

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envelopes[10] of both integrating filter outputs, and decision is based on whether the Mark or Space sample is the larger. The timing required to make the sampling and quenching operations occur exactly at the end of each received baud implies knowledge of the transmitter modulation timing. This is readily available, because of the degree of the transmitter-receiver synchronization already required of the reference sources.

III. Theoretical Foundations

In this section we present the communication-theoretical arguments that led to the Rake receiver design. Two rather distinct mathematical treatments have been pursued, which between them encompass a fairly realistic and complete set of assumptions. As yet, no unified theory has been forthcoming which includes the entire set of assumptions.

Both analyses are based on statistical decision theory [27, 28] and both lead, with some intuitive extension, to the Rake receiver previously described. We shall first discuss the simpler and more physical argument, and later briefly touch upon the more abstract analysis that actually first suggested the Rake configuration.

In the application of statistical decision theory, it is customarily assumed that the transmitter, its signals, and the channel are specified, at least on a statistical basis. Here, we shall assume that the receiver has knowledge of the possible transmitted waveforms and their *a priori* probabilities, and complete statistical knowledge of all random elements in the transmitter and channel. We seek the best such receiver. Generally the "best" receiver is taken to be that which achieves the minimum average "cost," suitably defined, of wrong decisions made by the receiver [27]. In the communication problem under study it appears that the appropriate cost function is simply the probability of error—that is, a Mark being mistaken for a Space is no worse nor better than the inverse error. This assumption becomes even more reasonable when we add the additional condition that the *a priori* probabilities of transmitting a Mark or a Space are to be equal. It has been shown that the corresponding optimum receiver, called by Siegert the Ideal Observer [29] is one that, from the received waveform, computes the *a posteriori* probabilities that a Mark, or a Space, was transmitted, and decides in favor of the symbol having the larger probability.

A. Analysis for Multipath Assumed to Be Perfectly Measurable

Using this philosophy, it is straightforward to find the optimum receiver in the case of a channel perturbed solely by additive white Gaussian (thermal or shot-effect) noise [30, 31]. Assuming that the Mark and Space waveforms have equal energy, the Ideal receiver in this case simply cross correlates the received waveform with the two stored references, and bases its decision on the symbol yielding the larger correlation. (Thus the simple correlation receiving system of Fig. 1, used previously for explanatory purposes, is seen to be an optimum system against white noise under the foregoing assumptions.)

The introduction of an arbitrary linear or nonlinear filter into the channel, preceding the noise, can be accommodated by a simple extension of this result, provided that the receiver knows or can measure the filter characteristic exactly. It is apparent that the optimum receiver in this case first passes its local reference waveforms through filters identical to the one in the channel and then performs cross correlations as before. (We assume that there is a negligible difference in the energies of the filtered reference wave forms.)

The above requirement that the receiver have complete knowledge of the channel filter is rather unrealistic for an ionospheric channel, since exact measurement of such a randomly-varying filter characteristic cannot be made so long as any noise is present. Fortunately, the multipath "filter" usually changes fairly slowly, and, furthermore, can be considered linear, so that good measurement accuracy can be achieved by a number of procedures, such as the cross-correlation method outlined previously. It is then plausible to employ the results of the measurement in constructing filters to process the stored waveforms for correlation, as just stated, ignoring the small measurement errors.[11]

We now examine more carefully requirements on the cross-correlation method of measuring the multipath characteristic described in Section II-B, and the process of applying it to correct the stored waveforms. We assume transmitted signal $x(t)$ to have a band-limited spectrum $S(\omega)$, and the multipath to be varying so slowly relative to the transmission bandwidth that a quasi-stationary analysis is allowable. (The usual Fourier-transform relations are assumed to hold between correlation functions and power spectra on a finite-observation-interval basis.) Then in order to measure everything about $H(\omega)$ (the "instantaneous" transfer function of the multipath) that is relevant, it is only necessary, and only possible, for the receiver to measure that portion $H_M'(\omega)$ of $H_M(\omega)$ occupied by the transmission.[6]

$$H_M'(\omega) = \begin{cases} H_M(\omega); & \omega \text{ in transmission band} \\ 0 & ; \text{ elsewhere.} \end{cases} \tag{1}$$

$H_M'(\omega)$ may be found from the cross-correlation function $\phi_{xw}(\tau) = \overline{x(t)w(t+\tau)}$ of the transmitted signal $x(t)$

[10] The phases of the signals on the output buses are fixed and known, so that in principle synchronous (coherent) detection could be employed to some advantage [4, 47]. In terms of probability of baud decision error, however, the equivalent transmitter power advantage of synchronous over envelope detection is only 1.14 db at an error rate of about one teletype character in three lines of copy, and becomes even less at smaller error probabilities. See [18], (13) and (39) with $\lambda = 0$ (orthogonal signals).

[11] Such a procedure closely parallels a suggestion of W. L. Root and T. S. Pitcher [32]. See also [14], pp. 12–14, where the use of a frequency-group transmission like that provided by the shift-register is suggested to accomplish the measurement necessary to correct the reference waveforms.

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and the received signal $w(t)$.[12] (It is assumed that $\phi_{xw}(\tau)$ is measured with an effective integration time appropriate to the rate of change of $H_M'(\omega)$.) Although $x(t)$ is of course not available to the receiver, $\phi_{xw}(\tau)$ can be found from correlating the received signal with the sum of both stored waveforms, providing their long-term cross-correlation function is zero for all τ shifts. Now the Fourier transform of this $\phi_{xw}(\tau)$ is the cross-spectral power density spectrum, $\Phi_{xw}(\omega)$ which is equal to $S(\omega)H_M(\omega)$ [35]. So,

$$H_M'(\omega) = \begin{cases} \Phi_{xw}(\omega)/S(\omega); & \omega \text{ in transmission band} \\ 0 \quad ; & \text{elsewhere.} \end{cases} \tag{2}$$

If $S(\omega)$ is now assumed to be constant within the band, (2) says that the impulse response $h_M'(t)$ of the receiver filters that should correct each stored symbol is given directly by

$$h_M'(t) = K\phi_{xw}(t) \tag{3}$$

where K is a constant. Finally, since $h_M'(t)$ has a spectrum limited to the transmission bandwidth W, it is only necessary to measure $\phi_{xw}(\tau)$ at points $1/2W$ apart in τ, or, since $S(\omega)$ is a bandpass spectrum, to find the envelope and phase of $\phi_{xw}(\tau)$ at points $1/W$ apart [33]. Even when the effect of channel noise is included in the measurement, any closer spacing of the sampling points gains nothing, since the error in measuring $\phi_{xw}(\tau)$ for a particular received $w(t)$ is, as a function of τ, also limited to bandwidth W.

The complete band-limited impulse response $h_M'(t)$ may be found by interpolating from its sample values, using interpolation functions of the form (sin t/t). Likewise, the stored symbols which are to be passed through $h_M'(t)$ can be expressed as a series of samples with interpolations of the same form. By virtue of the orthogonality of these interpolation functions, it is easy to show that the convolution resulting from filtering the stored waveforms with $h_M'(t)$ is exactly equivalent to multiplying the sampled values of the stored signal with those of $h_M'(t)$ and summing the products. Hence both the measurement and correction operations can be performed on a discrete basis, using multiple-tap delay lines with taps $1/W$ apart.

A block diagram of this Ideal receiver is shown in Fig. 5. Difference-frequency correlation[2] is employed to obtain the phase at the sampling taps. The sampled values of $\phi_{xw}(\tau)$ are obtained directly from the tap outputs they multiply, in order to keep the local filtering in alignment with the channel multipath filter. As mentioned earlier, both stored waveforms contribute equally to the measurement of $\phi_{xw}(\tau)$.

To show that this result is equivalent to the receiver described in Section II, we first rearrange the multiplying and adding elements of Fig. 5, as we may do freely

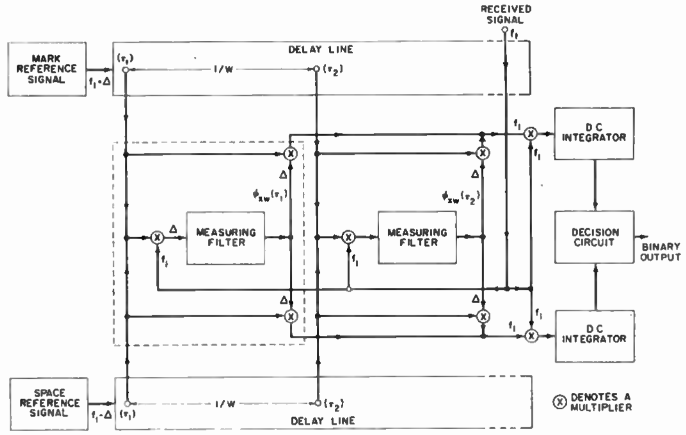

Fig. 5—Diagram of Ideal, delayed-reference receiver. The f's and Δ's appear as in Fig. 2. The $\phi_{xw}(\tau_i)$ are the multipath measurements. Measuring filters are tuned to Δ and have bandwidth approximately equal to the multipath fluctuation rate.

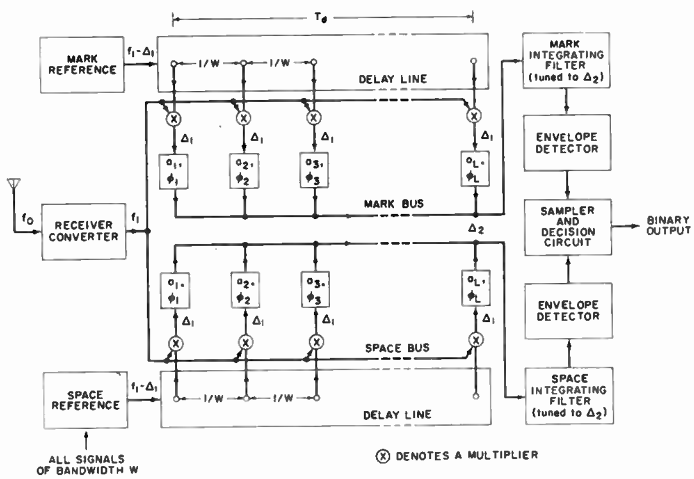

Fig. 6—Simplified block diagram of Ideal, delayed-reference receiver. This is Fig. 5 rearranged, employing the terminology of Fig. 2.

because of their linearity. We arrive at the configuration shown in Fig. 6 where the terminology is the same as that of Fig. 2. The double-difference frequency scheme of Fig. 2 has also been introduced, but results in no basic change in the operation. Envelope detectors are now indicated as well, but are incidental. In arriving at the final equivalence between Fig. 6 and Fig. 2, we study the signals and noise on the Mark and Space buses. At this point the only difference is whether the delay line appears in the incoming signal or in each reference signal. We will term the schemes of Fig. 2 and Fig. 6 the *delayed-signal* and *delayed-reference* configurations, respectively.

Finally, it is necessary to show that, independent of the form of the reference signals employed, the output snr from the integrating filters is substantially the same for both configurations, under the assumption that the length of the delay line T_d is significantly smaller than the baud length T.[13] Each integrating filter responds to

[12] G. L. Turin has shown that under certain conditions cross correlation is the optimum measuring operation for white noise present in the channel. See [34].

[13] We are indebted to Prof. R. M. Fano for the interpretation given in the present paragraph.

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signals only within about $\pm 1/T$ of the frequency Δ_2. Therefore, the noises adding on the common buses can be considered sinusoids of frequency Δ_2 having a fluctuation period no shorter than T, regardless of the form of reference signal. The only difference between the tap circuit contributions of the delayed-signal scheme (Fig. 2) and those of the delayed-reference scheme (Fig. 6) is that the latter are staggered in time by various fractions of T_d, and since such staggering is therefore small compared to the significant fluctuation period of the contributions, we conclude that the noise outputs of the two configurations are equivalent.[14]

There are three practical advantages of the delayed-signal scheme over the delayed-reference scheme. First, one delay line instead of two is required. Second, in the latter configuration, corresponding taps in the Mark and Space lines would have to be adjusted to and kept in phase coincidence. Third, coherent intersymbol interference (eliminated in the delayed-signal scheme) is still present in the latter scheme, as can be verified by repeating the same line of reasoning given in Section II-A.

In this analysis we have ignored the consequences of imperfect measurement of the path structure. The question of what form the best receiver takes when the path fluctuations become rapid cannot be answered by this simple reasoning. The next part summarizes the theoretical study that originally led to the Rake configuration. This analysis allows the paths to fluctuate, without assuming at the outset an explicit measurement function in the receiver.

B. *Analysis for the Assumption of a Discrete Path Structure with Known Time Delays* [42]

In order to extend the analysis to include random variations in the multipath structure, we must at present make the restriction that there are a finite number of discrete paths whose time delays are known *a priori* to the receiver.[15] The assumptions with which Section III began apply here, as well as the condition that the additive interference be white Gaussian noise. The paths are taken to be of the Booker-Gordon "scatter" type [39, 40], and to vary independently of each other, the transmitted signal, and the noise.[16] It is further assumed that the complete statistical description of each path is available to the receiver, in the form of $\phi_p(\tau)$. The quantity $\phi_p(\tau)$ is the "correlation function of the pth path," placing in evidence its average strength and rate of variation.

For this model, the corresponding Ideal receiver for the detection of isolated bauds has been deduced [42]. The exact solution, in open form, is given in (20) through (26) of [42]. (Unless otherwise stated, all numbers in this paragraph refer to [42].) An approximate closed-form solution appears in (49), and an analog computer for this equation is shown which bears a near-equivalence to the Rake configuration, having a tapped delay line, multiplying and integrating elements, and filters. The filters are matched in bandwidth to corresponding path stabilities, and may be viewed as performing the measurement-correction function of Section III-A in the present paper. The outputs to the decision circuit are designated by $\log_e L_k$; $k=1$ for Mark, 2 for Space. A bias block shown in the computer, and given by the last term in (49), may be presumed the same for Mark as Space, and hence eliminated from the decision.

By lumping together the Mark and Space measuring filters, leaving the resulting filters unquenched after each baud decision, and finally by placing the delay line taps at intervals of $1/W$ apart, to capture all paths that may arise,[17] the configuration of Fig. 5 in the present paper is reached. The final Rake receiver of Fig. 2 then follows in the same manner as in Section III-A.

IV. Experimental Realization of Rake System

In this section we shall describe the construction of an experimental prototype Rake system and mention some of the problems encountered in its operation. Since a large portion of the circuitry is quite conventional, considerable attention will be devoted only to those elements of the system that may be novel in realization or application.

[14] There is an aspect of the configuration of Fig. 2 that is related to the familiar notion of the *matched filter* (or conjugate filter) [15, 36]. With a simple rearrangement of the multipliers and adders, the delay line can be seen to be part of a filter matched to the multipath characteristic; that is, the filter has an impulse response that is the time reverse of the path structure. The output of this filter would then be fed directly to a single pair of Mark and Space correlators. This illustrates the fundamental difference between the Rake receiver and schemes which seek to pass the received signal through a filter that is the *inverse*, rather than the *conjugate*, of the multipath transfer function.

[15] In an important related study by Turin [37, 38], a result is obtained for the situation in which the strengths and delays of a number of discrete paths are presumed unknown to the receiver *a priori* [38], p. 160. This result is derived from a formula [[38], (24)] which can yield the Rake receiver in a manner similar to the approach of Section III-A.

Both Turin's study and ours [42] assume knowledge of certain channel parameters which must ultimately be measured in some manner. However the concept of symbol correction using a cross-correlation method of "instantaneous" path measurement, as employed in the Rake receiver, materializes as a unit (with some intuitive modifications mentioned in the text shortly) from the analysis of [42]. Turin, Root, and Pitcher [32], and Section III-A of this paper, on the other hand, logically unite the rather distinct operations of path measurement by correlation, and symbol correction using path measurement.

[16] We neglect the possible presence of a specular component which may be encountered, in addition to "scatter" (random component) in ionospheric propagation below the maximum usable frequency (muf) [41]. It has been shown by Turin [38] that at small signal-to-noise ratios such specular components play a dominant role in conveying information. From observations during recent experiments, however [43], we do not feel that conclusive knowledge is yet available of the relative amplitudes of the specular and scatter components to be expected in below-muf propagation. This question is being studied further by M. Balser and others in Lincoln Lab., Mass. Inst. Tech., Lexington Mass.

[17] Turin's work ([37], pp. 40–44) indicates that when the path time delays are unknown *a priori*, the weighting of the tap contributions (Section II-B) to the output buses of Fig. 5 should be a nonlinear function of the measurement. A very rough approximation to this refinement is considered in Section IV-F.

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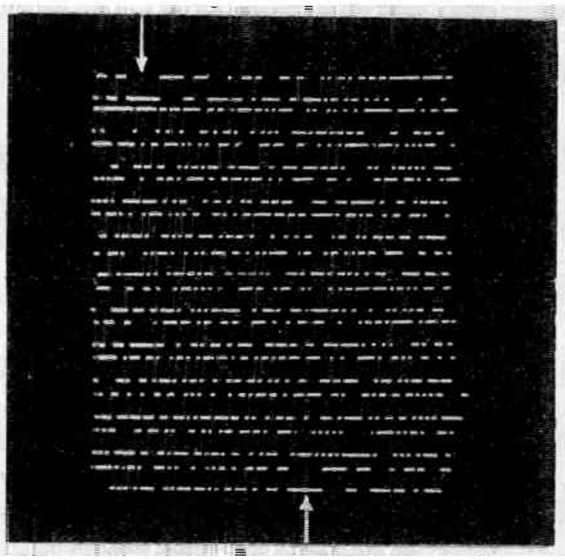

Fig. 7—Maximal-length shift-register sequence. $n=10$ stages, period $m=1023$. The entire sequence is shown, reading from left to right and top to bottom, with some overlap. The beginning and end are shown by arrows.

A. Transmitter

The transmitter shift-register (Fig. 4) has $n=10$ stages, and the binary sum of the last stage and fourth from last stage is fed back to the input stage to yield a maximal-length sequence of period $m=2^{10}-1=1023$. The sequence is shown in Fig. 7. The driving clock pulses occur at a 120-kc rate, so that the repeat time of the sequence is $1023/(120\times 10^3)=8.525$ msec, which is longer than normal ionospheric multipath duration.[18] The pulser passes positive and negative pulses of 0.1 μsec duration into a four-pole Tchebycheff filter having a 10-kc half-power bandwidth, centered on 455 kc. The filter output is shown in Fig. 8(a), next page, and is observed to have large envelope fluctuations. The spectrum of the output of this first filter is presented in Fig. 8(b). Following the band-pass limiter the spectrum is spread, as shown in Fig. 8(d), and in order to suppress the spectral "tails" the limited signal is passed through a second Tchebycheff filter, 10 kc wide and centered on the spectrum. The waveform and spectrum of the resulting final wide-band "carrier" are shown in Fig. 8(e) and 8(f), respectively. The envelope is seen to be considerably more uniform than that of Fig. 8(a).

Frequency-shift modulation of the "carrier" is conveniently accomplished between the band-pass limiter and the second filter. The injection for this conversion is a tone of average frequency 155 kc, so that the second filter is actually centered on 300 kc. The deviation from 155 kc is $\pm 2/(22\times 10^{-3})\approx \pm 90.90$ cps (arbitrarily + for Mark, − for Space), corresponding to an exact integral multiple of the reciprocal of the 22-msec baud length, as required for orthogonality. (The output of the conventional, motor-driven teletype distributor is here electronically retimed to provide bauds of this length, with accurately determined end points [26].)

The 300-kc signal from the second filter is heterodyned to the allocated transmission frequency. Harmonics of a 100-kc Western Electric 0-76/U crystal clock, from which the 120-kc shift-register driving pulses and the baud-timing are also derived, are employed to perform the major portion of the frequency translation. Other crystal oscillators accomplish the remaining interpolation.

B. Receiver Reference Sources

The Mark and Space reference sources at the receiver are identical to those at the transmitter, except that there is a fixed downward translation of frequency by about 20 kc just ahead of the second filters which are now centered on 435 kc. (In Fig. 2 and Fig. 3, $f_1=455$ kc, $\Delta_1=20$ kc, $\Delta_2=9$ kc.) The Mark-reference injection frequency is (20 kc+90.90 cps) and the Space-reference injection is (20 kc−90.90 cps). The phase of the 181.80 cps Mark-Space deviation at the receiver (*i.e.*, the tone that would be obtained if Mark and Space references were multiplied together) must be kept in good alignment with the transmitter deviation. This is required so that modulation produces no phase transients in the receiver measuring filters, as explained in Section II-B. The required high order of deviation-frequency agreement and stability is obtained by first providing crystal oscillators at (155 kc−90.90 cps) and (20 kc−90.90 cps) for the Space frequencies at the transmitter and receiver, respectively. For Mark, the 181.80 cps$\approx 4/(22\times 10^{-3})$ deviation from these frequencies is obtained at both transmitter and receiver from a harmonic of the 22 msec baud timing using single-sideband modulation. Once the phasing of the transmitter and receiver 181.80 cps deviations is brought into agreement, the precision of the 100-kc clocks from which the baud timing is derived is sufficient to maintain it for many hours.

C. Conversion and Delay of Received Signal

The rf signal is received and translated to the standard intermediate frequency of 455 kc by a conventional receiver (R390/URR). Double conversion is employed, the major portion of the frequency translation being performed by the harmonics of the receiver 100-kc crystal clock. As at the transmitter, other crystal oscillators supply the remaining interpolation.

The IF signal is fed into a cascade of two ultrasonic delay lines, of a novel type,[19] each 1.5 msec long and having about 50-kc bandwidth at 455-kc center frequency. This multipath of up to 3 msec duration can be accommodated. One of the lines is shown in Fig. 9.

[18] Measurements of the spectrum of this sequence in the region from 30 cps to 10 kc were made using a General Radio Model 736-A Wave Analyzer. The amplitudes of the spectral lines, spaced $1/(8.525\times 10^3)\approx 117.30$ cps apart, were found to be uniform within ±1 per cent.

[19] These lines were designed and built under the guidance and with the invaluable assistance of R. M. Lerner, H. Penfield and L. P. Romano, of Lincoln Lab., M.I.T., Lexington, Mass.

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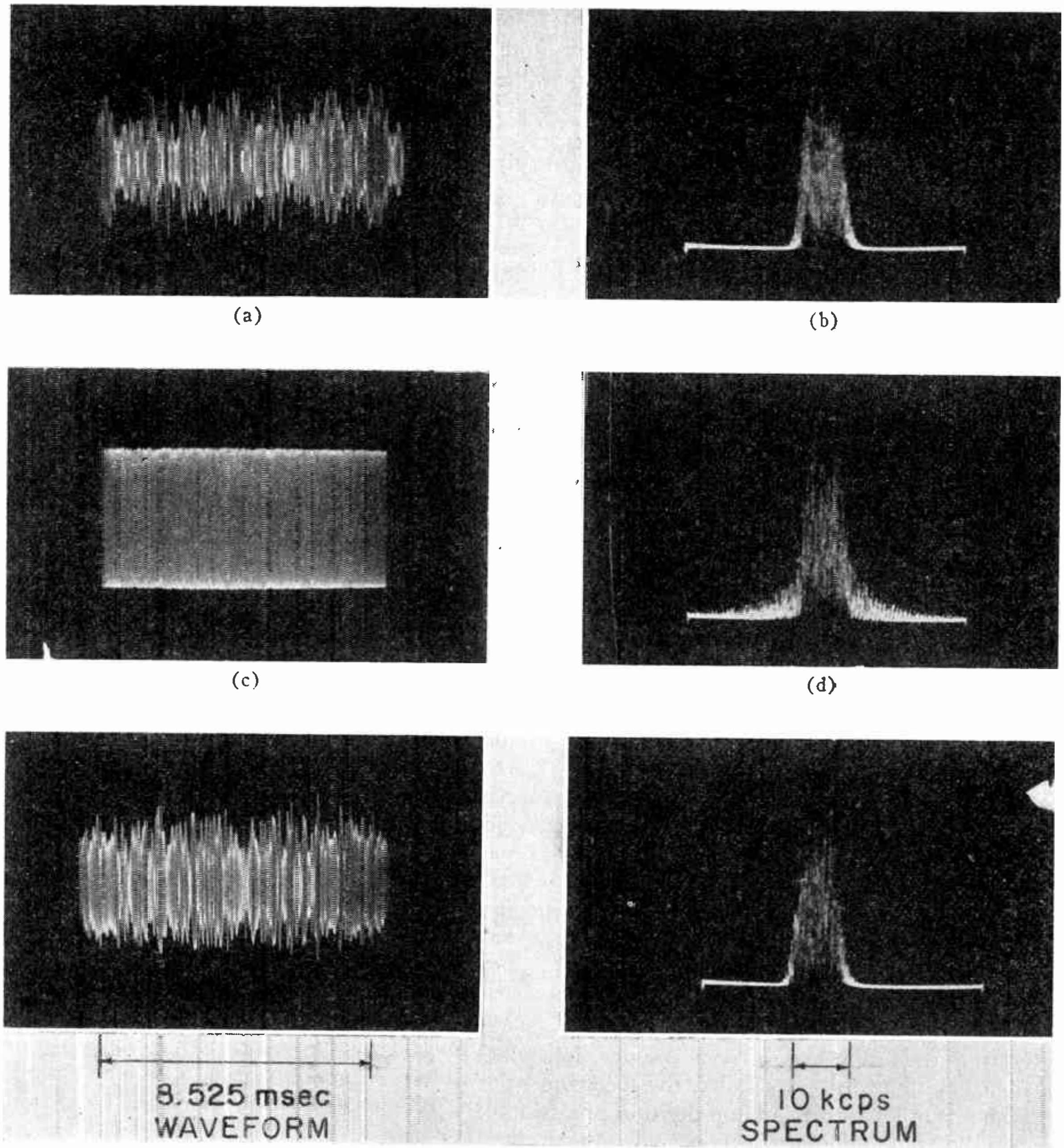

Fig. 8—Reference source waveforms and spectra. (a) and (b) First filter output. (c) and (d) Limiter output. (e) and (f) Second filter output.

D. Tap Circuit

The following parameters of the tap-circuit block diagram of Fig. 3 have been established for the experimental Rake system. From the preceding remarks, the center frequency f_1 of the received signal, after conversion and delay, is 455 kc. The reference sources are centered at 435 kc, so that Δ_1 is 20 kc.[20] The reference frequency (and therefore the output bus frequency) Δ_2 has been chosen to be 9 kc, after consideration of the strengths and locations of various intermodulation products and the ease of suppressing them by series or parallel traps. The measuring filter is therefore centered on $\Delta_1-\Delta_2=11$ kc, and its bandwidth is made 1 cps to correspond roughly to the average rate of ionospheric fluctuations.

The multipliers and mixer are single type 6AS6 tubes, which have already been widely used as analog multipliers. Operated on a difference-frequency basis with proper biases on the control and suppressor grids, they are very linear over a dynamic range (at their output) of

[20] The 455 and 435-kc figures just mentioned refer to Mark and Space signals alike. However, the difference between Mark and Space is a 181.80-cps shift of the internal structure within the nominal band. Therefore if Mark is received, for example, the Mark multiplier A produces a $\Delta_1=20$-kc tone, but Space multiplier B produces a tone 181.80 cps away from Δ_1. The latter is suppressed in the Space integration (see next section).

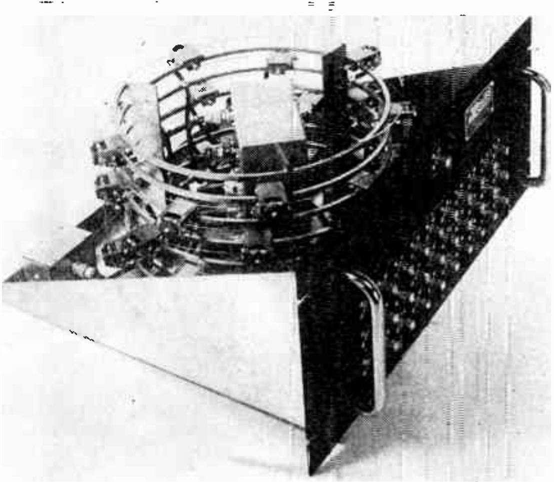

Fig. 9—Delay line. 1500-μsec helical ultrasonic line with the driving circuits in the center, driving transducer at back left, and termination at top right. The center frequency is 455 kc, bandwidth 50 kc; the injection gain input-first tap 10 db, insertion loss of wire element 28 db, ratio of signal to spurious, 25 db. Line uses 1/32-inch diameter invar rod, driven piezoelectrically in longitudinal mode through exponentially tapered matching section. Driving transducer is quarter-wavelength cylinder of barium titanate. Low-Q output coils are biased by permanent magnets. Rod is terminated in winding of tape clamped in rubber (not shown).

100 db. The gains in the tap circuit are set so that as the received signal grows large the grids of the second multipliers reach saturation at approximately the same point. The measuring filter is an 11-kc NT-cut, flexure-mode bar crystal, driven in the series-resonant mode from a cathode follower. The complete assembly of the 30 plug-in tap circuits and two delay lines, which comprises the major part of the Rake receiver, is shown in Fig. 10.

E. Integrating Filters and Decision Circuitry

The integrating filters have high-Q coils enclosed in temperature-stabilized ovens. They are tuned to 9 kc and employ Q-multiplication circuits for good integration linearity. The quenching is accomplished by diodes and is completed in a few hundred microseconds.[21]

Envelope detectors, connected in series opposition, are applied to the two integrating filter outputs, and the net dc voltage is applied to the decision circuit, which consists of a limiting dc amplifier, sampler, and Schmitt trigger driving the teletype printer.

F. Auxiliary Features and Devices

Effective operation of the Rake system with a minimum of manual intervention calls for the use of automatic frequency control (afc), automatic gain control

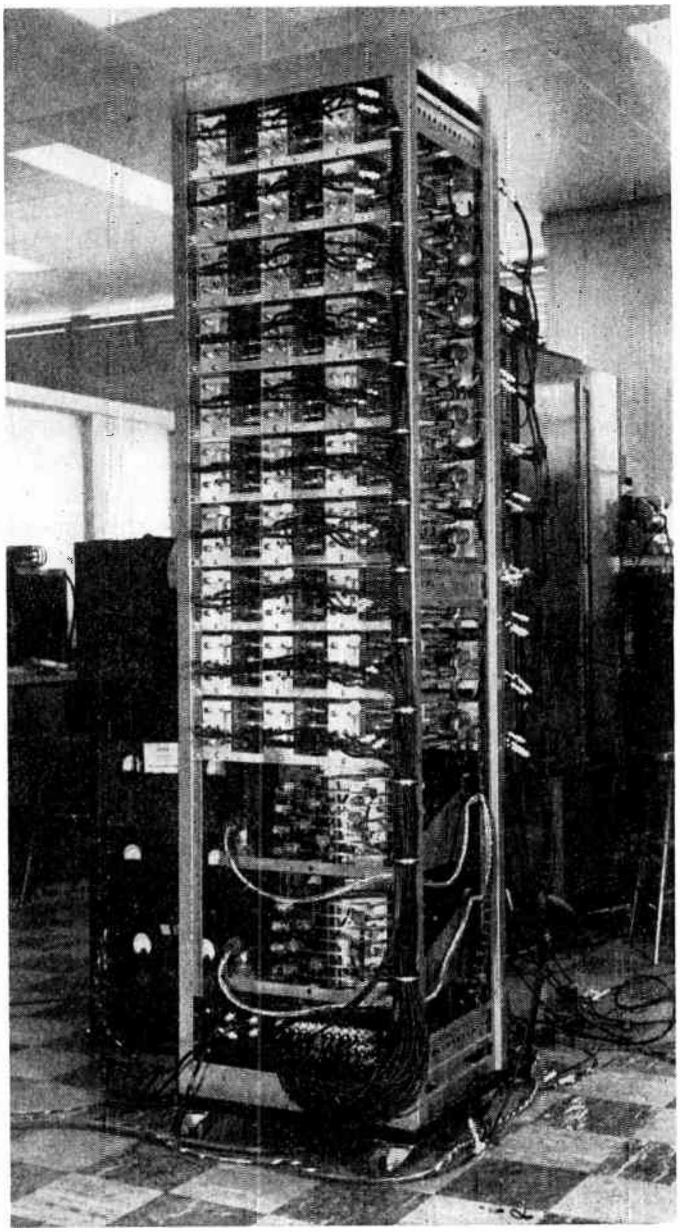

Fig. 10—Rake receiver rack. Inputs are through cables in rear; output buses run across front. Two delay lines and commutator located at bottom.

(agc), and a means for observing the multipath structure. These auxiliary devices will now be discussed.

For satisfactory operation, over-all frequency alignment of the system must be kept to within a fraction of the measuring-filter bandwidth, regardless of drift in the transmitter and receiver converters and ionospheric Doppler shift. The frequency-sensitive element for the required afc can be realized by comparing the phase of the sum of the two Rake bus outputs with that of the injected frequency Δ_2. The sharp phase characteristic of the measurement filters thus provides a sensitive indication of frequency alignment. If it should happen that different paths have widely different Doppler shifts, this scheme may have to be modified considerably.

[21] Integration patterns resemble those shown in Fig. 3 of [8]. For example, if the output of the Mark filter is observed, a linear buildup is seen when a Mark is transmitted, while a sequence of four small scallops results when Space is transmitted. The latter waveform corresponds to the frequency separation of four reciprocal-baud-lengths between Mark and Space (see Section IV-A).

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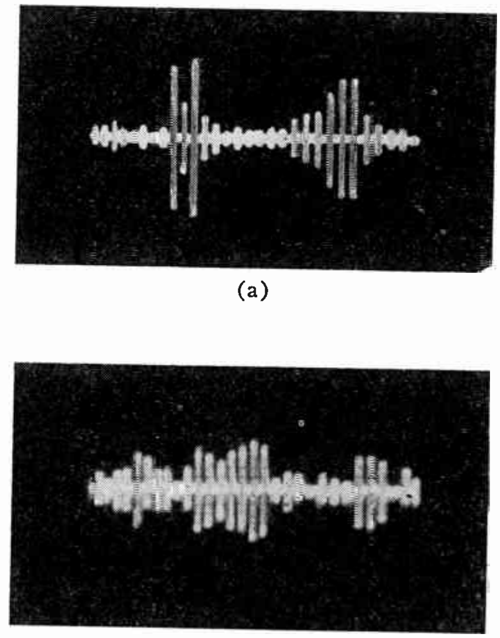
Changes in the received signal level appear as level changes on the Mark and Space output buses in a square-law relationship because of the weighting. Since the integrating filters will not accommodate a very great dynamic range, agc must be employed so that such level fluctuations are not great.

Three additional manual adjustments are needed from time to time. First, care must be taken that the entire multipath structure is contained within the multiple-correlation interval T_d; this is accomplished by adding or subtracting occasional shifting pulses from the receiver reference shift register. Second, when only a few tap circuits are responding to significant paths (for example, when there is only one ionospheric path), it is well to disconnect the remainder from the buses. This eliminates the small amount of noise these circuits contribute due to noise fluctuations in their measuring filter outputs (which ideally should be zero when no path is present but are not because of the nonzero measurement filter bandwidth).[17] Third, the bandwidth of the measuring filters should be kept in correspondence to the path stability, although this is a far-from-critical adjustment.

These three manual adjustments made during operation, together with such initial adjustments as are required when the system first goes "on the air," are facilitated by an auxiliary feature of the Rake receiver. The multipath measurement is presented continuously on an oscilloscope by means of a commutator which samples the measuring filter outputs, as indicated in Fig. 3. The operator can thereby study the ionospheric behavior at all times, simultaneously with the reception of intelligence. Since dynamic path phase information is also contained in the measuring filter output, it is possible to observe multipath phase stability over long distances, by employing a suitable horizontal sweep on the oscilloscope. Similar observations have already been reported elsewhere [43].



(a)

(b)

Fig. 11—Path structures observed on Rake receiver. Both observations made in 12-mc band with 3 milliseconds observation interval. (a) Fine structure in left group and broadness of right group indicate two closely spaced pairs of paths, with interpath separation of 1.4 milliseconds. (b) Path continuum immediately after a sudden ionospheric disturbance.

V. Test Results

On-the-air tests of the Rake system over a transcontinental circuit were conducted during the summer of 1956. The transmitter was situated at the U. S. Army Radio Station AAG in Davis, Calif., where a maximum undistorted power of 20 kw was available from a Collins FRT-22 single-sidebanded transmitter. The receiver was located at the Army receiving facility in Deal, N. J. Both transmitter and receiver employed rhombic antennas of about 12-db gain. Transmissions were made near 8, 12, and 17 mc at various times of day.

Fig. 11 shows some typical path structures observed by commutating an oscilloscope along the measurement outputs of the successive Rake receiver tap circuits. Fig. 11(a) represents the usual situation in which discrete paths are present, while Fig. 11(b), taken immediately after a sudden ionospheric disturbance, presents the rarer case of an apparent continuum of paths. It was possible to observe a long continuum of paths of up to 5 msec duration by operating the system in the region of high absorption below the so-called lowest usable frequency (luf).

Fig. 12 presents an observation of the selective fading of the wide-band signal, as seen on a spectrum analyzer. On one occasion, a two-path structure was encountered near 17 mc in which there appeared to be a differential Doppler shift between the two paths equal to 1.5 cps. This was measured by tuning the highly stable injection to the Rake receiver (R390/URR) for maximum response to first one path and then the other, and confirmed by observing on the spectrum analyzer the rate at which the selective fades drifted across the transmission band. (See also Fig. 8 of [9].)

Satisfactory performance of the system was obtained for a wide variety of multipath structures. On one occasion 1800 lines of teletype copy (representing nearly six hours of operation) were printed at the receiver without error. Although experimental evidence is not yet conclusive, it appears that the Rake system performs best at frequencies well below the maximum usable frequency (muf) where there is enough multipath spread that there is small probability of a simultaneous fade on all paths. Once, when the multipath extended over 5 msec, satisfactory printing was obtained with a Rake observation interval of only 1.5 msec. This indicates that it is not absolutely necessary to make use of all significant paths, although performance against noise naturally improves as more paths are included.

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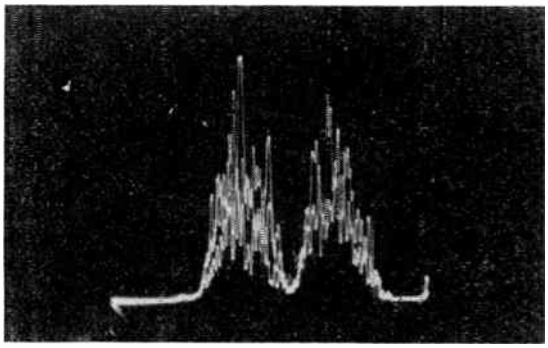

Fig. 12—Selective fading as observed on a Panadaptor. The individual frequency components are not resolvable. The width of the spectrum is 10 kc.

Very rapid (≈10–20 cps) fluctuations of the multipath structure were occasionally encountered during disturbed conditions, and this resulted in poor system performance, chiefly because of the inability of the measurement filters (whose bandwidth had not been made adjustable) to respond rapidly enough.

A brief test was made which confirmed that a delay-line tap spacing of 100 μsec was as efficient as 50 μsec, and that a spacing of 150 μsec resulted in marked degradation in performance, in accordance with the sampling analysis of Section III-A.

It was of interest to compare the performance of the Rake system with that of an fsk system operating under the same conditions. A standard frequency shift of 850 cps was employed, and fsk receivers of both the conventional (CV-116/URR) and predicted-wave [26] type were employed. The fsk receivers were operated both with and without dual space-diversity, the diversity combining being accomplished through a common limiter in the CV-116, and through diode combining in the predicted-wave receiver.

Comparisons between the Rake and fsk systems were made on the basis of probability of teletype character error (five bauds make up a character) during times when the only additive channel disturbance was additive noise. The transmitter was calibrated to establish equal average power levels for the two systems, and alternate transmissions of groups of ten lines each (lasting about two minutes) were made using Rake and fsk at various levels. Times when the only additive channel interference was solely noise were rare, and consequently only a small amount of comparison data was taken. Two error runs are presented in Fig. 13. In Fig. 13(a) comparison is made with the conventional CV-116 fsk receiver, while in Fig. 13(b) predicted-wave reception is included as well. (It should be noted that comparison at these high error rates cannot be extrapolated to the error rates of commercial interest.) Both these runs employed space-diversity reception with the fsk transmission, but with Rake space diversity is not required and was therefore not used.

Experiments were made to determine the extent to which the information rate could be sped up by decreasing the baud length. This was an attempt to take advantage of the absence of coherent intersymbol interference in the Rake receiver. It was found that the decreased integration time (baud length) caused a prohibitively high error rate (≥1 error per line) for values of $T \leq 5$ msec (or $TW \leq 50$).

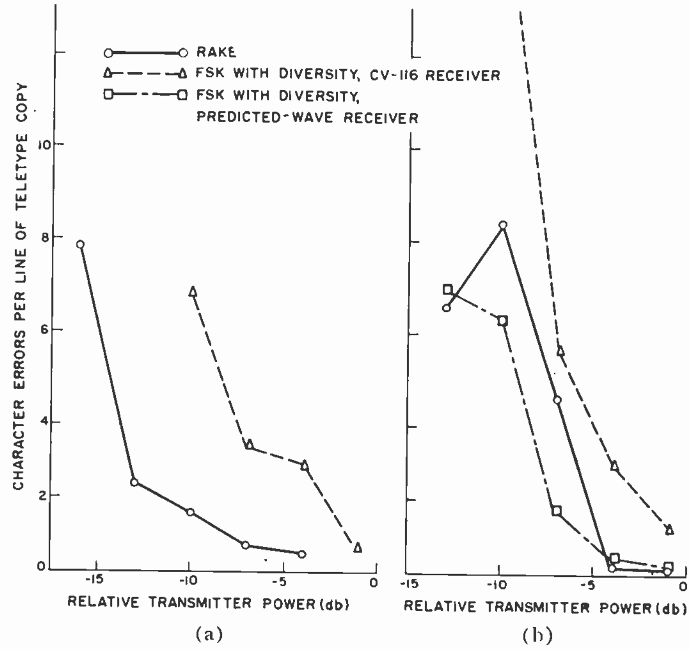

Fig. 13—Comparisons of Rake vs fsk. (a) Observed 2100 EST, 12-mc band. (b) Observed 0100 EST, 8-mc band.

VI. Conclusion

The preceding pages have outlined the features of a new technique for dealing with a channel perturbed by random multipath disturbances and additive noise. This Rake system derives its name from the operations performed at the receiver, where a number of correlation detectors are provided at uniform delay increments, with their outputs added. The addition takes place only after each correlated output has been suitably weighted and phase-corrected for maximization of the snr of the sum. By employing wide-band signals, such a procedure allows algebraic rather than vectorial addition of the signals from the various paths, thus eliminating the degradation usually associated with selective fading. This method of detection, weighting, and combining was originally suggested by the study of optimum probability-computing receivers for multipath and noise (Section III-B).

It was pointed out that the original concept, developed in the communication-theoretical derivations of Section III, can be modified by inserting the required delays in the incoming signal rather than the reference signal. The result of this alteration is that a reconstitution or collapsing of the multipath delays is achieved which effectively eliminates intersymbol interference. Thus the receiver output behaves as though there were

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a single propagation path of high strength, rather than a series of weaker paths spread in time. It should be possible to exploit this feature to send information in bauds whose durations are equal to, or even shorter than, the multipath duration.

The procedure of detecting, weighting, and combining signals with different delays clearly has an analog in the frequency domain. Using again a wide-band signal, one can build a receiver that divides the spectrum up into segments (which need be no narrower than the fine structure of the selective fading, that is, the reciprocal of the multipath duration), then detects, weights, and adds their contributions.[22]

Although it is clear that the selective fading and intersymbol interference resulting when multipath is present in conventional systems can be effectively disposed of, a new and undesirable effect appears in the Rake system—a form of noise in the output. This is most easily seen when we consider the "matched-filter" interpretation of the Rake receiver.[14] From this aspect, the channel multipath is cascaded with a time-reversed duplicate of itself at the receiver. This combination yields a new multipath characteristic of twice the original spread, but one in which there is now a strong central path formed of the time-reconstituted contributions of the original channel paths. Correlation detection is then employed to isolate this path and suppress the others on either side of it, thus effectively eliminating the spread. While there are thus no coherent contributions from any paths but the strong central one, the remainder of the "new" paths do produce fluctuation in the integrating filter outputs, by virtue of the finite correlation time allowed (the baud length T). This fluctuation is virtually identical to that produced by channel noise.

It can be shown[23] that the snr at the integrating filter output is proportional to TW, the product of the baud length by the reference signal bandwidth. Here the noise[24] is the sum of that due to interference in the channel, and that due to the correlation fluctuation just mentioned. Experiments with the prototype system in which W was fixed at 10 kc showed that even with no channel noise present, satisfactory error rates were attained only for TW about 50 or larger.

We may interpret this number as follows. For $W=10$ kc, bauds as short as 5 msec can be used. This figure should be substantially independent of the duration of the multipath provided the Rake delay line has a length at least equal to this duration, because of the delay reconstitution. Because the length of the delay line of the prototype system was only 3-msec, the 5-msec limitation on baud length made it impossible to demonstrate the full possibilities of time reconstitution. On the other hand if a bandwidth of 50 kc could have been employed, then 1-msec baud lengths should have been usable over normal ionospheric multipath.

It is recognized that at present the Rake system, like any high-order frequency-diversity scheme, is relatively uneconomical of bandwidth. On the other hand, the system performs with high reliability, and in fact may work best, over a channel having extensive multipath. Thus Rake systems may prove useful in providing satisfactory communication at those frequencies, well below the "optimum working frequency," which would normally be abandoned on account of multipath. There is also promise that the Rake technique can be extended to the multiplexing of a considerable number of independent, wide-band transmissions into the same frequency band [49].

Another of the important directions for future work on systems of this type is the reduction of the minimum TW product required for a given error rate. Several approaches look promising. Among them are: 1) the search for reference signals with reduced short-time correlation fluctuations. 2) Schemes of cancelling out these fluctuations. The fluctuation waveform depends only on the reference waveforms (known exactly), the multipath characteristic (known almost exactly) and the sequence of Marks and Spaces transmitted (this can be deduced at the receiver with low probability of error by using a step-by-step hypothesis test). So far neither of these approaches has produced workable results.

Much improvement is probably possible in equipment size and complexity. An ingenious series of ideas has been proposed by Sunstein, Steinberg, and Ehrick [44] for time-sharing a single tap unit rather than building a multiplicity of them. Also, it would be desirable to make some present manual operations automatic, such as the complete suppression of those tap circuits not responding to any significant path.

The assumption has been made throughout that the fluctuation period of the multipath is long compared to the symbol lengths, although this is not demanded in the original development of Section III-B. In cases where the fluctuation period becomes comparable to or less than the symbol length, the Rake receiver takes a particularly simple form: instead of weighting each correlator output by a long-term measurement of itself, one need only pass it through a filter identical to the measurement filter of the present tap circuits and square the filter outputs [45, 46].

One final observation should be made: the Rake system as presently conceived does not appear to be directly adaptable to the transmission of analog, rather than digital, information. Further study may extend the

[22] Such a scheme was conceived independently by Ehrick and Steinberg [45].

[23] By properly interpreting the correlator snr result [(13) of [17]], an equation can be derived for Rake output snr. It consists of the product of the dimensionless quantity TW with a time-varying quantity depending only on the multipath, the spectral shape of the reference, signal and noise, and the ratio of powers of incoming signal and noise.

[24] Making the baud length equal to or less than the multipath duration is a violation of the assumption under which one can attribute equal output snr to the two forms of Rake system: the delayed-signal, and delayed-reference forms (Section III-A). The delayed-reference scheme has the smaller output noise because various portions of this noise add now with delays greater than the pertinent fluctuation period (the baud duration).

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applicability of the Rake technique in this and other directions.

Acknowledgment

The development of the Rake system has been the result of the stimulation, support, and hard work of a number of people. We would particularly like to note the contributions of R. S. Berg, C. W. Bergman, J. W. Craig, R. E. Gay, W. C. McLaughlin, W. B. Smith, and C. A. Wagner to the system development and testing, and the assistance of the U.S. Army Signal Engineering Laboratories (Fort Monmouth, N. J.) and the Sixth Army (San Francisco) in providing field test facilities. We would like to record our great indebtedness to W. B. Davenport, Jr., whose guidance in earlier work underlying the Rake system made this concept possible. This paper reflects in many ways the influence of R. M. Fano who has spent many hours and not a few graduate students (including the authors) in pursuit of solutions to the multipath problem.

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Low Noise Tunable Preamplifiers for Microwave Receivers*

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Summary—An investigation of noise reduction in backward-wave amplifiers has yielded two significant results. 1) This type of amplifier has been demonstrated to be capable of very-low noise figures and therefore constitutes an entirely new class of microwave receiver tubes. Tube noise figures less than 6 db for a 25 per cent tuning range and less than 4.5 db for a 10 per cent tuning range of the amplifier's narrow pass band have been attained. 2) Tube noise figures of well *under 4 db* have been measured, which are the lowest recorded to date on any type of microwave tube. This performance results from a special low-noise gun which was developed for use with hollow electron beams and which features a new type of beam launching mechanism.

The experimental *S*-band tubes and "Christmas-tree gun" are described. Detailed noise performance of the tubes is presented, as well as other data relating to the operation of the tube as a receiver component. On the basis of these experiments, it is concluded that still lower noise figures are possible using the basic concepts of this new gun, not only for backward-wave amplifiers, but also for other types of microwave tubes.

INTRODUCTION

A fundamental requirement in virtually all microwave systems is maximum receiver sensitivity. This need has not only been intensified with the advent of new types of broad-band systems but is now also coupled with the demand for receivers possessing greater flexibility. Modern receivers must often be capable of operation over appreciable portions of an octave, sometimes with almost instantaneous tuning—and this must be accomplished while maintaining very-low noise figures

In view of these considerations, an investigation of noise reduction in backward-wave amplifiers was undertaken in an effort to adapt the unique voltage-tuned filter characteristics of this type of tube[1,2] to receiver applications. It was evident that if the noise figure of this device could be lowered to values approaching the ultimate theoretically predicted limit[2–4] (6–7 db) it would constitute a new type of microwave receiver tube whose characteristics would potentially make possible the evolution of a special class of receivers featuring both high sensitivity and great versatility.

Some other more fundamental considerations initially motivated this investigation. Previous experiments on the nature of noise in electron beams have all utilized small cylindrical beams.[5,6] Although the generalized one-dimensional theory predicts identical minimum noise figures for all beam-type amplifiers,[4] extensive experiments on noise reduction have been concerned almost exclusively with the conventional traveling-wave tube.[7,8] In addition to the application of present concepts of beam noise to a different type of tube, the use of the backward-wave amplifier permits the investigation of a range of parameters and beam geometry not

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heretofore encountered. The beams employed in helix-type tubes are annular and of relatively large diameter ($\gamma a \sim 5$). Interaction with the circuit field utilizes a space charge mode having a sinusoidal azimuthal variation. The interaction impedance is typically more than an order of magnitude smaller than that of forward-wave amplifiers, and the optimum conditions which must be realized for minimum noise figure are different.[9] Thus, aside from the practical importance of low noise, *per se*, the investigation was directed toward a study of the assumptions and validity of the existing noise theory under these new conditions.

The phase of the investigation reported in this paper concerns the experiments which establish the backward-wave type of amplifier as a new class of receiver tube. Following a brief comparison of the properties of backward-wave amplifiers and traveling-wave tubes, as related to receiver problems, the design considerations and features of the experimental tubes are described. These tubes were designed for operation at *S* band and include both the single-helix and cascade circuit configurations. Since the general characteristics of backward-wave amplifiers have been discussed previously in the literature, the noise performance of the tubes is emphasized here.

The most important feature of these tubes is the unique noise reducing property of the hollow beam electron gun. Noise figures comparable with those of the best traveling-wave tubes and crystal mixers have been attained over tuning ranges greater than 25 per cent. Tube noise figures of *4 db and lower* have been measured at the center of the band. This value is several db under the usually accepted theoretical limit for this tube. Although the mechanism responsible for this performance is not yet fully understood it is apparent that it results from the flexibility provided by the special gun geometry. This, together with the fact that the cathode techniques were rather crude compared with those customarily employed in low-noise traveling-wave tubes, leads to an optimistic viewpoint regarding the ultimate noise figures that might be achieved.

It is suggested that the rapidly tunable selective preamplification afforded by this type of tube may lead to practical application of the trf receiver concept at microwave frequencies.

General Characteristics

The backward-wave amplifier can be regarded as an active electronically tuned filter having narrow bandwidth and high gain. These characteristics complement those of the conventional traveling-wave tube and make the low-noise backward-wave amplifier attractive as a receiver tube in those types of systems where the inherently large bandwidth of the traveling-wave tube limits the maximum attainable threshold sensitivity. For purposes of illustration consider the properties of several possible types of microwave preamplifiers, as indicated in Fig. 1. It will be assumed that the various amplifiers can be made to have identical noise figures and arbitrarily high gain.

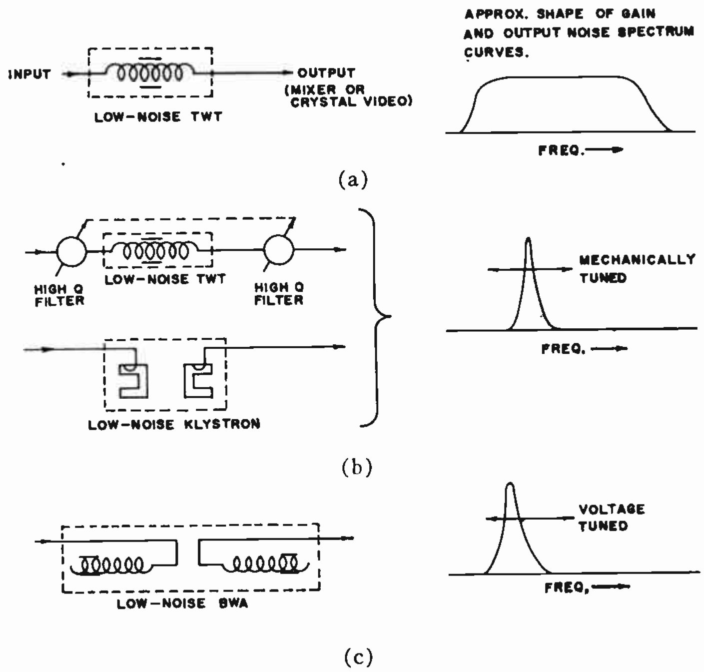

Fig. 1—Schematic illustrations of several possible types of microwave preamplifiers. The principal gain and noise output power characteristics of these configurations are indicated by the curves at the right.

The first configuration, consisting simply of a low-noise traveling-wave tube, comprises a wide open receiver which accepts and amplifies all input signals in a wide frequency range. Equally important, however, is the fact that it generates noise power over this same band. Such a preamplifier can be used in two types of receivers. In a crystal-video receiver (*i.e.*, after sufficient rf gain the signal is detected and passed through a video amplifier), the over-all sensitivity is limited primarily by the output noise power integrated over the large band of the traveling-wave tube. In this case, the fact that the tube noise figure may be very low is of relatively minor importance.

On the other hand, if employed as the input stage of a superheterodyne receiver, the first detector stage acts as a filter for most of the noise. Here the wide open characteristic of the receiver has been eliminated and broad-band operation is achieved on a time-sharing basis by rapid tuning of the local oscillator. Since the frequency of the IF amplifier is usually very small compared with the bandwidth of the traveling-wave tube, the noise at the image frequency will still result in a deterioration of over-all receiver sensitivity of about 3 db.

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An improvement in the design from the viewpoint of maximum sensitivity would be obtained by placing narrow-band filters both preceding and following the traveling-wave tube, as shown in Fig. 1(b). The first filter rejects noise and other spurious signals occurring in the broad frequency band of the amplifier and thus eliminates saturation and intermodulation products which could result. The second filter rejects the internally generated tube noise in all but the pass band of interest, centered on the carrier frequency. Although this arrangement achieves a sensitivity limited only by tube noise figure and required system bandwidth, it is restricted to systems operating on a single frequency or, at best, to systems in which a rapid tuning rate is unimportant. A low-noise klystron would, in principle, have the same characteristics although its mechanical tuning range would be less than that possible with the traveling-wave tube.

A low noise backward-wave amplifier combines the features of these preamplifier configurations. Its useful bandwidth corresponds to filters having Q values of the order of 100 to 1000, which restricts the frequency range in which signals can be amplified and internal noise generated to values consistent with the requirements of most systems. It can be tuned almost instantaneously over a frequency range approaching that of most traveling-wave tubes. Since it operates on a regenerative mode, the bandwidth can be adjusted electronically with an accompanying variation of gain. In fact the device in itself constitutes a microwave analog of the familiar low-frequency trf-type amplifier but with the important advantage of freedom from mechanical tuning and tracking of the selective circuits.

Description of Experimental Tubes

The noise figure of a beam-type amplifier is a unique function of the noise pattern on the beam, as characterized by the noise power standing-wave ratio and the position of the standing-wave minimum relative to the circuit.[4] In the case of the backward-wave amplifier, the optimum values of these parameters,[9] computed on the basis of the usual one-dimensional theory, vary over a considerable range as the pass band of the tube is tuned in frequency. This arises from the rapid variation in plasma wavelength as the beam voltage is changed. In order to maintain a low noise figure the electron gun must therefore constitute an extremely flexible transducer for the noise-excited space-charge waves between the cathode and circuit.

Another complicating factor in the design of low-noise guns for helix-type backward-wave amplifiers (at least at S band) is the very low range of current densities typically employed in the annular beams. The plasma wavelength in the precircuit drift region is of the order of 10 to 20 inches, leading to an optimum spacing between the gun and helix of more than 6 inches, as calculated from models similar to those used in the design of low-noise traveling-wave tubes.[10,11] Since this spacing is impractical it was decided to move the gun as close as possible to the circuit (approximately 0.85 inch) and to consider methods of increasing the inherent flexibility of the transducer section of the gun as a phase shifter for the noise space-charge waves.

A key feature which is made possible by the use of high-γa annular beams is a relatively large cross-sectional beam area. This results in the possibility of attaining very high values of *total* gun perveance and thus furnishes a degree of freedom not obtainable with conventional solid beams. With high perveance the first accelerating anodes can be operated at extremely low potentials and still produce a required total beam current. In this region the plasma wavelength is short and the phase shift through this section is very sensitive to small changes in average beam potential. The high degree of flexibility thus attained permits fine adjustment of optimum initial noise conditions on the beam at the circuit entrance with a minimum over-all gun length.

Another degree of flexibility, essential to the gun's performance, was obtained by disposing a special control electrode around the sides of the cathode. This electrode, which really acts as an auxiliary anode, is the most important feature of the gun. It allows the perveance to be varied over a considerable range. Thus, the gain, which is a sensitive function of beam current, can be maintained constant for a wide range of voltage on the first accelerating anode. The latter potential, which is of the order of several volts, can then be adjusted for optimum transformation of the space-charge waves independently of parameters in the circuit region.

At the same time, the geometry of the control electrode was designed so that it could profoundly alter the potential distribution in the immediate vicinity of the cathode in such a way as to permit possible manipulation of the basic noise characteristics on the beam as well as the excitation mechanism of the space-charge waves. Thus, the underlying assumptions and validity of the one-dimensional noise theory can be examined under a wide range of conditions.

A sketch of the low-noise gun employed in this investigation is shown in Fig. 2. The mean cathode diameter is 0.335 inch and the area of the emitting annulus is 0.1 cm^2. The special control electrode faces the edges of the cathode. It is followed by four anodes which constitute the space-charge wave transducer.

In the intital guns, the five electrodes inside of the beam were each suspended by three equally spaced grid wires extending radially through the beam from the outer anodes. These grids intercepted a minimum of 6 per cent of the total cathode current. In spite of this

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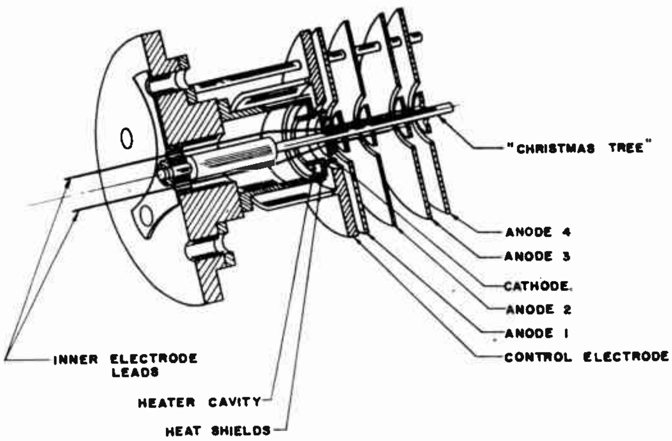

Fig. 2—Sketch of the "Christmas-tree" low-noise electron gun.

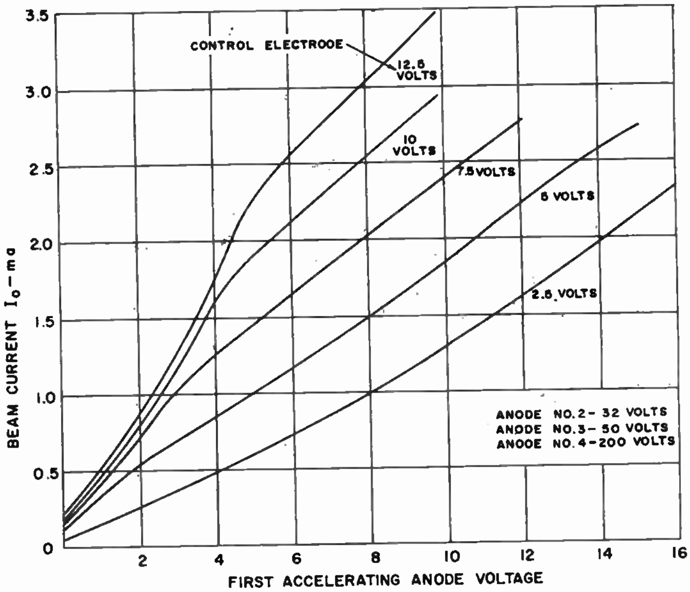

Fig. 3—Typical dc characteristics of the "Christmas-tree" gun. Usual low-noise operation occurs at a beam current of about 1 ma with the control electrode approximately 7 volts positive with respect to the cathode.

high interception, noise figures of approximately 11 db were attained on both a single helix tube and a cascade amplifier. It was felt that this clearly demonstrated the principal features of the gun since the contribution to the measured noise figure due to partition noise was estimated at about 4-db minimum.

In order to eliminate interception and hence partition noise in the vital gun region, the novel type of construction illustrated in Fig. 2 was developed. The electrodes inside the beam are mounted on a "Christmas tree" which protrudes through the center of the cathode. The individual leads to these electrodes are insulated and brought out at the base of the gun. The potentials of these leads introduce a negligible perturbation of the field at the position of the beam. Since the electrodes are symmetrically disposed and close to the thin beam, the accelerating field is virtually constant over the beam's cross section, resulting in a stiff transformation of dc beam impedance through the transducer region.

Typical curves of the dc gun characteristics are shown in Fig. 3. The potential of the control electrode is *positive* with respect to the cathode and the potentials of the second, third, and fourth anodes are held at their usual operating values. Regarded as a triode, the gun has a low value of mu, as evidenced by the magnitude of current at zero voltage on the first anode. Typical low-noise operation occurs prior to the onset of saturation in the neighborhood of 1-ma beam current. Here, at 7.5 volts on the control electrode, the total microperveance (defined in terms of the first anode potential) has a value of about 190.

The first two tubes using this low-noise gun were single-helix amplifiers. Noise figures somewhat under 7 db at 12-db gain were attained on each of these tubes[12] in the neighborhood of 2600 mc, which is very close to the theoretically predicted minimum value. A striking characteristic of these tubes was that the noise figure was a typically increasing function of current, *i.e.*, the best noise figures were invariably realized at relatively low values of gain or, equivalently, at small ratios of operating current to start oscillation current. The variation was much larger than could be predicted theoretically and seemed to be an intrinsic property of either the tube or the gun. This behavior was used to establish design criteria for low-noise cascade amplifiers.

In a cascade tube the input circuit by itself constitutes a single-circuit amplifier. This section was designed to operate at a low value of gain (under 10 db) in the region in which the best noise figures were obtained in the preliminary experiments on single-circuit tubes. The signal-to-noise ratio on the beam is essentially established in this section of the amplifier. The output section must then be designed to operate closer to the oscillation threshold in order to achieve the desired high gain. This was accomplished by increasing the length of this circuit and thus reducing the start-oscillation current in this region. The two sections thus operate at different values of the critical ratio of operating-to-starting current.

The tube pictured in the lower half of Fig. 4, following page, is typical of the low-noise cascade amplifiers discussed in this paper. No effort was directed to minimizing the physical size of these experimental models. The model shown at the top of Fig. 4 illustrates the sort of size reduction that can be achieved.

The helices in the experimental tubes have a mean diameter of 0.409 inch and a pitch of 0.125 inch. Fig. 5 shows the resulting nominal tuning curve for the tubes. Here frequency refers to the center frequency in the narrow pass band of the tube at a particular beam voltage.

The lengths of the input and output helices are 5.5 and 7.5 inches, respectively. The glass envelope was

[12] M. R. Currie and D. C. Forster, "Experiments on noise reduction in backward-wave amplifiers," Proc. IRE, vol. 45, p. 690; May, 1957.

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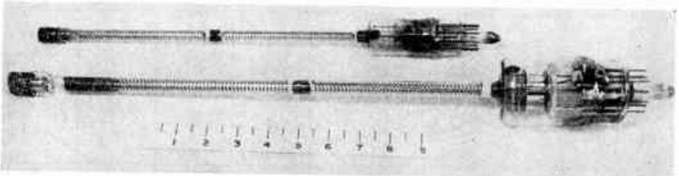

Fig. 4—Low-noise backward-wave amplifiers. The larger tube is the research model described in this paper. The smaller tube is a later model which has been greatly reduced in size and scaled for operation at the center of *S* band.

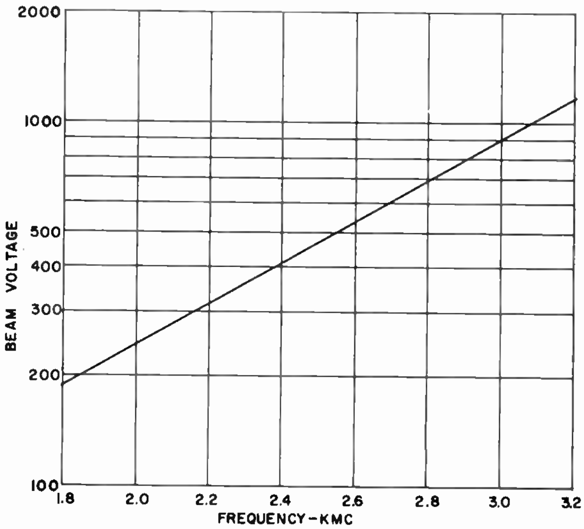

Fig. 5—Tuning curve of the experimental low-noise backward-wave amplifiers.

made purposely thin so that the lossy terminations at the extreme ends of the tube could be applied externally. Thus, for example, it was possible to change the effective length of the output circuit relative to the input circuit and investigate noise behavior for a considerable range of currents and relative electronic gains in the two sections.

No attempt was made to minimize the cold insertion loss of the helices, *e.g.*, by silver or copper plating techniques. The theoretical noise figure of backward-wave tubes is particularly sensitive to distributed circuit loss.[9] Moreover, this loss should be minimized over the entire length of the input circuit because the beam noise and signal are injected at opposite ends and a continuous feedback occurs between beam and circuit.

It should also be emphasized that the cathode and processing techniques employed in these tubes were relatively crude. Experience with traveling-wave tubes[13] and triodes has shown that special types of cathode nickel, very uniform activation, and extremely smooth and dense cathode coatings are essential to achieving minimum noise generation. Since the investigation was initially aimed at studying first-order effects, no special techniques were employed. The nickel was of a nondescript variety. The cathode coating was not particularly uniform; its thickness was not controlled and the edges were often ragged. Inspection of typical cathodes after activation often revealed small glazed areas and patchiness.

In short, it appears that there is room for considerable refinement in techniques before the full noise capabilities of these tubes and this special type of low-noise gun are known.

Method of Measurement

A number of precautions were taken to insure the accuracy of the noise figure measurements. After initial adjustment of the tubes with a direct reading system similar to that used by Peter,[14] final optimization and detailed measurements were performed manually using the system shown in Fig. 6. This system permits several methods of measurement which can be used to check the data.

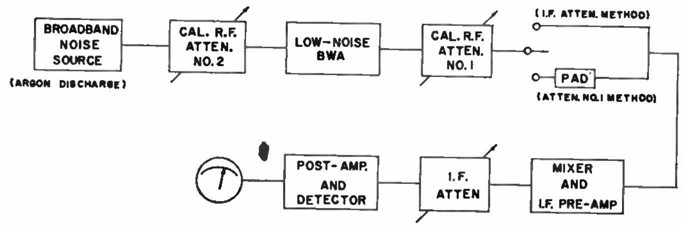

Fig. 6—Block diagram of the noise-figure measurement system. Three types of measurements were employed to minimize experimental errors.

The bulk of the data was obtained by means of the calibrated waveguide attenuator designated No. 1 in Fig. 6. Here, the measured noise figure is independent of receiver noise figure ($\sim$12 db) regardless of tube gain. A 6-db pad was used to isolate the mixer from the tube. These data were checked using a waveguide-below-cut-off attenuator at the intermediate frequency. The latter method measures the noise figure of the tube and mixer combination and agrees closely with the first method at relatively high values of gain. The IF attenuator is in itself a secondary standard.

A possible source of error, particularly in backward-wave amplifiers, is a change in gain due to a change in source match with the noise source in its fired and unfired conditions. Attenuator No. 2 was employed in the familiar power-doubling method[14] to eliminate this source of error. Here the noise source is maintained in the fired condition. This attenuator can also be varied over a wide range while monitoring the output noise power from the tube in order to detect any possible change in tube gain due to slight mismatch presented by the noise source in the unfired state.

Subjective errors in reading the output meter were minimized by averaging as many as a dozen measurements at a given point. The rf attenuators were cali-

[13] E. W. Kinaman and M. Magid, "Very Low-Noise Traveling-Wave Tube," paper presented at Second Annual Technical Meeting on Electron Devices, Washington, D. C.; October, 1956.

[14] R. W. Peter, "Direct-reading noise-factor measuring systems," *RCA Rev.*, vol. 12, pp. 269–281; June 1951.

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brated as a function of frequency and continually compared with the IF attenuator. Various argon discharge lamps were substituted in the noise source; an excess noise of 15.3 db was assumed for these lamps. Other possible sources of error (mismatches, variation in ambient temperature, finite insertion loss of fired noise source, etc.) were in a direction to subtract somewhat from the measured noise figure but were not taken into account. The error in absolute noise figure is believed to be less than ± 0.3 db and is probably on the conservative side.

The noise figure measurements are referred to the input transducer of the tube (*i.e.*, the insertion loss of the coaxial input lead was subtracted out). In a packaged unit the noise figure would be several tenths of a db higher than the tube noise figure; however, tube noise figure is a measure of the inherent capability of the tube.

Experimental Results

The general results obtained with single-circuit amplifiers incorporating the "Christmas-tree" gun have already been discussed. It should be mentioned in addition that there were occasional indications of considerably lower noise figures on these tubes. In fact, at one frequency, independent measurements using different noise sources indicated a noise figure under 5 db at approximately 10-db gain. These results were unfortunately not readily reproduced. It was apparent that the nature of the cathode was changing with time, probably due to barium migration effects. However, it was felt that some unusual mechanism of noise reduction was present which could be studied more readily in cascade type tubes where the higher stable gains permit more accurate measurements.

A total of four cascade amplifiers of the type pictured in the lower half of Fig. 4 was constructed in the course of this study. The first two (denoted A and B) were identical except for a slight change in over-all gun length. These tubes were used to study the noise performance of the gun as well as various loss configurations on the circuits. In the other tubes (C and D) the cathode geometry was varied in order to obtain possible frequency scaling information.

The most significant results of the investigation are presented in Fig. 7. Here, the noise figure of Tube A is shown as a function of midband frequency. The over-all gain is maintained at a constant value of 20 db. It should be emphasized that the potentials of the various noise-transducer electrodes were *not programmed so as to obtain a minimum possible noise figure* at each frequency in the curve of Fig. 7. These potentials were maintained constant at the values which minimized the noise figure at 2.53 kmc. Thus the conditions of operation under which these data were obtained correspond closely to those which would be employed in receiver applications.

From a practical viewpoint, this curve demonstrates

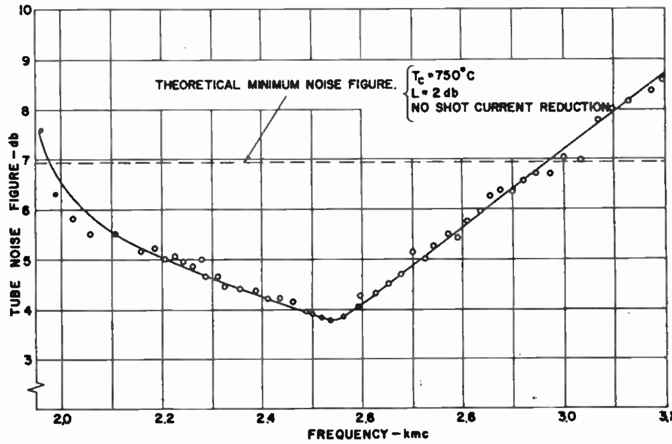

Fig. 7—This curve summarizes the most important results of the investigation. The noise figure of Tube A is shown as a function of frequency with the gain held constant at 20 db. The noise figure was optimized at 2.53 kmc and the gun potentials were not programmed as the pass band of the tube was tuned. The loss configuration for this curve corresponds to Case 3 of Fig. 12.

that the backward-wave amplifier is capable of very-low noise figures and therefore is a promising new type of microwave preamplifier. A tube noise figure of less than 6 db was measured over a tuning range approaching 30 per cent of the midrange frequency. A *maximum* tube noise figure of 4.5 db was obtained over a 10 per cent tuning range.

It was found that the dip in the curve could be translated somewhat by optimizing the noise figure at some other frequency. For example, if optimization were carried out at 2.8 kmc, the noise figure at that point might drop from 5.6 db to perhaps 5 db. However, this would bring up the noise figure at 2.5 kmc and the average noise figure of the over-all curve would be higher. Optimum performance was invariably obtained for initial adjustment around 2.5 kmc. Recent data indicate that a simple type of programming of the third and fourth anodes might increase the frequency range for a maximum allowable noise figure. This would be accomplished simply by connecting these electrodes to a voltage divider across the beam-voltage power supply, resulting in a linear increase of anode potentials with beam voltage.

Of considerable fundamental interest is the extremely low noise figure measured at the center of the tuning range. The lowest value is 3.7 db and is the lowest noise figure obtained to date on any type of microwave tube. It is about 3 db less than the theoretically predicted minimum noise figure of 6.9 db. This minimum value was calculated under the usual assumption of full uncorrelated shot noise at the initial excitation plane of the noise space-charge waves in the vicinity of the potential minimum.[9] It also assumes a cold insertion loss of 2 db for the input helix and a cathode temperature of 750°C. The factors which could, within the limits of a one-dimensional theory, account for the very low measured noise figure are a reduction in the magnitude of

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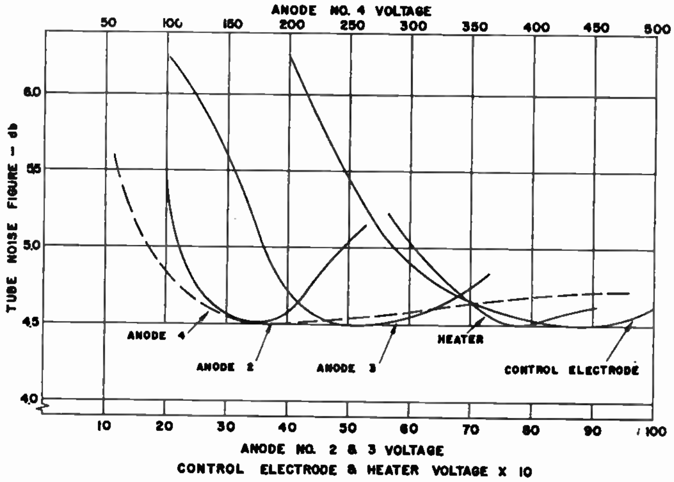

Fig. 8—Curves showing the optimization of noise figure with respect to the potentials of the various electrodes of the "Christmas-tree" gun. The frequency is 2.7 kmc. The potential of anode 1 is of the order of several volts and varies with gain.

either shot current or kinetic noise voltage (*i.e.*, velocity fluctuations), or the introduction of statistical correlation between these quantities at the excitation plane, or a combination of these factors. Indeed, based on detailed experiments on the Christmas-tree gun at the optimum point of operation, there are arguments to support all of these possibilities as well as some others. This phase of the investigation is continuing and will be reported at a later time.

Typical optimization curves are shown in Fig. 8. The center frequency in this case is about 2.7 kmc. The critical phase shift through the transducer, required for minimum noise figure, is established primarily in the low voltage section of the gun, as evidenced by the fact that the noise figure is successively more sensitive to the potentials of the fourth, third, and second anodes. Beyond the first anode, therefore, the gun behaves similarly to the multiregion guns used in conventional traveling-wave tubes.[11] It was often found that more than one set of anode potentials would minimize the noise figure. That is, if the potential of one of the anodes is varied (within limits) the gun is sufficiently flexible to set up a compensating phase shift and to establish essentially the same over-all transformation of the space-charge waves.

Minimum noise figure is dependent on a fine balance of potentials in the noise excitation or "input" region of this gun, consisting of the cathode, control electrode, and first accelerating anode. The latter electrode typically operates below 5 volts. Since it exerts primary control of the beam current it is initially adjusted to a particular value. The heater power and potential of the control electrode must then be adjusted in order to achieve the unique balance needed to optimize noise figure.

It is regarded as highly significant that the control electrode operates at a higher potential than that of the first anode. This establishes a saddle point of potential in the cathode-anode region and allows the beam to drift at very low velocity. It is expected that an analysis of this region will show an enhancement of correlation between the basic noise quantities in the beam and hence a considerable reduction in the minimum noise figure which can be theoretically predicted.

The *relative* variation of noise figure with frequency shows good qualitative agreement with theory. Table I shows calculated values of the space-charge parameter (QC), optimum noise power swr (η_{opt}) and angular distance between the noise standing wave and the circuit (ψ_{opt}) at several frequencies.

TABLE I

	2.0 kmc	2.5 kmc	3.0 kmc
QC	0.4	0.16	0.08
η_{opt}	1.8	2.7	4
ψ_{opt}	$-116°$	$-60°$	$-40°$

Although the magnitude of the angular distance ψ should ideally *decrease* with increased frequency, the increasing plasma wavelength moves the standing-wave minimum away from the gun and *increases* the magnitude of ψ. Thus, it can be assumed that if the correct conditions are established at 2.5 kmc this angle is close to its least desirable values at 2.0 kmc and 3.0 kmc. The variation of noise figure with respect to its minimum value can then be estimated.[9] This calculation, which is independent of initial noise quantities,[15] predicts an increase in noise figure of 2 db at 2.0 kmc and 5.3 db at 3.0 kmc as compared with the measured values of 2.3 db and 3.4 db, respectively. Except for the fact that the measured minimum noise figure was about 3 db less than the theoretical value, the variation of noise figure vs frequency is therefore approximately what was estimated and is believed to be typical of the characteristics to be expected from low-noise backward-wave amplifiers employing single-filar helix circuits. The low-noise tuning range could be increased by choosing a circuit (*e.g.*, the bifilar helix)[16] whose impedance (and hence QC, η_{opt}, and ψ_{opt}) is more nearly constant with frequency.

Although it was possible to focus all of the experimental tubes satisfactorily at magnetic field strengths of only several hundred Gauss, much higher fields were required for optimum low-noise performance. The measurements reported in this paper were taken at about 1300 Gauss. At 700 Gauss the noise figure typically deteriorated by about one db and above 1000 Gauss only slight improvement was noted. Under all of these conditions the total current interception was less than 0.1 per cent. It is not known whether this means that a high magnetic field is *inherently* necessary for very-low noise

[15] S. Bloom, "The effect of initial noise current and velocity correlation on the noise figure of traveling-wave tubes," *RCA Rev.*, vol. 16, pp. 179–196; June, 1955.

[16] P. K. Tien, "Bifilar helix for backward-wave oscillators," PROC. IRE, vol. 42, pp. 1137–1143; July, 1954.

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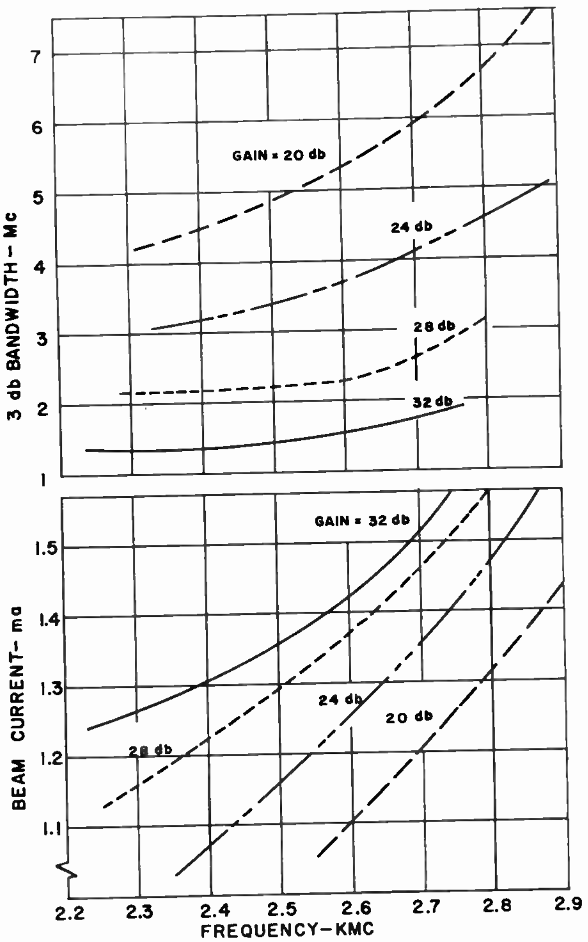

Fig. 9—Average gain-bandwidth characteristics of Tube A.

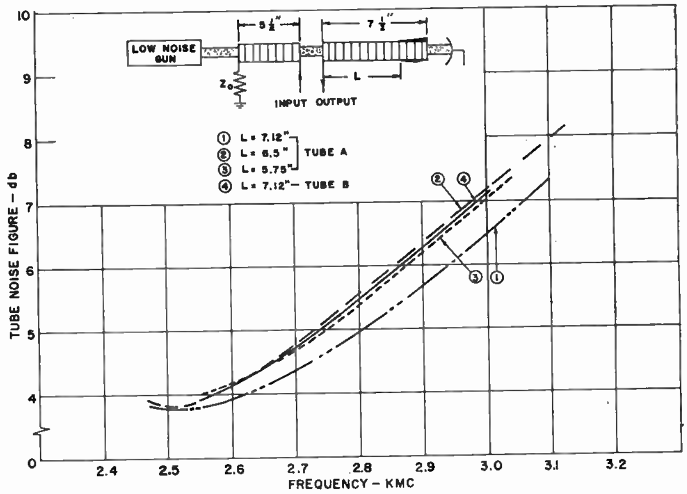

Fig. 10—Measured noise figure as a function of frequency for various effective lengths of the output helix. Gain is held constant at 20 db. As L increases, beam current is reduced and bandwidth decreases. Curve No. 4 was obtained on a second tube and illustrates the reproducibility of the results.

operation in this type of tube. In view of the extremely close spacing between the beam and the circuit, it is probable that better electron optics will reduce the magnetic field requirements.

In all cases the noise figure was very sensitive to reflected primary electrons from the collector, which are focused back to the gun and add a very noticeable component of shot noise to the beam. These effects were minimized in the measurements by biasing the collector to a value greater than 1000 volts positive with respect to the circuit. This problem also exists in conventional traveling-wave tubes but is more critical in backward-wave amplifiers because of the proximity of the beam to the circuit and the rapid radial variation of the rf field. In practice such effects can be reduced by coating the collector and distorting the magnetic field so as to trap the reflected primaries.[17] These improvements may well have the additional effect of reducing the necessary focusing field.

Composite gain and bandwidth curves for Tube A are shown in Fig. 9. Fluctuations due to matching have been smoothed out. These curves are typical and illustrate the nature of programming with frequency which probably would be necessary in some receiver applications. In order to maintain constant gain the beam current must be increased with frequency by varying the potential of the first anode. This mode of operation is seen to result in an increasing bandwidth-vs-frequency characteristic. On the other hand, programming for constant bandwidth results in an increasing gain-vs-frequency characteristic. Since the noise figure of Tube A was relatively insensitive to gain in the range from 20 to 30 db, the noise figures corresponding to various operating points on the gain-bandwidth curves are therefore all close to those plotted in Fig. 7. The above characteristics pertain to one specific tube; a reasonable range of bandwidths (*e.g.*, at 20-db gain) can be achieved by appropriate modifications in tube design.[18]

At a given value of beam voltage the noise figure was found to be essentially constant within the sharply peaked response curve of the tube. This is in agreement with theoretical calculations.

Fig. 10 summarizes the data obtained by varying the effective length of the output circuit. As indicated in the schematic drawing of the tube, this was accomplished simply by changing the length of the termination at the collector end. There are two general effects which result from a decrease in length of the output helix. First, the start-oscillation current of this section is increased. The beam current must then be increased to obtain a particular over-all gain. Second, because the increased current permits the input section to operate closer to its own oscillation threshold, the electronic gain of this section is higher relative to the total gain of the tube. Another incidental effect is an increased gain-bandwidth product at higher currents.

[17] R. W. Peter and J. A. Ruetz, "Influence of secondary electrons on noise factor and stability of traveling-wave tubes," *RCA Rev.*, vol. 14, pp. 441–452; September, 1953.

[18] M. R. Currie and D. C. Forster, "The gain and bandwidth characteristics of backward-wave amplifiers," IRE TRANS., vol. ED-4, pp. 24–34; January, 1957.

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For the three lengths of output helix shown in Fig. 10 ($L=7.12$ inches, 6.5 inches, and 5.75 inches) the nominal values of beam current at 20-db gain and 2.5 kmc were 0.8 ma, 1.2 ma, and about 2 ma, respectively. The minimum attainable noise figure (at 2.5 kmc) is seen to increase somewhat with increased beam current. Moreover, optimization was invariably simpler and more easily reproduced in Case No. 1 than for the other cases, and the noise performance with frequency was improved. Two possible explanations for this behavior are as follows: 1) The start-oscillation current of the input circuit should be optimally about twice that of the output circuit and/or 2) optimum conditions in the immediate vicinity of the cathode (*i.e.*, the noise input region) can inherently be attained only at very low currents for this particular gun geometry.

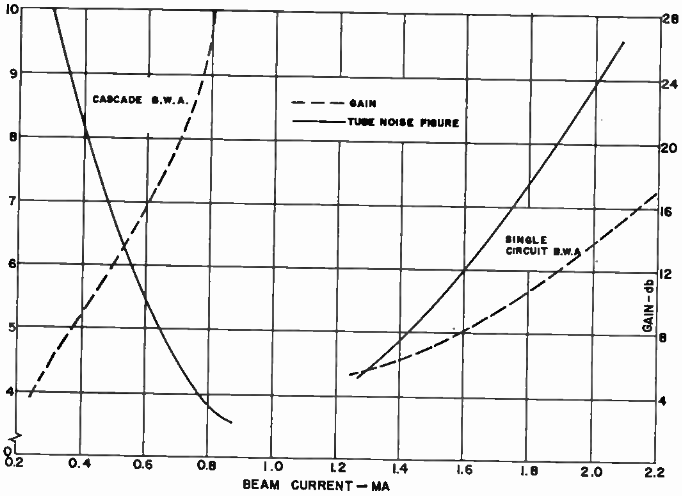

Fig. 11—Gain and noise figure as a function of beam current for 1) the input helix of Tube A acting as a single-circuit backward-wave amplifier and 2) for Tube A acting as a complete cascade amplifier.

The above conclusions are also indicated by the typical data of Fig. 11. Here, the noise figure and gain of the input helix operated as a single-circuit backward-wave amplifier by itself are shown as a function of current along with similar curves for the over-all tube. As mentioned earlier, the noise figure of the single-circuit amplifier typically decreases at low values of gain. An opposite characteristic was obtained on the cascade configuration. Minimum noise figure is seen to occur in the vicinity of 1-ma beam current; here, the gain of the input circuit is very low. At lower currents there is insufficient gain in this section to exert any control of the signal-to-noise on the beam as it enters the output circuit. These data substantiate the design philosophy used for the cascade low-noise tubes and agree with the tentative conclusions of the preceding paragraph.

The purpose of Tube B was to determine the reproducibility of low-noise performance from tube to tube. A modified spacing between the third and fourth anodes of Tube B introduced a serious lens effect which precluded obtaining smooth data as a function of frequency. However, the fourth curve of Fig. 10 represents an average of the data and shows a noise performance essentially equivalent to that of Tube A. It is concluded that the very-low noise figures of this type of tube can be easily reproduced.

Upon examination of the data in Figs. 7 and 10 a question naturally arises as to why the minimum attainable noise figures on Tubes A and B always occurred at 2.53 kmc and whether or not this optimum frequency can be shifted. A simple explanation, consistent with all of the other data, is the following: Because of a still undetermined fundamental mechanism in the cathode region of the gun, maximum noise reduction occurs for a unique set of potentials on the control electrode and first anode. This not only establishes an optimum range of beam current but also limits the phase shift of the noise waves between the cathode and the circuit to a narrow range of values (since, as discussed previously, the low voltage section of the gun exerts primary control on this phase shift). Thus, for optimum noise-wave excitation at the input plane, the swr (η) and phase (ψ) of the standing wave of noise power at the circuit entrance are essentially determined. The *optimum* values of these parameters are rapidly varying functions of the space-charge parameter QC which, in turn, varies rapidly with frequency in a backward-wave tube. Therefore, the minimum attainable noise figure will occur at that frequency (corresponding to a unique value of QC) for which the theoretical values of η_{opt} and ψ_{opt} most nearly coincide with the values of η and ψ produced by the gun.

This explanation was tested experimentally by scaling the mean cathode diameter so that the value of QC (about 0.16) obtained on Tube A at 2.53 kmc would occur at a new frequency. Table II lists the changes in cathode dimensions.

TABLE II

	Tubes A and B	Tube C	Tube D
Cathode O.D.	0.350 inch	0.370 inch	0.340 inch
Cathode I.D.	0.320 inch	0.340 inch	0.310 inch

All other dimensions of the tubes were maintained constant and, in each case, the gun electrodes were disposed symmetrically about the cathode.

At a given frequency, the value of QC for Tube C is considerably less than that of Tube A because the beam is much closer to the helix (high impedance), and vice versa for Tube D. Accordingly, it was expected that the frequency of minimum noise figure would be decreased for Tube C and increased for Tube D. This behavior is clearly indicated by the experimental data shown in Fig. 12. All of the curves are for 20-db gain. The optimum frequency of Tube C was not reached because of the lack of an appropriate local oscillator at the time of measurement; however, it clearly lies below that of Tube

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A. The minimum noise figure of Tube D has been successfully scaled upwards in frequency by about 10 per cent.

The above experiments illustrate a convenient means of trimming the design of low-noise backward-wave amplifiers so as to produce minimum realizable noise figure in the particular tuning range of interest in a receiver application. The results also demonstrate the flexibility of this type of tube as a noise detector in fundamental investigations of noise in electron beams, *i.e.*, the pass band is tuned for the minimum attainable noise figure which can, in turn, be related to the inherent noisiness of the beam.

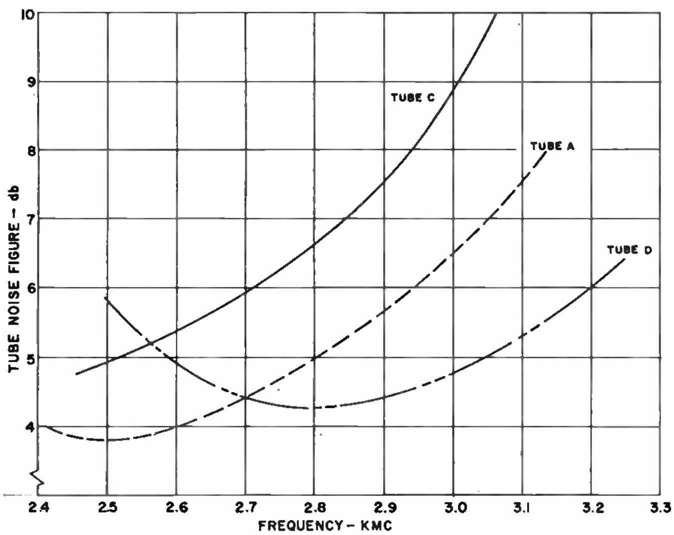

Fig. 12—These experimental curves were obtained from three different tubes and illustrate how the point of minimum noise figure can be shifted in frequency by changing the mean cathode diameter. The gain for all three curves is 20 db.

Conclusion

The tubes described are initial research models. Further development will be directed toward minimizing the magnetic field and improving mechanical tolerances so as to eliminate any fluctuations in gain as the tubes are tuned. The reproducibility of very-low noise figures on the first few tubes is encouraging. The life of these tubes should be long because of the small current density at the cathode. One model was operated for several hundred hours with no signs of deterioration.

Some of the applications of the low noise backward-wave amplifier as a receiver component remain to be investigated experimentally. As mentioned previously, its natural application is in simple trf-type receivers, although it also offers additional versatility to conventional superheterodyne receivers. Its other characteristics may also prove to be advantageous from a systems viewpoint. For example, when the beam current is gated off, the tube acts as an isolator with almost arbitrarily high attenuation. This suggests applications in very fast-acting duplexing.

The possibilities of scaling the very-low noise figures to other frequency bands are believed to be very good, as indicated by results on a slightly scaled tube (shown at the top of Fig. 4). The experiments have shown that, beyond the first anode, the noise transducer and circuit regions behave essentially as predicted by the one-dimensional theory. This part of the tube can certainly be scaled. However, since the assumptions of the usual theory are not valid in the cathode region, the fundamental mechanism of noise reduction here must be more completely understood before the scaling characteristics can be predicted in detail. Research on this problem is being continued.

The experiments described strongly suggest that further investigation might well lead to still lower noise figures for beam-type microwave amplifiers. There have been occasional measurements of lower noise figures on the research tubes but these results have not been highly reproducible. It is also improbable that these tubes were adjusted precisely for minimum noise because of the large number of interdependent parameters (*i.e.*, potentials).

Reduction of the cold insertion loss of the circuit and refinement of cathode and processing techniques will improve the noise performance. More important, none of the really critical dimensions of the Christmas-tree gun have been systematically varied to optimize the noise reduction in the cathode or noise excitation region. It would be a remarkable coincidence indeed if optimum geometry were achieved in the first design.

It is evident that the unique noise reduction produced by the special geometrical and potential configuration in the cathode region is a first-order effect. All of our investigations to date indicate that the gun employed in these experiments is merely one embodiment of a general means for effecting a basically new excitation mechanism of the space charge waves in the vicinity of the cathode. For example, although the Christmas-tree structure is convenient for use with large diameter beams it is not essential to the gun's performance. When eliminated, an annular beam originating predominantly from the outer edge of the cathode results. Thus, annular beam guns of this general type can be applied to conventional traveling-wave tubes. The parameters relating to the noise pattern on the beam (η and ψ) remain much more constant with frequency for this type of tube. Therefore, having adjusted for the very-low noise figure which can be attained with this type of gun, it will then be possible to maintain this low noise figure over a wide frequency band.

In summary, a new type of low noise amplifier has been demonstrated which may satisfy the increasingly apparent need for rapidly tunable selective preamplification at microwave frequencies. Also, based on the results of this investigation, we believe that the future will see microwave tubes with significantly lower noise figures than have been attained to date.

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Atmospheric Noise Interference to Short-Wave Broadcasting*

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Summary—In order to determine the different parameters necessary for assessing the interfering effect of atmospheric noise to short-wave broadcasting, a systematic physical analysis is made of how the atmospheric noise impulse, as *heard* by the ear, arises and how it causes annoyance to the listener of broadcast programs. Hence, criteria are developed both for measurement and estimation of atmospheric noise. The paper thus provides the necessary additional physical background for the author's papers on "Measurement of atmospheric noise interference to broadcasting" and "Noise power radiated by tropical thunderstorms" but leaves the final conclusions and results of the two papers unchanged. Although the paper is thus restricted in scope, it is believed that the general principles emerging from the discussion should be of wider application.

I. Introduction

A METHOD[1] for the measurement of atmospheric noise interference to broadcasting has been reported. The method was evolved on the basis of the results of extensive, long, and tedious experiments. Some of the criteria adopted in the paper like the choice of bandwidth, the time constants of the noise meter, the frequency response characteristics of the output unit, the method of calibration, etc., require a proper physical explanation.

The noise power[2] radiated by tropical thunderstorms has been calculated. To bring out the full significance of each stage of this calculation, it is necessary to give an integrated physical picture of how the acoustic impulse as heard by the ear arises from the radiations that originate from the electrical discharges associated with lightning flashes. As a result of experimental and theoretical investigations, now it is possible to eliminate the defects enumerated above and that is the principal object of this paper.

As the physical principles involved in the subject matter of this paper are difficult to visualize, they are described in some detail. Further, extremely simple mathematical methods have been adopted to add to the clarity of the description. By restricting the discussion to the short wave band *viz.*, 2.5–20 mc. only, we are required to consider only one type of lightning discharge and this makes the exposition clearer.

II. Analysis of the Problem

The subject matter of the paper is presented in the following order. It is shown first that atmospheric noise is a statistical phenomenon and, therefore, any simplified discussion is possible only by considering the idealized statistically valid representation. A brief description is given of the phenomenon of lightning discharge, and the numerical parameters associated with the type of discharge of interest to this paper are then reproduced. A time plot of the radiation arising from a typical flash is deduced and what a broadcast receiver is expected to pick up is discussed.

In order to develop further the subject matter, there follows a detailed discussion of the characteristics of the ear and the ordinary broadcast receiver. The conclusions from such discussions are utilized to develop the concept of the acoustic impulse that arises from a lightning flash and is heard by the ear. From this concept, the essential requirements of an objective noise meter are enumerated. A brief description is given of how the noise values, as deduced from the data collected experimentally, are to be estimated. It is shown that, for purposes of estimation, the only quantity to be deduced from lightning discharge data is the equivalent radiated power corresponding to the acoustic impulse as heard by the ear. This power is calculated next on the basis of the conclusions of the previous sections. Finally, a concluding section briefly reviews the entire problem.

III. Statistical Nature of Atmospheric Noise

The source of all atmospheric radio noise appears to be the natural electrical discharges associated with thunderstorms. The number of thunderstorms occurring in a region during a season varies from year to year. The growth and decay of such thunderstorms during a day show variations from thunderstorm to thunderstorm. The different parameters associated with the electrical discharges accompanying a thunderstorm vary from discharge to discharge. Therefore, atmospheric noise arises from a natural phenomenon which shows statistical variations. Hence, noise levels at a place can only be deduced by collecting and assessing data on a statistical basis. Similarly, the estimation of such noise levels must have a statistical basis. If the measured values are to agree with the estimates, the statistical basis for both must be identical.

Owing to the statistical nature of the phenomenon, an individual thunderstorm, an individual electrical discharge or an individual impulse as heard by the ear has no significance. There are various ways of dealing with statistical phenomena. In this paper, we choose one of the simplest of these. A physical parameter which shows statistical variations has always a median value. Sup-

* Original manuscript received by the IRE, September 3, 1957.
† L.D. College of Engineering, Ahmedabad. 9, India.
[1] S. V. C. Aiya, "Measurement of atmospheric noise interference to broadcasting," *J. Atmos. Terrest. Phys.*, vol. 5, pp. 230–242; September, 1954.
[2] S. V. C. Aiya, "Noise power radiated by tropical thunderstorms," Proc. IRE, vol. 43, pp. 966–974; August, 1955.

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pose we build up a physical picture of a phenomenon by assigning median values for each of the physical parameters associated with the phenomenon. Then, this physical picture will be regarded as an idealized statistically valid representation of the phenomenon and we will utilize this representation for discussing any typical case. The adjective, "typical," when used in the rest of the paper stands for "the idealized statistically valid representation" or its equivalent. The physical analysis that follows is based on a discussion of such "typical" cases.

IV. The Lightning Discharge

A thunderstorm is a localized thermodynamical process in the atmosphere. It is invariably accompanied by electrical discharges. Such discharges occur within the cloud, from the cloud into the air, and from the cloud to ground. These discharges are complicated physical phenomena and show differences in characteristics depending on the type of discharge, etc. Changes of current occurring in the discharging and charging processes give rise to radiation. The frequency distribution of energy radiated depends on the characteristics of a discharge, etc. Therefore, different types of discharge are responsible for radiation that contributes significantly to noise in different frequency bands.[3,4] One type of discharge, *viz.*, the discharge within the cloud, is principally responsible for noise in the short-wave band (2.5–20 mc). We are concerned with this discharge only for purposes of this paper.

This particular type of discharge has been discussed in detail.[2] Certain numerical values have been assigned to the different parameters associated with it. These have all been re-examined[4] on the basis of the recently published literature. It is found that only one quantity has to be changed and that is the median value of the number of strokes per flash. This was taken as three but it should be four as given by Schonland.[5] This change is incorporated in this paper without further discussion.

Discharges within the cloud are the most common types of electrical discharges associated with tropical thunderstorms. Also, they are quite frequent at higher latitudes. Besides, frequently there is a discharge within the cloud even in the case of discharges striking the ground. The height of the cloud base is generally about 2 km above ground level[6] and, therefore, the discharges in this case can be considered to occur practically in free space. Since we are concerned with such discharges in this paper, some essential details pertaining to them are given in the next section.

[3] S. V. C. Aiya, "Noise radiation from tropical thunderstorms in the standard broadcast band," *Nature*, vol. 178, p. 1249; December, 1956.

[4] S. V. C. Aiya, "Atmospheric noise radiators," to be published.

[5] B. F. J. Schonland, "The lightning discharge," "Handbuch der Physik," S. Flügge Marburg, ed., Springer-Verlag, Berlin, Germany, vol. 22, pp. 576–628; 1956.

[6] D. J. Malan and B. F. Schonland, "The distribution of electricity in thunderclouds," *Proc. Roy. Soc., A.*, vol. 209, pp. 158–177; October, 1951.

V. Electrical Discharges Within the Cloud

When the electric field at some point in the cloud exceeds the disruptive strength of the dielectric, a discharge occurs and this leads to the initiation of a lightning flash. The flash is intermittent and consists of a number of strokes. Each stroke is made up of a leader stroke and a recoil of low intensity and long duration. The recoil is of no significance for radiation in the short-wave band. The leader stroke is practically vertical and is not continuous. It consists of a large number of steps. The time duration of each step is extremely short but the time interval between steps is much longer. The steps are approximately vertical and of short length so that they can be considered as equivalent to short dipoles in free space. When a discharge occurs in a step, there is growth of current and the rate of change of this current is responsible for radiation. Obviously, the maximum rate of change of current gives rise to the peak electric fields.

An idealized statistically valid representation of such a flash has been given[2] and the statistical median values assigned to the different parameters enumerated above are given in Table I.

TABLE I

Data For A Typical Cloud Discharge

T_f	= over-all duration of a flash = 0.2 second
n	= number of strokes per flash
	= number of stepped leaders per flash = 4
T_s	= time interval between strokes in a flash
	= time interval between stepped leaders = 0.04 second
T_L	= over-all duration of a stepped leader = 0.001 second
τ	= time duration of a step = less than a microsecond
T_i	= time interval between steps = 0.000074 second
ν	= recurrence frequency of steps = 13,500 cps
X	= max. rate of change of current in a step
	= 10^{10} amperes per second
l	= length of a step in a stepped leader = 67 meters

In each step, there is a discharge and the current changes give rise to the radiation of an impulse. A series of such impulses arise from each stepped leader. As there are four such leader strokes in each flash, four trains of impulses are radiated by a flash. Assigning numerical values as given in Table I, the complete time plot of the radiation arising from a typical cloud discharge is given in Fig. 1. Fig. 1 gives the correct physical representation of the radiation from a typical cloud discharge as deduced from the available data on lightning discharges.

A Fourier analysis of an impulse shows that it consists of a large number of components of different amplitudes and frequencies. Radio waves of different frequencies display different propagation characteristics; this applies equally to the different Fourier components of the impulses radiated by lightning flashes. Therefore, these different Fourier components travel via the ground, via the ionosphere, via the troposphere, or as an optical ray in exactly the same manner as other radio waves of corresponding frequencies.

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From the standpoint of noise, we are concerned with only those Fourier components which, after being received on an aerial, get through into the receiver. An ordinary receiver tuned to a frequency, f, picks up all the Fourier components which lie within the bandwidth, B, of the receiver. B/f is generally small. Hence, we are concerned with only the Fourier components, which lie within a narrow band, B, round about f. The amplitude of the Fourier components varies with frequency but, for the narrow band we are concerned with, we can consider the amplitude constant. Further, for all these Fourier components in this narrow band, we can regard the propagation characteristics to be identical with that of frequency, f.

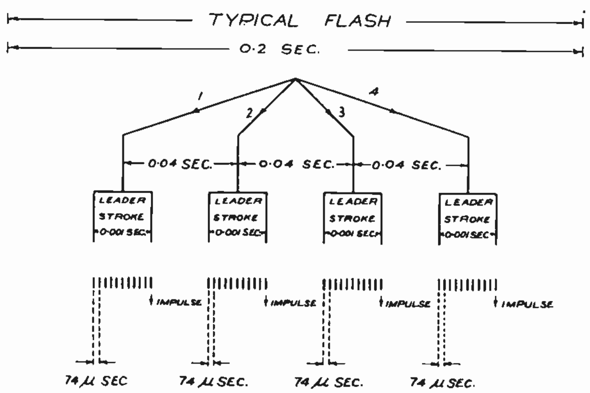

Fig. 1—Radiations from a typical flash in a cloud discharge.

Therefore, when viewed from the standpoint of the receiver, the impulse consists of a number of Fourier components lying within a bandwidth, B, at a frequency, f, and the amplitudes of these Fourier components are approximately equal. This impulse is thus different from the impulse radiated by the source.

Further analysis of the problem leading to the acoustic impulse has to take account of the effect of the receiver and the ear. Therefore, the characteristics of a broadcast receiver and the average human ear are given in the sections to follow. As some facts relating to the ear have a bearing on the effect of the receiver, a description of the ear is given first.

VI. The Typical Ear

The characteristics of the ear have a bearing on our evaluation of the interfering effect of atmospheric noise which is essentially impulsive noise. In the subsections to follow, the different aspects of the problem are discussed to the extent necessary for purposes of this paper.

A. Concept of an Impulse

An impulse may last a fraction of a microsecond, a few milliseconds, and so on. These are mathematical possibilities. But, when we think of an acoustic impulse, we have to take account of the behavior of the ear. The ear operates as an integrating device. Stuedel[7] has observed that the apparent loudness of an impulse as assessed by the ear is determined by the pressure integrated over a period of 0.0003 second in the region of its peak value. Therefore, the correct measure of what the ear notices is the average value of the amplitude over a period of 0.0003 second. The result of Stuedel quoted is for loudness greater than 50 phons. The conclusions of Stuedel's investigations, etc. have been more generally put by Davis[8] as follows: "The response of the ear to single impulsive sounds appears to depend mainly upon the maximum impulses which the ear experiences in rather less than a thousandth of a second."

The conclusions of Davis appear to be more representative of a listener's experience. We conclude, therefore, that the average value of the amplitude of sound impulse over a period 0.001 second gives the proper measure of the apparent loudness of an impulse as assessed by the ear.

We will now consider the implications of what has been said for purposes of this paper. A lightning discharge radiates four trains of impulses per flash. Each train has a number of impulses lasting a fraction of a microsecond and the over-all duration of the train of impulses is 0.001 second. From the acoustical point of view, each one of these individual impulses is of no significance. What matters is the average amplitude due to one complete train which lasts 0.001 second. That is, we are only concerned with the average value of the amplitude arising from a complete stepped leader.

B. Response of the Ear to Impulsive Noise

The characteristics of the ear for judging impulsive noise have been investigated. They are discussed by Davis[8], and the important conclusions are as follows:

1) The ear responds to different magnitudes of impulses in a logarithmic manner.
2) For recurring impulses, the ear judges the intensity in a cumulative manner, *i.e.*, subsequent repetitions of an impulse would add to the loudness as heard by the ear.
3) Between impulses, there is always some leakage of intensity of sound as heard by the ear.
4) Impulses separated by more than a second do not have any cumulative effect, *i.e.*, there is almost complete leakage of sound intensity in about a second
5) The full strength of any sound is judged in about 0.2 second

The conclusions of Davis are simply but more quantitatively incorporated in the following results of the author:[1]

[7] U. Stuedel, "The sensation of loudness and its measurement," *Hochfrequenztech. u. Electroacoust.*, vol. 41, pp. 116–128; 1933.

[8] A. H. Davis, "An objective noise meter for the measurement of moderate and loud, steady and impulsive noises," *J. IEE*, vol. 83, pp. 249–260; 1938.

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a) The charging time constant of the ear is 10 milliseconds.
b) The discharging time constant of the ear is 500 milliseconds.

From b), it follows that the leakage of sound in the ear is such that its value falls to 10 per cent of the original in 1.15 seconds and this agrees with 4). It further follows from b) that sound as heard by the ear remains above $\frac{2}{3}$ its original value for about 0.2 second. This carries the implicit meaning of 5). Further 1), 2), and 3) are automatically included in a) and b) but the latter have the advantage of being quantitative and much more explicit.

Now, a lightning flash lasts 0.2 second and consists of four trains of impulses, each train lasting 1 millisecond. As explained in Section VI-A, a train of impulses becomes equal to one effective impulse with an average value of the amplitude. Four *such* impulses will have a cumulative effect as the ear judges the full strength of sound in 0.2 second and the cumulative effect of the four has to be obtained by taking note of the fact that each of these four equivalent impulses lasts 0.001 second and that the charging and discharging time constants of the ear are as given under a) and b). This is a very important result that has to be incorporated into calculations, and hence, estimations.

Turning now to any technique of measurement, the output unit which is used for taking readings in any noise meter must have charging and discharging time constants as given in a) and b) if the noise meter is to indicate what the ear feels.

Summarizing, we may say that the *acoustic* impulse as heard by the ear arises from one complete flash and is the integrated effect of the four trains of impulses radiated during a flash.

C. Sensitivity of the Ear

The sensitivity of the ear depends on frequency and is maximum in the frequency range, 1000–3000 cps. It is, therefore, extremely important that the Fourier components of noise within this frequency range are properly reproduced. This fact indicates that, in choosing bandwidths for the rf side of receivers, the choice of 3 kc each side of the carrier is most desirable.

D. Noise Interference

In a study of the interfering effect of noise, interest is centered on the investigation of noise in the presence of signal. In such cases, the noise wave form gets superposed on the signal wave form. Subjective tests have revealed that, for a normal ear, the smallest difference of sound intensity than can be detected is about 10 per cent of the original sound intensity at small volumes and about 25 per cent for large volumes.[9] Therefore, for the first case,

1) Power of a signal that is audible
$= E$ watts
2) Next level of power giving a sound audibly greater than in a)
$= (1.10)\ E$ watts
3) The next level
$= (121)\ E$ watts, and so on.

From 1) and 2), it follows that if the signal amplitude is over 20 db above noise amplitude, noise will not be noticed by the ear. This is an important conclusion for evaluating the standards of satisfactory service.

From 2) and 3), it follows that the ear notices the first increase of noise when noise power has changed from 0.10 E to 0.21 E. This shows that a normal ear notices increase of noise in the presence of signal in steps of 3 db. (The latter will be 3.5 db when we take the large volume criterion.) The ear does not notice changes in noise when it is 3 db either way about a mean value. It follows that variations of noise amplitude within 6 db from a maximum value can be considered constant for purposes of judgment by the ear. This, we will call the 6-db criterion.

Examining the above conclusion with that of Section VI-C, we conclude that a bandwidth of 6 kc at 6 db down is suitable for a receiver employed as a noise meter. Further, for relating the measured values with estimates, no correction for the 6 db down criterion is necessary in calculations employed for estimations of noise.

We have considered so far the way in which the characteristics of the ear affect the design of a noise meter and the calculations involved in noise estimations. We have to remember also that the noise heard by a listener is what gets out of the loudspeaker of a receiver. To this extent, we have to examine how an ordinary receiver affects the noise impulse in its passage through the different stages of a receiver. This is examined in Section VII.

VII. The Broadcast Receiver

The general design of receivers for amplitude modulated signals is now practically standardized. Such receivers are of the superheterodyne type and consist of: high-frequency stages made up of rf, mixer or converter and IF stages, a detector, af voltage and power amplifier stages and, an output transformer and a loudspeaker. The effect of each of these will now be discussed.

A. High-Frequency Stages

The principal point of interest is the fact that these stages employ tuned circuits. Increase in the number of rf or IF stages increases the number of tuned circuits. The effect of tuned circuits on impulses has been widely discussed. We will focus our attention to the specific case of the impulses as they arise from lightning flashes. A flash gives rise to four trains of impulses as indicated in Fig. 1. Each impulse lasts a fraction of a microsecond

[9] S. V. C. Aiya, C. G. Khot, K. R. Phadke, and C. K. Sane, "Tropical thunderstorms as noise radiators," *J. Sci. Ind. Research (India)*, vol. 14B, pp. 361–376; August, 1955.

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and this fact is important. Further, as discussed in the previous section, we are primarily concerned with the output wave form, etc. of the complete train of impulses. In such cases, it has been shown that:[10]

1) The time integral of the output envelope is independent of the damping and number of tuned circuits.
2) The output waveform is independent of the wave shape of the impulse if its duration is less than the order of the reciprocal of the bandwidth, but is dependent only on its time integral.

As discussed in Section VI, we require the time integral of the output envelope, and the bandwidth is of the order of 6 kc. It follows, therefore, that the high-frequency stages have no effect.

B. The Detector

Modern receivers employ diodes with excellent detection characteristics. In any first approximation, therefore, the effect of the detector can be ignored. However, a proper calculation has to take account of one aspect of detection and that is the efficiency of rectification of modulation, η, because it will be shown later that atmospheric noise impulse due to one leader stroke as a whole becomes equivalent to a 100 per cent modulated signal.

$$\eta = V_d/m \cdot V_c \tag{1}$$

where

m = depth of modulation or modulation factor,
V_d = peak modulation frequency voltage developed across the diode load,
V_c = peak amplitude of the carrier.

In actual designs, the value of η varies from 0.7 to 0.9 for all well designed receivers when moderate or large signals are considered. It can be higher for small signals. Ordinary reception corresponds to moderate or small signals when tuning in the short wave band. Laboratory designed and commercial receivers have been extensively used by the author and a value of 0.9 for η is found to be the best average for the conditions stated and this value will be used in this paper.

C. The AF Stages

For moderate and low volumes, the fidelity of the af stages is practically flat from 100 to 10,000 cps. Hence, they do not affect the wave form of the noise impulse as heard.

D. The Output Transformer and Loudspeaker

These are the limiting factors in the af performance characteristics of receivers. Cone type dynamic loudspeakers are ordinarily employed in receivers. For speakers of this type, the response curve falls off below 100 and above 4000 cps, both on the axis and at 30° off the axis. For even very good speakers, the fall is certainly rapid below 100 and above 5000 cps. Therefore, any output unit used in a noise meter must have corresponding characteristics, *i.e.*, flat frequency response from 100 to 5000 cps. This has actually been found, in practice, to be satisfactory.[11]

VIII. The Interfering Acoustic Impulse

For the particular type of lightning discharge with which this paper is concerned, the time plot of the actual radiation from a flash is given in Fig. 1. All the relevant details are given in the figure. It will be seen that there are four trains of impulses in a flash and the time duration of the impulses themselves is extremely small, being of the order of a fraction of a microsecond. It has been shown in Section VI-A, that for purposes of the apparent loudness as judged by the ear, we are concerned with the average value of the amplitude arising from one complete train of impulse, *i.e.*, the average amplitude due to a stroke in a flash. Therefore, from the acoustical point of view, the four trains of impulses of Fig. 1 are equivalent to four rectangular pulses of 0.001 second duration. This is shown in Fig. 2. We will now examine the implications of Fig. 2. From the standpoint of interfering effect, a typical flash gives rise to four rectangular pulses of 0.001 second duration at intervals of 0.040 second. A rectangular pulse of the type shown in Fig. 2 can be regarded approximately as corresponding to a 100 per cent modulated signal on the basis of our conceptions of amplitude modulation.

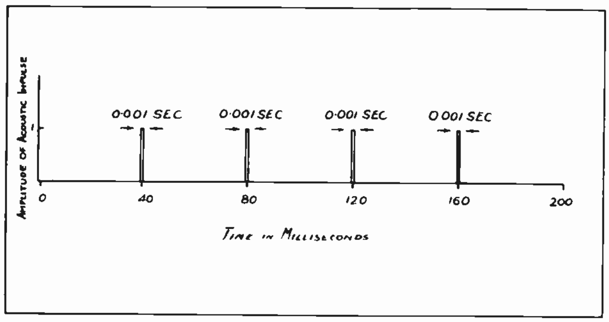

Fig. 2—Equivalent acoustic impulses arising from a typical flash.

The recurrence frequency of these rectangular pulses is 25 cps. If we regard this rectangular pulse of 0.001 second duration as corresponding roughly to half of a complete 100 per cent modulated envelope, it follows that the equivalent modulating frequencies probably correspond to 500 cps and its harmonics. These facts are suggestive of a possible approach to the design of noise suppressor circuits for being employed after the detector stage in receivers and will form the subject matter of a future communication.

[10] H. A. Thomas and R. E. Burgess, "Survey of Existing Information and Data on Radio Noise in the frequency Range, 1–30 mc," H.M. Stationary Office, London, Eng., Radio Res., Special Rep. No. 15, 1947.

[11] S. V. C. Aiya and K. R. Phadke, "Atmospheric noise interference to broadcasting in the 3 mc band at Poona," *J. Atmos. Terrest. Phys.*, vol. 7, pp. 254–277; October, 1955.

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For the calibration of noise meters in which measurements are carried out after detection, a suitable af modulating frequency has to be chosen. From the discussion in the previous paragraph, it follows that 500 cps is perhaps suitable. Theoretically and from the practical standpoint, there is not any significant difference between the use of 500 or 400 cps. Further, 400 cps is used as the modulating frequency in all receiver testing, etc. and there is provision for this in all standard signal generators. Therefore, the choice of 400 cps as the modulating frequency for signal generators employed for the calibration of noise meters appears to be a scientifically justifiable step.

It has been explained in Section VI-B, that the acoustic impulse as heard by the ear arises from one complete flash and is the integrated effect of the four rectangular pulses of Fig. 2. Such a cumulative effect has to be evaluated after taking into account the charging and discharging time constants of the ear. For this purpose, Fig. 3 has been drawn and it makes the exact effect clear.

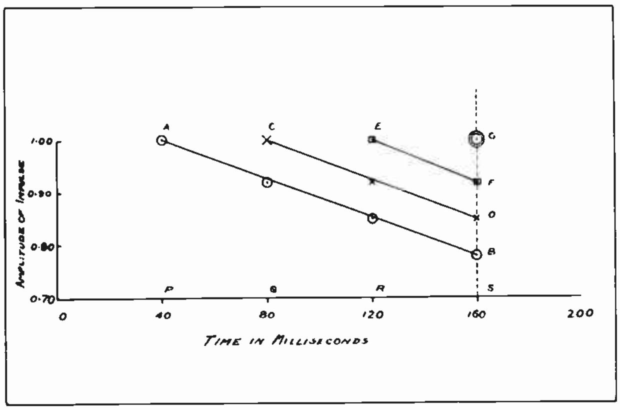

Fig. 3—Decay curves of the four acoustic impulses arising from a flash.

The four rectangular pulses of Fig. 2 arise at points P, Q, R, and S in Fig. 3. The first three pulses decay in magnitude as shown by lines AB, CD, and EF in Fig. 3. It is assumed that the discharging time constant of the ear is 0.5 second and that the interval between the pulses is 0.040 second. The ear judges the full strength of sound in about 0.2 second. It follows, therefore, that so far as the ear is concerned, the cumulative effect of the four pulses is felt at S in Fig. 3 and its magnitude is

$$0.78 + 0.85 + 0.92 + 1.00 = 3.55. \tag{2}$$

Now, suppose there is a carrier of voltage amplitude, A. Then, for a 100 per cent modulated signal, the af voltage amplitude would be A but having regard to the effect of the detector as explained in Section VII-B, it would be $(0.9)A$.

But, the pulses as shown in Fig. 2 last 0.001 second while the charging time constant of the ear is 0.01 second. Therefore, an af amplitude of $(0.9A$ would affect the ear only to the extent of $(0.09)A$.

Now, suppose the equivalent carrier amplitude of each pulse in Fig. 2 is A. Then, for purposes of the effect on the ear, each pulse becomes equivalent to one having an amplitude of $(0.09)A$. Further, since the four pulses produce a cumulative effect on the ear as indicated in (2), the amplitude of the equivalent carrier which gives rise to the acoustic impulse as it affects the ear is given by

$$B = (0.32)A. \tag{3}$$

Eq. (3) gives an extremely important result and it has to be examined in detail. If we examine Fig. 2, we find that there are four rectangular pulses, each of 0.001 second duration in a total period of 0.2 second which corresponds to the duration of the flash, *i.e.*, physically a carrier amplitude, A, corresponding to that of a pulse exists for only four milliseconds in a total period of 200 milliseconds. What is the equivalence of this from the standpoint of continuous waves? From the acoustic standpoint, the effect that is felt in a period of 0.2 second is equivalent to the effect of a continuous wave. Therefore, if we calibrate a noise meter using a continuous wave signal generator, the four pulses of amplitude, A, would give a reading corresponding to an amplitude, B, as recorded by a continuous wave signal generator. In other words, if we calculate "A" from lightning discharged data, we must reduce the value by multiplying by 0.32 to get "B" which corresponds to the amplitude of the equivalent acoustic impulse. Putting it in another way, we may say that we can calculate the field due to a stroke in a flash. Then, to get the field that corresponds to the acoustic impulse and hence to continuous wave signal generator calibration, we must multiply by 0.32 the field due to a stroke in a flash. It has to be emphasized that we are not interested in the electric field at a point due to a lightning discharge but in the value of the field corresponding to the equivalent acoustic impulse evaluated on the basis of continuous wave equivalence as broadcasting employs continuous waves. It has also to be pointed out that calculating the average value in 200 milliseconds of fields due to four one millisecond pulses without reference to the receiver and ear would be highly incorrect and would have no practical significance.

In the last paragraph we have discussed the acoustically averaged equivalent value of the carrier which lasts for only four milliseconds in 200 milliseconds and the acoustical averaging is done for 200 milliseconds in which the ear judges the full strength of sound. There is yet another aspect of the problem as examined from the af point of view only. If we regard each rectangular pulse of Fig. 2 as a 100 per cent modulated signal, we get an af amplitude of magnitude, A. But, because of the receiver and ear characteristics the effect felt by the ear corresponds to $(0.32)A$. That is, from the standpoint of continuous modulation of a continuous signal, the equivalent modulation is only about 32 per cent and *not* 100 per cent. Therefore, the level of modulation to be

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chosen at the time of calibration should be about 32 per cent for the standard signal generator. Since there is no essential difference between 32 and 30 per cent modulations in engineering evaluations and more particularly in statistical phenomena, the choice of 30 per cent for the level of modulation is scientifically justifiable.

Thus, there are two consequences of (3). The first corresponds to our deducing the equivalent magnitude of a continuous wave carrier corresponding to the four rectangular pulses. The second is for deducing the level of continuous wave modulation corresponding to the actual effect on the ear. It is difficult to elaborate in greater detail these *dual* consequences arising from the same equation but it is presumed that, with a little effort, it should not be difficult to visualize the same.

We can now deduce the requirements of an objective noise meter which measures the acoustic impulse as it arises from a typical flash, *i.e.*, a noise meter which measures atmospheric noise as a source of interference to broadcasting. This will be done in Section IX. Then follows an account of how the noise levels as deduced from measurements with such a noise meter can be estimated. Such a technique requires the value of noise power radiated by a typical flash but corresponding to the equivalent acoustic impulse as discussed in this section. This is deduced from the lightning discharge data.

IX. Requirements of an Objective Noise Meter

As a source of interference to broadcasting, we are interested, not in any general value of atmospheric noise, but in the acoustic impulse which arises from a flash. Therefore, measurements have to be carried out after detection, *i.e.*, on the af side. The other essential requirements follow automatically from the discussions in Sections VI, VII, and VIII. We must use a superheterodyne receiver having a bandwidth of 6 kc at 6 db down. The af output must be fed to an output unit having a charging and discharging time constant of 10 and 500 milliseconds respectively. The frequency response of the output unit must be flat from 100 to 5000 cps. As it is to record impulses, there must be some overload protection in the form of a logarithmic response.

The indicating meter to be used in the output unit must be such that its response should become relatively unimportant compared to the other time constants. But, if possible, it is desirable to have for the indicating meter, a time constant approximately equal to the time required for the ear to judge the full strength of sound, *i.e.*, 0.2 second. (The indicating meter used by the author and his collaborators in noise measurements had this time constant approximately.)

It follows, from Section VIII, that the noise meter must be calibrated by using continuous signals from a standard signal generator and that these signals must be modulated to a level of about 30 per cent by a note of frequency as near 500 cps as possible. The use of 30 per cent modulation by a 400 cps note as used in receiver testing, etc. is permissible.

It was noticed experimentally that ten acoustic impulses per minute have an annoyance value for the listener of broadcast programs.[1] Hence, the average value of the ten highest impulses per minute is taken as a measure of noise. It has not yet been possible to find a suitable explanation for this value of ten. Fortunately, however, there is a scientific co-relation from the natural phenomenon of lightning. It has been observed that all types of thunderstorms give rise to at least 10 flashes per minute during the period of their peak activity.[9] Further, when they radiate during peak activity more than 10 flashes per minute, the average of 5, 10, 15, or 20 impulses arising therefrom have about the same value when a large number of thunderstorms give rise to noise at a place and this frequently happens in the short-wave band.

The requirements of a noise meter have been deduced here from scientific considerations and all these are satisfied by the noise meter previously described.[1,11] Other details including the method of collection and assessment of data are given in the references quoted.

X. Noise Estimations

From the actual measurements with the noise meter described in Section IX, noise levels can be deduced in different ways. The most suitable way recommended is to give the median values of noise at a place for a particular sector of day during one season. We will refer to this quantity as noise level. It is not possible to deduce noise levels at all places from actual measurements. Therefore, a technique must be developed to estimate such noise levels at a place with reasonable accuracy. It is adequate if the technique gives estimates which agree with measured values within 3 db. A variation up to 6 db has to be tolerated in some cases.[11]

A technique for estimating the noise levels as measured has already been described[11] and it is found that there is very satisfactory agreement between estimates and measured values. A brief description of the technique is given below.

Atmospheric noise arises from lightning discharges associated with thunderstorms as explained in Sections IV and V. From the data collected by weather offices for years, the distribution of thunderstorm days over the globe is known.[12–14] The data is scanty for many regions and is not necessarily very reliable in all cases. But, they are reasonably adequate for dealing with the

[12] C. E. P. Brooks, "Distribution of Thunderstorms Over the Globe," Meteorological Office, London, Eng., Geophys. Memo. and Prof. Notes, No. 24; 1925.

[13] "World Distribution of Thunderstorm Days, Part I—Tables," World Meteorological Organisation, Geneva, Switzerland; 1953.

[14] S. V. C. Aiya, "Distribution of thunderstorm days on the land mass of India," *J. Sci. Ind. Research*, vol. 13A, pp. 314–317; July, 1954.

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statistical problem for many regions. From such data, the mean center of thunderstorm activity in a season responsible for noise at a place during a particular sector of day can be located. At this position, it is supposed that there is a short vertical dipole in free space radiating a carrier power of P watts and carrying a modulation by a 400 cps note at a level of 30 per cent. A step in a stepped leader which is of short length and high above the ground is responsible for radiation from a flash. Hence, the assumption of a short vertical dipole in free space is justified. The concept of 30 per cent modulation by a 400 cps note implies the carrying over of the calibration procedure dictated by the interfering acoustic impulse to the idealized radiator.

Using the above conception, the ground or sky wave calculations can be carried out as usual. Since we mostly deal with sky wave noise in the short wave band, we can restrict the discussion to sky waves. The unabsorbed field intensity, E, in $\mu V/m$, at point distant, r, in millions of meters fron the mean center, is given by

$$E = 212\sqrt{P/1000} \cdot \frac{\sin\theta}{r} \cdot \qquad (4)$$

θ is the angle the direction of radiation makes with the axis of the dipole. If there is absorption, similar calculations can be carried out by following the procedure given by CRPL.[15] If "P" is the statistical median value for power, the value obtained for "E" in (4) gives the estimate of noise level. Hence, the problem of estimating noise levels gets reduced to evaluating the value of P from lightning discharge data on the basis of the discussions in this paper, praticularly the one in Section VIII. The value of P is deduced in the next section.

XI. Power Radiated by a Flash

A flash radiates four trains of impulses as shown in Fig. 1. As explained in Sections VI-A and VIII, we are interested in the average amplitude arising from such a train of impulses. Such impulses occur at random as the radiation is a statistical phenomenon. The recurrence frequency of the impulses is ν (see Table I). The bandwidth of the receiver, B, is 6000 cps. For such a case, the root mean square amplitude has been calculated[2] and is found to be given by

$S(\omega_0)$ = root mean square amplitude due to stroke in a flash,

$$= \sqrt{2\cdot\nu\cdot B}\cdot\frac{X\cdot l}{\omega_0} \qquad (5)$$

where

$$\omega_0 = 2\cdot\pi\cdot f. \qquad (6)$$

and "f" is the frequency to which the receiver is tuned.

[15] "Ionospheric Radio Propagation," Central Radio Propagation Lab., Nat. Bur. Stds., Superintendent of Documents, Washington, D. C. NBS Circular 462; 1948.

A step in a stepped leader, it has been explained, is the radiator of the impulse and it is equivalent to a short vertical dipole. Taking account of this fact and correcting for the gain factor of the antenna, it has been shown[2] that the peak electric field intensity due to a stroke in a flash, *i.e.*, due to one train of impulses, is given by

$$E_1 = \frac{30}{c\cdot r}\cdot S(\omega_0)\cdot\sqrt{(1.5)\cdot\sin^2\theta} \quad \mu V/m \qquad (7)$$

where

c = velocity of light in meters per second,
r = distance of the place from the source in millions of meters.

We must now examine the physical significance of (7). If median values are used for all the quantities in $S(\omega_0)$, the equation gives the median value of the peak field intensity. The way in which we have deduced $S(\omega_0)$ and the concept of peak field intensity imply that some form of distribution of the values of the field intensity exists. This is but natural in any statistical phenomenon. But, we have gone a stage further. We have implicitly assumed that this distribution is sinusoidal. It has to be pointed out that the statements just represent the physical implications of the actual calculations and no more.

Now, a stroke consists of a number of steps and each step radiates an impulse. The steps follow each other in time sequence. The time duration of a complete stroke is 1 millisecond. The question that arises is whether we can provide a possible physical background to idealize the sinusoidal distribution of field intensities to an actual sinusoidal time distribution of field intensities during the period of one millisecond. A stepped leader has a commencement and an end and during the entire interval between the commencement and end, there are steps. It is reasonable to assume that a step at commencement of the leader stroke radiates a small power. Then, increasing power is radiated by successive steps till this power reaches a maximum, after which the power radiated by successive steps decreases. In other words, the time distribution of the magnitude of the electric fields arising from the impulses in the stroke is such that the field intensity increases to a maximum and then decreases. Now, applying the implied sinusoidal concept mentioned earlier, we may say that the resultant pulse arising from a stroke in a flash, *i.e.*, due to a train of impulses is a sinusoidal one. Hence, our idealization leads to the result that a stroke in a flash radiates a sinusoidal pulse of 1 millisecond duration.

Thus, the envelope of the noise field during 1 millisecond is a sinusoidal one. But, what we consider to be the significant measure of the envelope is the average value of the amplitude of the corresponding rectangular pulse. This follows from an examination of Fig. 2. There are several methods for deducing the amplitude of the equivalent rectangular pulse that corresponds to a sinu-

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soidal one. All such methods are approximate to varying degrees of accuracy. Such methods have been described and discussed.[16] Two such methods are really important. The first method is based on the principle of conservation of energy and charge and the second on the minimum departure of areas. The value of K, the ratio of the amplitude of the equivalent rectangular pulse to the peak amplitude of the sinusoidal pulse is found to be 0.837 and 0.860 respectively by the two methods. The first method is unassailable from the theoretical standpoint. The second method gives an rf spectrum that is a better approximation to the actual pulse. It is difficult to choose between the methods scientifically for the case that is being discussed in this paper. Further, the actual difference between the values of K is not large. It is, therefore, best to choose the mean value of 0.85 for K and say that the result is accurate to within ± 2 per cent. In fact, the final result in this section corresponds to that of the statistically idealized case and a better order of accuracy can never be expected.

Taking K as 0.85, the average value of the electric field due to one train of impulses, E_2, can be deduced from (7) as

$$E_2 = (0.85)E_1. \tag{8}$$

Eq. (8)[17] gives the amplitude corresponding to a rectangular pulse similar to any one of those shown in Fig. 2. Or, to put briefly, it gives the value that corresponds to A in (3). But, it is B that really corresponds to the amplitude of the equivalent carrier as judged by the ear. Therefore, if E is the field intensity corresponding to B, then E is given by

$$E = (0.32)\cdot(0.85)\cdot E_1. \tag{9}$$

An expression for P follows from (4) and (9) and inserting the numerical values as given in Table I and elsewhere, we find that

$$P = \frac{45.39}{f^2}\cdot \tag{10}$$

Hence, P may be taken as given by

$$P = 45/f^2 \text{ watts} \tag{11}$$

where "f" is the frequency in mc.

[16] W. H. Bostick and J. V. Lebacqz, "Pulse duration and amplitude," "Pulse Generators," ed. by G. N. Glasoe and J. V. Lebacqz, Mass. Inst. Tech. Rad. Lab. Ser., No. 5, McGraw-Hill Book Co., Inc., New York, N. Y., pp. 710–722; 1948.

[17] There is another way of deducing the result of (8). Due to a stroke in a flash, a large number of impulses are radiated at random. Hence, the resultant noise can be considered as equivalent to fluctuation noise. The rms field intensity E', due to a stroke is deduced from (7) as $E_1/\sqrt{2}$. Then, the amplitude of the envelope of fluctuation noise which is what we require and which should correspond to E_2 is given by $(1.25)E'$ and E_2 becomes $(0.88)E_1$. This method is not strictly correct as it fails to incorporate the concept of a sine wave pulse that the derivation of (7) implies and secondly, the steps in a leader are only about 10 to 20 and this cannot justify our regarding the resultant effect as equivalent to a fluctuation. Anyway, even this approach will not significantly alter the final result.

It will be seen that the result of (11) is the same as deduced and used earlier.[2]

Systematic measurements of noise by the method described in this paper have been carried out for several years at Poona, India (18.31 N, 73.55 E) for the period of day, 18–24 hours IST and noise levels have been deduced from such measurements for the different seasons of the year in the 3-, 5-, 9-, and 13-mc bands. Noise levels have also been estimated by the method described in Section X. The estimates agree extremely well with the noise levels deduced from actual measurements.[11,18,19]

XII. Conclusion

Atmospheric radio noise arises from electrical discharges associated with thunderstorms. Such discharges are complicated statistical phenomena and different types have different characteristics. It is first necessary to ascertain which specific type is mainly responsible for noise in a certain frequency band. Then, the numerical data corresponding to an idealized statistically valid representation of this particular type have to be deduced from the available experimental results on lightning discharges.

Atmospheric noise is a source of interference and, as such, the problem has to be examined from the standpoint of its interfering effect. This is only possible with reference to a specific service. The specific service selected for this paper is short wave broadcasting. For broadcasting, we employ continuous waves and they carry varying levels of modulation. The noise wave form gets superposed on the sound wave form and causes annoyance to the listener of broadcast programs.

The ear is the ultimate judge in listening and it hears what comes out of the loudspeaker. It judges the full effect of sound in 0.2 second. Hence, any evaluation over a period of 0.2 second is, for practical purposes, equivalent to the effect of continuous waves. The apparent loudness of sound as judged by the ear arises from the integrated effect over a period of 0.001 second. Therefore, it is the average amplitude over a period of 0.001 second that is of significance in dealing with impulses. This has the effect of converting all impulses of 0.001 second duration, whatever their form, into corresponding rectangular pulses. In its response to impulsive noise, the ear behaves like a circuit having a charging and a discharging time constant.

Taking into account the considerations as outlined, the radiations from a typical lightning flash can be analyzed and this shows that a flash gives rise to one acoustic impulse. It is this acoustic impulse that has to be measured as the source of annoyance to the listener

[18] K. R. Phadke, "Atmospheric noise interference to broadcasting in the 5 mc band at Poona," *J. Inst. Telecomm. Eng. (India)*, vol. 1, pp. 136–146; September, 1955.

[19] C. K. Sane, "Atmospheric Noise Interference to Broadcasting in the 9 and 13 Mc Bands at Poona," Ph.D. dissertation, University of Poona, Poona, India; 1957.

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and this measurement has to be carried out in terms of continuous signals carrying a certain level of modulation. The appropriate level of modulation and the modulating frequency follow automatically from the analysis. These considerations define the requirements of an objective noise meter.

The noise field strength as deduced from such measurements has to be estimated. This requires the corresponding idealized radiator and the power radiated from such an idealized radiator has to be deduced from the lightning discharge data.

Although the paper is thus restricted to examining the problem of atmospheric noise interference to short wave broadcasting, the general principles emerging from the discussion may be of wider application. It is clear that a generalized mathematical treatment of the problem based on Fourier analysis and statistics is, by itself, of perhaps not a great value. Similarly, a pure physical analysis leading to an evaluation of a general expression for noise power at the detector may serve no useful purpose. Engineering evaluations of noise have to take account of the equipment which picks up the noise and which is affected by it and the actual experimental conditions under which the effect is produced. In the particular case examined in this paper, the equipment consists of the broadcast receiver and the human ear and the experimental condition is that noise comes along with a program.

Theory of Junction Diode and Junction Transistor Noise*

A. VAN DER ZIEL†, FELLOW, IRE AND A. G. T. BECKING‡

Summary—**A. van der Ziel has given formulas for shot noise in junction diodes and junction transistors for transistors in which:**

1) All current is carried by one type of carrier.
2) The carrier flow is one-dimensional.
3) The recombination is by volume recombination.

These equations are here proved with the help of a corpuscular approach without any significant restrictions except that the individual holes can be treated as independent.

The emitter and collector currents are then split into various parts for which the noise spectrum can be obtained by relatively simple reasoning.

Introduction

As SHOWN by van der Ziel and others,[1–3] the noise behavior of a junction diode can be expressed by a current generator i in parallel to the junction, such that

$$\overline{i^2} = 4kTGdf - 2eIdf. \tag{1}$$

Here G is the junction conductance and I the junction current; the latter is taken positive for forward bias and negative for back bias. If the junction has an appreciable series resistance, one would expect full thermal noise of this resistance. These predictions have been verified by various authors; probably the most accurate verification was carried out by Champlin.[4]

The noise behavior of a junction transistor can be expressed by two noise current generators,[1,3] a noise generator i_e across the emitter junction and a noise current generator i_c across the collector junction, such that

$$\overline{i_e^2} = 4kTG_edf - 2eI_edf \tag{2}$$

$$\overline{i_c^2} = 2eI_cdf \tag{3}$$

$$\overline{i_e^*i_c} = 2kTY_{ce}df \tag{4}$$

where the asterisk denotes a conjugate complex quantity. Here I_e is the emitter current, I_c the collector current, G_e the emitter conductance and Y_{ce} the emitter-collector transfer admittance. If any of the contacts has an appreciable series resistance, one would expect full thermal noise of such a resistance. This is especially important for the base resistance $R_{b'b}$ but in some cases it may be necessary to take into account the thermal noise of the series resistance of the emitter contact. The equivalent noise circuit is shown in Fig. 1.

Eqs. (1)–(4) (or their equivalents) were proven rigor-

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‡ Research physicist at the Philips Res. Labs., Eindhoven, The Netherlands, died June 26, 1957 after a long illness. This paper is the final result of a discussion on noise in semiconductor devices during Dr. Becking's visit to the Univ. of Minnesota in June, 1956.

[1] A. van der Ziel, "Theory of shot noise in junction diodes and junction transistors," Proc. IRE, vol. 43, pp. 1639–1646; November, 1955; "Theory of shot noise in junction diodes and junction transistors," vol. 45, p. 1011; July, 1957.

[2] A. Uhlir, "High-frequency shot noise in p-n junctions," Proc. IRE, vol. 44, pp. 557–558; April, 1956. Correction, p. 1541; November, 1956.

[3] W. Guggenbuehl and M. J. O. Strutt, "Theory and experiments on shot noise in semiconductor junction diodes and transistors," Proc. IRE, vol. 45, pp. 839–854; June, 1957.

[4] K. S. Champlin, "A Study of Shot and Thermal Noise in Silicon P-N Junction Diodes," Univ. of Minnesota, M. Sc. thesis. 1955.

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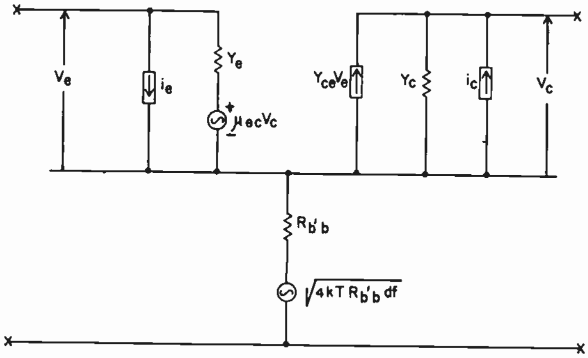

Fig. 1—Equivalent circuit of a transistor.

ously by van der Ziel[1] for a one-dimensional model in which all current is carried by holes. Since the equations do not give any reference to the particular model used, it might be hoped that they are of general validity. A simpler but much less rigorous proof was given by Guggenbuehl and Strutt, who were also the first to put (4) in the form presented here. Eqs. (1)–(4) agree reasonably well with experiment;[5,6] the experimental data were not as accurate as in the diode case, however. A more accurate test is now under way at the University of Minnesota.

Despite the fact that the noise behavior of junction diodes and junction transistors is now reasonably well understood, there is still a need for a proof of (1)–(4), more rigorous than Guggenbuehl's and Strutt's treatment and more general than van der Ziel's proof. The proof to be given here is an extension of earlier work by Uhlir.[2]

The basis of this proof is the fact that a carrier crossing a p-n junction transition region gives a very sharp current pulse when it *becomes* a minority carrier or when it *ceases to be* a minority carrier. The charge transferred in this pulse is $\pm e$, the sign depends on the sign of the carrier charge and on the direction of current flow. The current pulses of the individual carriers can be considered independent.

To clarify this, consider a p-n junction with two ohmic contacts in which all current is carried by holes. Let the junction be biased so that holes are injected into the n region. If a hole enters the n region across the barrier, its space charge will be neutralized in a very short time ($\simeq 10^{-12}$ seconds) by a slight rearrangement of the electron distribution in the n region; ultimately, an electron will enter through the ohmic contact to make the n region outwardly neutral, but that does not give a noticeable amount of noise. The passage of a hole into the n region thus gives a current pulse of very short duration; the total displaced charge per pulse is $+e$. In the same way the departure of a hole out of the n region back into the p region gives a sharp current pulse of opposite polarity. No effect is noticed in the external circuit when a hole and an electron recombine in the n region.

Application to Junction Diodes

First we assume that all current is carried by holes and we divide the minority carriers entering or leaving the n region into three groups (Fig. 2) that are considered independent.

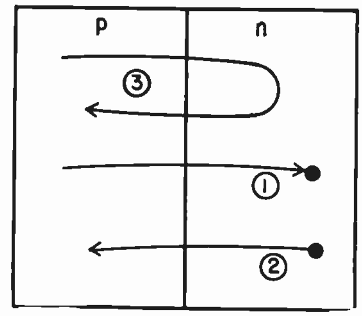

Fig. 2—Hole flow in a junction diode.

1) Holes entering the n region and dying there.
2) Holes generated into the n region and passing into the p region.
3) Holes passing into the n region and flowing back again.

The holes of groups 1) and 2) give rise to single, random, sharp current pulses and the holes of group 3) give rise to two sharp current pulses of opposite polarity and random time delay.

Let I be the dc current and I_o the saturated back current. The holes in group 2 give a contribution $-I_o$ to the dc current; since group 3) gives no contribution, group 1) must give a contribution $(I+I_o)$ to the dc current.

Before turning to the noise, let us consider the junction admittance $Y=(G+jB)$ in more detail. The applied ac voltage modulates the number of holes of group 1) and group 3); however, at low frequencies group 3) gives a negligible contribution to Y. Since the hole density just inside the n region follows the applied voltage practically instantaneously, the holes of group 1) give a contribution G_o to Y at all frequencies, where G_o is the dc junction conductance. Since $(I+I_o)=I_o \exp(eV/kT)$, where V is the dc voltage applied across the junction, the dc diode conductance is:

$$G_o = e\frac{(I+I_o)}{kT}. \tag{5}$$

Another contribution to the admittance Y is due to the capacitance C_T of the junction transition region. Thus the holes of group 3) must give a contribution $(Y-G_o-j\omega C_T)$ to Y. It is well known that the diode conductance G increases with increasing frequency; our discussion shows that the part $(G-G_o)$ must be attributed to the holes of group 3).

[5] G. H. Hanson and A. van der Ziel, "Shot noise in transistors," Proc. IRE, vol. 45, pp. 1538–1542; November, 1957.
[6] E. G. Nielson, "Behavior of noise figures in transistors," Proc. IRE, vol. 45, pp. 957–962; July, 1957.

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We turn now to the noise. Since groups 1) and 2) give rise to sharp, independent, random current pulses of charge e, the currents $(I+I_o)$ and $(-I_o)$ give full shot noise at all frequencies. Hence, the contributions of groups 1) and 2) to $\overline{i^2}$ are:

$$\overline{i_1^2} = 2e(I + I_o)df \tag{6}$$

$$\overline{i_2^2} = 2eI_o df. \tag{7}$$

Group 3) consists of holes that return to the p region. This return is due to diffusion, that is, due to interaction with the lattice vibrations; since this interaction is of thermal origin, one would expect group 3) to give full thermal noise of the incremental conductance $(G-G_o)$ at all frequencies. Hence, their contribution to $\overline{i^2}$ is:

$$\overline{i_3^2} = 4kT(G - G_o)df. \tag{8}$$

A more rigorous proof of this equation is given in the Appendix.

Adding the contributions of the three groups yields:

$$\overline{i^2} = \overline{i_1^2} + \overline{i_2^2} + \overline{i_3^2}$$
$$= 2e(I + I_o)df + 2eI_o df + 4kT(G - G_o)df. \tag{9}$$

Substituting (5) into (9) yields (1). Note that nothing has been said about the geometry of the junction; thus (9) and hence (1) hold for all geometries and not only for van der Ziel's one-dimensional model.

One might object to treating the holes of group 1) and of group 3) as independent, since it is not known in advance whether a hole passing from the p region into the n region will belong to group 1) or to group 3). We observe, however, that each hole will either belong to group 1) or to group 3) and the behavior of one hole is not influenced by another hole. This is sufficient for the validity of our derivation.

It is essential for the derivation of (9) that the field strength in the n region is zero. If an appreciable field strength occurs, the current at the beginning of the n region is no longer given by diffusion alone, but there is also a field term caused by the hole injection. The occurrence of this field means that the passage of individual holes across the transition region perhaps may be considered no longer as a series of independent, random events. An appreciable field strength in the n region can be expected if the injected hole density is comparable with or larger than the equilibrium electron density in the n region (large injection levels). Therefore it is not certain that (9) and hence (1) will be valid in that case. Experimental work on this problem might clarify the situation and might indicate how large the deviation from (1) or (9) actually is. Probably the effect is very small.

Finally we drop the assumption that all current is carried by holes. It is obvious that (1) is also true if all current is carried by electrons. The case that the current is carried partly by electrons and partly by holes requires further consideration.

Let

$$I = I_n + I_p; \qquad I_o = I_{no} + I_{po}$$
$$G = G_n + G_p; \qquad G_o = G_{no} + G_{po} \tag{10}$$

where I_n, I_{no}, G_n, and G_{no} are the contributions of the electrons to I, I_o, G, and G_o, respectively, and I_p, I_{po}, G_p, and G_{po} are the corresponding contributions of the holes. Treating individual current pulses as independent, one would expect

$$\overline{i^2} = \overline{i_n^2} + \overline{i_p^2} \tag{11}$$

where $\overline{i_n^2}$ and $\overline{i_p^2}$ are the contributions of the electrons and of the holes to $\overline{i^2}$, respectively. In analogy with (9) we have:

$$\overline{i_n^2} = 2e(I_n + I_{no})df + 2eI_{no}df + 4kT(G_n - G_{no})df \tag{11a}$$

$$\overline{i_p^2} = 2e(I_p + I_{po})df + 2eI_{po}df + 4kT(G_p - G_{po})df. \tag{11b}$$

Substituting into (11) yields (9) and hence (1), so that this equation is now also proven for the case in which part of the current is carried by electrons. Except for the case mentioned above, (1) should thus be generally valid.

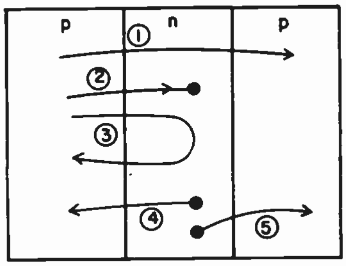

Fig. 3—Hole flow in a p-n-p junction transistor.

Application to Junction Transistors

First we consider a p-n-p transistor in which all current is carried by holes. The holes entering or leaving the base region now may be split into five groups (Fig. 3) that are considered independent.

1) Holes injected by the emitter and collected by the collector.
2) Holes injected by the emitter and dying in the base region.
3) Holes injected by the emitter and returning to the emitter.
4) Holes generated in the base region and collected by the emitter.
5) Holes generated in the base region and collected by the collector.

It is hereby tacitly assumed that the collector is biased so that it does not inject any holes into the base region.

Let group 4) give a contribution $(-I_{ee})$ to the emitter current I_e and group 5) a contribution I_{cc} to the col-

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lector current I_c. If β_o is the dc collector efficiency, group 1) gives a contribution $\beta_o(I_e+I_{ee})$ to the dc emitter current, group 2) gives a contribution $(1-\beta_o)(I_e+I_{ee})$, and group 3) gives no contribution. Since the holes of groups 1) and 5) are collected by the collector, the collector current I_c may be written

$$I_c = \beta_o(I_e + I_{ee}) + I_{cc} = \beta_o I_e + I_{co} \tag{12}$$

where

$$I_{co} = \beta_o I_{ee} + I_{cc} \tag{12a}$$

is the collector saturated current.

Before turning to the noise, we consider first the emitter admittance $Y_e = G_e + jB_e$ and the transfer admittance Y_{ce} in more detail.

An applied ac voltage v_e modulates the number of holes of groups 1) to 3); however, at low frequencies the holes of group 3) give a negligible contribution to Y_e. Since the hole density just inside the n region follows the applied voltage practically instantaneously, the holes of groups 1) and 2) give a contribution G_{eo} to Y_e at all frequencies, where G_{eo} is the dc emitter conductance. Since $(I_e+I_{ee}) = I_{ee} \exp\,(eV_e/kT)$, where V_e is the dc voltage applied across the emitter junction, the dc emitter conductance is:

$$G_{eo} = e(I_e + I_{ee})/kT \tag{13}$$

in analogy with (5); the part β_o of this conductance comes from the holes of group 1) and the part $(1-\beta_o)$ comes from the holes of group 2). Another contribution to the admittance Y_e is due to the capacitance C_{Te} of the emitter junction transition region. Thus the holes of group 3) must give a contribution $(Y_e - G_{eo} - j\omega C_{Te})$ to Y_e; the part $(G_e - G_{eo})$ of the emitter conductance G_e then is due to holes of group 3).

The signal transfer admittance Y_{ce} is due solely to the holes of group 1). Since these holes give a contribution $\beta_o G_{eo}$ to Y_e at all frequencies, the low-frequency signal transfer admittance Y_{ceo} is:

$$Y_{ceo} = \beta_o G_{eo} = \beta_o e(I_e + I_{ee})/kT. \tag{14}$$

If an ac signal v_e is applied to the emitter junction, the holes of group 1) give a contribution i_{e1} to the ac emitter current and a contribution i_{c1} to the ac collector current so that

$$i_{e1} = Y_{ceo} v_e; \qquad i_{c1} = Y_{ce} v_e = \frac{Y_{ce}}{Y_{ceo}}\, i_{e1}. \tag{15}$$

Here i_{e1} is frequency independent, since the hole concentration at the emitter side of the base region follows the applied emitter voltage practically instantaneously. At high frequencies $|Y_{ce}| < Y_{ceo}$ and Y_{ce} become complex. The reason is that diffusion is a random process; thus there is a random time delay between the passage of an individual hole of group 1) across the emitter junction and the passage of the same hole across the collector junction. We need this for our discussion of the noise properties of the transistor.

We now turn to the noise. Since the holes of groups 1) to 4) contribute to $\overline{i_e^2}$, we have in analogy with (9)

$$\begin{aligned}\overline{i_e^2} &= \overline{i_{e1}^2} + \overline{i_{e2}^2} + \overline{i_{e3}^2} + \overline{i_{e4}^2} \\ &= 2e(I_e + I_{ee})df + 4kT(G_e - G_{eo})df + 2eI_{ee}df \end{aligned} \tag{16}$$

where the first term is due to the combined effect of groups 1) and 2). Since the holes of groups 1) and 5) contribute to $\overline{i_c^2}$, we have in analogy with (9)

$$\overline{i_{c2}} = \overline{i_{c1}^2} + \overline{i_{c5}^2} = 2e\beta_o(I_e + I_{ee})df + 2eI_{cc}df = 2eI_c df. \tag{17}$$

Thus (3) has been proved. Substituting (13) into (16) yields (2).

The cross correlation $\overline{i_e^* i_c}$ can be due only to holes of group 1), since this represents the only group that contributes to both I_e and I_c. That is:

$$\overline{i_e^* i_c} = \overline{i_{e1}^* i_{c1}} \tag{18}$$

Since the holes of group 1) produce independent, random pulses in the emitter and in the collector circuits, we have:

$$\overline{i_{e1}^2} = \overline{i_{c1}^2} = 2e\beta_o(I_e + I_{ee})df = 2kTY_{ceo}df \tag{19}$$

which is found by substituting (14). At low frequencies i_e and i_c are fully correlated, since the holes of group 1) pass both junctions. This means that,

$$\overline{i_{e1}^* i_{c1}} = \overline{i_{c1}^2} \tag{18a}$$

in that case. At higher frequencies $\overline{i_{e1}^* i_{c1}}$ becomes complex and $\overline{i_{e1}^* i_{c1}}$ decreases with increasing frequency because of the random time delay between the passage of an individual hole across the emitter and across the collector junctions. Since there is no difference between the flow of ac current and between the flow of noise currents across the junctions, (15) should hold for the noise due to the holes of group 1). That is, if i'_{c1} is that part of i_{c1} that is correlated with i_{e1},

$$i'_{c1} = \frac{Y_{ce}}{Y_{ceo}}\, i_{e1}; \tag{20}$$

$$\overline{i_e^* i_{c1}} = \overline{i_{e1}^* i'_{c1}} = \frac{Y_{ec}}{Y_{ceo}}\, \overline{i_{e1}^2} = 2kTY_{ce}df \tag{20a}$$

which is found by substituting (19); this proves (4). A more rigorous proof is given in the Appendix.

Eqs. (2)–(4) now have been proved for an arbitrary p-n-p junction in which all current is carried by holes. The restriction for the validity of (2)–(4) is the same as for (1), *viz.*, that the field strength in the base region due to injected holes must be zero. This means that the equations may not be valid for high injection levels. More experimental work is needed to decide how important the deviations from (2)–(4) are in these cases. Probably the deviations are very small.

It is obvious that (2)–(4) are also true for n-p-n

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transistors in which all current is carried by electrons. The case of a p-n-p transistor in which part of the current is carried by electrons (or the case of n-p-n transistors in which part of the current is carried by holes), requires further discussion.

Consider a p-n-p transistor in which part of the current is carried by electrons. These electrons consist of four groups (Fig. 4).

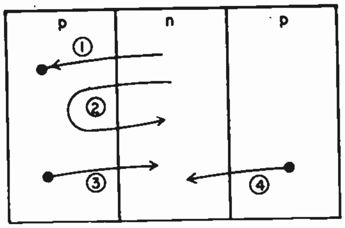

Fig. 4—Electron flow in a p-n-p junction transistor.

1) Electrons injected from the base region into the emitter region and dying there.
2) Electrons injected into the emitter region and returning to the base region.
3) Electrons generated in the emitter region and collected by the base region.
4) Electrons generated in the collector region and collected by the base region.

It is hereby assumed that the collector is biased such that no electrons are injected from the base region into the collector region.

Let the contribution of the electrons and of the holes be distinguished by the subscripts n and p, respectively, then:

$$I_e = I_{ne} + I_{pe}; \qquad I_c = I_{nc} + I_{pc} \tag{21}$$

$$I_{ee} = I_{nee} + I_{pee}; \qquad I_{cc} = I_{ncc} + I_{pcc} \tag{21a}$$

$$G_e = G_{ne} + G_{pe}; \qquad Y_{ce} = Y_{nce} + Y_{pce} \tag{21b}$$

$$i_e = i_{ne} + i_{pe}; \qquad i_c = i_{nc} + i_{pc}. \tag{21c}$$

Since all current pulses are independent and occur at random, i_{ne} and i_{pe} will be uncorrelated and so will i_{nc} and i_{pc}. Since no electron group contributes to both the emitter and the collector current, we have

$$Y_{nce} = 0 \quad \text{and} \quad \overline{i_{ne}{}^* i_{nc}} = 0. \tag{22}$$

Hence,

$$\overline{i_e{}^2} = \overline{i_{ne}{}^2} + \overline{i_{pe}{}^2}; \quad \overline{i_c{}^2} + \overline{i_{pc}{}^2}; \quad \overline{i_e{}^* i_c} = \overline{i_{pe}{}^* i_{pc}}. \tag{23}$$

In analogy with (2)–(4) we then have

$$\overline{i_{ne}{}^2} = 4kTG_{ne}df - 2eI_{ne}df$$

$$\overline{i_{nc}{}^2} = 2eI_{nc}df; \quad \overline{i_{ne}{}^* i_{nc}} = 0 \tag{23a}$$

$$\overline{i_{pe}{}^2} = 4kTG_{pe}df - 2eI_{pe}df$$

$$\overline{i_{pc}{}^2} = 2eI_{pc}df; \quad \overline{i_{pe}{}^* i_{pc}} = 2kTY_{pce}df. \tag{23b}$$

Substituting (23a) and (23b) into (23) and making use of (21)–(21b) yields (2)–(4) again. Except for the case mentioned above these equations are thus of general validity.

Appendix

A More Rigorous Proof of (8) with the Help of Carson's Theorem

Let identical events $y(t)$ occur at random at the average rate $\overline{N}$, let each event have a time constant τ and let

$$F(\omega, \tau) = \int_0^\infty y(t) \exp(j\omega t) dt \tag{24}$$

exist. Then the spectral density of the noise is:

$$S(f) = 2\overline{N} \mid F(\omega, \tau) \mid^2; \tag{25}$$

this is known as *Carson's theorem*. If an applied voltage modulates the rate of occurrence of these events instantaneously, then the system has an admittance,

$$Y = G + jB = \frac{\partial \overline{N}}{\partial V} F(\omega, \tau). \tag{26}$$

If there is a distribution in time constants τ

$$dN = \overline{N} g(\tau) d\tau \quad \text{with} \quad \int_0^\infty g(\tau) d\tau = 1 \tag{27}$$

then (25) and (26) must be replaced by

$$S(f) = 2\overline{N} \int_0^\infty \mid F(\omega, \tau) \mid^2 g(\tau) d\tau \tag{28}$$

$$Y = G + jB = \frac{\partial \overline{N}}{\partial V} \int_0^\infty F(\omega, \tau) g(\tau) d\tau. \tag{29}$$

In our particular case $y(t)$ is the current pulse due to a hole of group 3). Let $g(\tau)d\tau$ be the probability that a hole of group 3) stays in the n region for a time interval between τ and $(\tau + d\tau)$, and let $\overline{N}_3$ be the average rate of occurrence of pulses due to holes of group 3). Then,

$$F(\omega, \tau) = e[1 - \exp(-j\omega\tau)];$$

$$\mid F(\omega, \tau) \mid^2 = 2e^2(1 - \cos \omega\tau). \tag{30}$$

Substituting into (28) we obtain:

$$S(f) = 4e^2 \overline{N}_3 \int_0^\infty (1 - \cos \omega\tau) g(\tau) d\tau \tag{28a}$$

$$Y_3 = G_3 + jB_3 = e \frac{\partial \overline{N}_3}{\partial V} \int_0^\infty [1 - \exp(-j\omega\tau)] g(\tau) d\tau. \tag{29a}$$

Now we observe that the real part G_3 of Y_3 is equal to $(G - G_0)$ and that

$$\frac{\partial \overline{N}_3}{\partial V} = \frac{e}{kT} \overline{N}_3. \tag{31}$$

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Consequently,

$$(G - G_o) = \frac{e^2}{kT} \overline{N}_3 \int_0^\infty (1 - \cos \omega\tau) g(\tau) d\tau. \quad (32)$$

Since $S(f)$ and $(G-G_o)$ contain the same unknown expression, that expression may be eliminated and $S(f)$ may be expressed in terms of $(G-G_o)$. This yields

$$S(f) = 4kT(G - G_o) \quad (33)$$

which corresponds to (8).

A More Rigorous Proof of (20a) with the Help of a Modification of Carson's Theorem

Let $\overline{N}_1$ be the average rate of occurrence of holes of group 1) and let $h(\tau)d\tau$ be the probability that a hole of group 1) diffuses across the barrier in a time interval between τ and $(\tau+d\tau)$ such that $\int_0^\infty h(\tau)d\tau = 1$. The holes of group 1) give rise to sharp current pulses of charge e across the emitter and across the collector junction with a time interval τ between the two pulses. If all holes had the same diffusion time τ, we would have from Carson's theorem, as applied to cross-correlation terms

$$\overline{i_{e1}^* i_{c1}} = 2df\overline{N}_1 e^2 \exp(-j\omega\tau). \quad (34)$$

Taking into account the distribution in time constants, this becomes:

$$\overline{i_{e1}^* i_{c1}} = 2df\overline{N}_1 e^2 \int_0^\infty \exp(-j\omega\tau) h(\tau) d\tau. \quad (34a)$$

The rate $\overline{N}_1$ at which holes of group 1) are injected across the emitter is modulated instantaneously by the voltage V_e applied across the emitter junction. If all holes of group 1) had the same diffusion time τ across the base region, the transfer admittance Y_{ce} would be:

$$Y_{ce} = \frac{\partial \overline{N}_1}{\partial V_e} e \exp(-j\omega\tau) = \frac{e^2}{kT} \overline{N}_1 \exp(-j\omega\tau) \quad (35)$$

since $\partial\overline{N}_1/\partial V_e = e\overline{N}_1/kT$. Taking into account the distribution in time constants τ this becomes:

$$Y_{ce} = \frac{e^2}{kT} \overline{N}_1 \int_0^\infty \exp(-j\omega\tau) h(\tau) d\tau. \quad (35a)$$

We observe that (34a) and (35a) contain the same unknown expression; that expression may be eliminated and $i_{e1}^* i_{c1}$ may be expressed in terms of Y_{ce}. This yields

$$\overline{i_{e1}^* i_{c1}} = 2kT Y_{ce} df \quad (36)$$

which proves (20a) and hence (4).

Acknowledgment

The authors are indebted to Dr. K. M. van Vliet, University of Minnesota, for stimulating discussions that greatly contributed to this presentation. They acknowledge the encouragement and criticism of Dr. D. O. North, RCA Laboratories. Dr. G. Eichholz, Bureau of Mines, Ottawa, Canada, suggested the step leading to (15) and (20).

Ferrite Microwave Detector*

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Summary—**In treating the behavior of the magnetic moments of unbalanced electron spins in ferromagnetic materials under the action of an rf field, second-order terms in the alternating components are usually neglected. It is shown here that retention of certain second-order terms for one component of the magnetization predicts the possibility of using ferrites to detect an amplitude-modulated microwave signal. The demodulated envelope can be used to magnetostrict a long thin ferrite rod, the vibration of which can be observed by means of a polarized $BaTiO_3$ ceramic rod.**

The results obtained with an experimental model of this device indicate verification of the theory. Factors which are expected to improve performance are discussed.

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Introduction

THE phenomenon of the precession of the magnetic moments of unbalanced electron spins in ferromagnetic materials under the action of an rf field has been the subject of much investigation. The resonance associated with this precession has been thoroughly investigated and, based on the knowledge gained, many ferrite devices have been designed. Only a few of these, however, make use of the second-order effects inherent in the theory. Second-order effects have been considered by Ayres, Vartanian, and Melchor,[1] who investigated

[1] W. P. Ayres, P. H. Vartanian, and J. L. Melchor, "Frequency doubling in ferrites," *J. Appl. Phys.* vol. 27, pp. 188–189; February, 1956.

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frequency doubling, and by Pippin,[2] who derived expressions both for frequency doubling and mixing.

It is here shown that if second-order terms in one component of the magnetization are retained, the theory predicts the possibility of using ferrites to detect an amplitude-modulated microwave signal. The magnetization component alternating at the modulation frequency can be used to magnetostrict the ferrite. If the modulation frequency is adjusted to equal the mechanical resonant frequency of a ferrite rod, large vibrations will result. These vibrations are observed easily by means of a ferroelectric transducer bonded to the end of the ferrite rod or by a coil wound around the rod at a nodal point. The latter technique was used by Simon and Broussaud[3] when they first observed detection by ferrites at microwave frequencies.

The purpose of this paper is to present the theory of the detection phenomena and to show the data taken in its support. Also, the technique of observing the magnetostriction by a ferroelectric transducer as used in these experiments is discussed. Since this is an initial investigation of a physical phenomenon, detailed information pertaining to applications, such as the bandwidth and dynamic range of the detector, is not available as yet. The detector model described is not of optimum design, and factors which will improve the performance are discussed.

Theory

The model which is considered in deriving the second-order effects leading to detection is the conventional one used in ferromagnetic resonance theory; *i.e.*, a constant magnitude magnetization vector which is under the torque influence of a magnetic field is postulated. The magnetization vector is antiparallel and proportional to an angular momentum vector associated with the spin of the unpaired electron. Thus, the vectorial time rate of change in the angular momentum can be related simply to a change in the magnetization vector and the following relationship for the lossless infinite medium case results:

$$\dot{\vec{M}} = \gamma(\vec{M} \times \vec{H}) \tag{1}$$

where the "dot" notation is used for time derivative and

γ = magnetomechanical ratio of the unpaired electron; to be taken throughout as $|\gamma|$

$\vec{M}$ = magnetization

$\vec{H}$ = magnetic field.

It is useful to refer to the model in Fig. 1.

M_s is the saturation magnetization and is equal in magnitude to M. The lower case quantities (m) are the alternating field components of the magnetization.

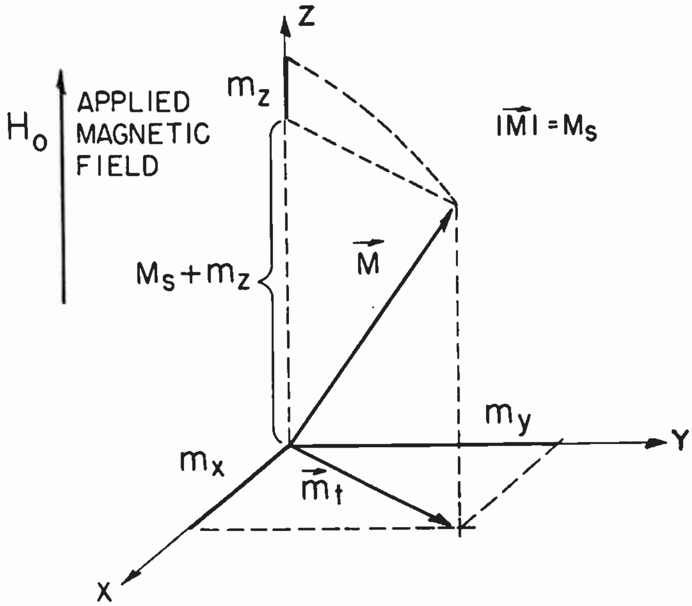

Fig. 1—Components of the magnetization.

$$\vec{M} = \vec{i}m_x + \vec{j}m_y + \vec{k}(m_z + M_s)$$

and

$$|M|^2 = m_t^2 + m_z^2 + 2m_zM_s + M_s^2 = M_s^2$$

$$\therefore m_z = -M_s\left[1 - \sqrt{1 - \frac{m_t^2}{M_s^2}}\right] \tag{2}$$

where

$$m_t^2 = m_x^2 + m_y^2.$$

The magnetic field $\vec{H}$ is composed of the static field $\vec{H_0}$ and the alternating field components $\vec{h} = \vec{i}h_x + \vec{j}h_y + \vec{k}h_z$.

First-Order Solution

The solution of (1), using real variables and neglecting cross products of the time-varying quantities, is

$$m_x = \alpha h_x - \beta \dot{h}_y \tag{3}$$

$$m_y = \alpha h_y + \beta \dot{h}_x \tag{4}$$

where the form of h_x and h_y may be

$$h_x = A_1 \sin \omega t + B_1 \cos \omega t \tag{5}$$

$$h_y = A_2 \sin \omega t + B_2 \cos \omega t. \tag{6}$$

The quantities α and β are

$$\alpha = \frac{\gamma^2 H_0 M_s}{\gamma^2 H_0^2 - \omega^2} \tag{7}$$

$$\beta = \frac{\gamma M_s}{\gamma^2 H_0^2 - \omega^2}. \tag{8}$$

Such a first-order solution yields a zero value for $\dot{m}_z$.

[2] J. P. Pippin, "Frequency doubling and mixing in ferrites," Proc. IRE, vol. 44, p. 1054; August, 1956.

[3] J. C. Simon and G. Broussaud, "Detection d'une onde hertzienne par une ferrite," *Compt. Rend.*, vol. 238, pp. 2294–2296; June 14, 1954.

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Second-Order Solution

However, the second-order approximation gives

$$\dot{m}_z = \gamma(m_x h_y - m_y h_x)$$

and using (3) and (4)

$$\dot{m}_z = -\gamma\beta(h_y\dot{h}_y + h_x\dot{h}_x). \tag{9}$$

Integrating

$$m_z = -\frac{\gamma\beta}{2}(h_y^2 + h_x^2) + C \tag{10}$$

where C is a constant.

In order to determine the variation with applied magnetic field of the detected signal for the case in which the A_n and B_n in (5) and (6) are slowly varying functions of time, these coefficients may be assumed to be constants equal to their respective peak values. The validity of such an assumption is dependent on how small is the value of their time variation compared with the time variation of the microwave signal. Since the microwave angular frequency (ω) in the experiments to be described is on the order of $2\pi \times 10^{10}$ radians/second and the modulation frequency $(\omega_A) 2\pi \times 10^4$ radians/second, the approximation is exact for all practical purposes.

Resort must be had to (2) in order to evaluate the constant C. Assuming that the time varying components of M are small compared with the static components so that $m_t^2 \ll M_s^2$, then

$$m_z \cong -\frac{m_t^2}{2M_s}. \tag{11}$$

Using (3) and (4),

$$m_z = -\frac{1}{2M_s}\left[\alpha^2(h_x^2 + h_y^2) + \beta^2(\dot{h}_y^2 + \dot{h}_x^2) + 2\alpha\beta(h_y\dot{h}_x - h_x\dot{h}_y)\right]. \tag{12}$$

By considering (5) and (6) and the definitions of α and β from (7) and (8), the above can be reduced to

$$m_z = -\frac{\gamma\beta}{2}(h_x^2 + h_y^2) - \frac{\beta^2}{2M_s}\left[\omega^2(A_1^2 + B_1^2 + A_2^2 + B_2^2) + 2\gamma H_0\omega(A_2B_1 - A_1B_2)\right]. \tag{13}$$

The constant in (10) has been evaluated and the value of m_z is completely determined to a second-order approximation.

By expanding the terms in (13) and collecting the time-varying terms the double frequency components are found to be

$$m_{z\,(\text{microwave})} = \frac{-\gamma^2 M_s}{2(\gamma^2 H_0^2 - \omega^2)} D \sin(2\omega t + \delta) \tag{14}$$

where

$$\delta = \cos^{-1}\left[\frac{A_1A_2 + B_1B_2}{D}\right]$$

and

$$D = \Big[\left[(A_1 + B_2)^2 + (A_2 - B_1)^2\right] \cdot \left[(A_1 - B_2)^2 + (A_2 + B_1)^2\right]\Big]^{1/2}.$$

The dc or audio-frequency components are

$$m_{z\,(\text{audio})} = \frac{-\gamma^2 M_s}{8}\left[\frac{(A_1 + B_2)^2 + (A_2 - B_1)^2}{(\omega + \gamma H_0)^2} + \frac{(A_1 - B_2)^2 + (A_2 + B_1)^2}{(\omega - \gamma H_0)^2}\right]. \tag{15}$$

The relationship (14) has been derived and used in slightly different form by Ayres, Vartanian, and Melchor.[1]

In the frequency doubling term it is seen that for a circularly polarized microwave ($A_1 = \pm B_2$; $A_2 = B_1 = 0$) the amplitude D becomes zero. The physical interpretation of this fact is discussed by Pippin as well as by Melchor, Ayres, and Vartanian in a later paper[4] and will be reviewed below.

The detected signal is seen to have its largest value for positive circular polarization; *i.e.*, $A_1 = -B_2 (A_2 = B_1 = 0)$. For the opposite sense, $A_1 = B_2$, the term with $(\omega + \gamma H_0)$ in the denominator is nonzero, but does not go through a resonance. It will be noted also that a change in the sign of the magnetic field H_0 is equivalent to a change in the sense of polarization, which is as it should be.

The physical picture of the phenomena outlined above is illustrated in Fig. 2. With the rf field off, $\vec{M}$ is aligned with the dc field and with the transverse rf field applied the magnetization vector fans out into precession about the dc field. Fig. 2(a) illustrates the precession of the magnetization vector for a circularly polarized driving field or for a linear driving field at resonance for the lossless case. Here there is a variation in m_z at the audio rate as the modulated microwave field alternately fans the magnetization vector out into circular precession and allows it to relax to a position parallel to the applied dc field. There is no rf variation in m_z for this case. Fig. 2(b) depicts the precession of the magnetization vector for a noncircularly polarized driving field. In this case, as $\vec{M}$ rotates through one cycle m_z varies through two maxima and two minima values. Hence the double frequency component. Again, the audio variation arises from the fanning-out and relaxation of $\vec{M}$ and is here seen to be somewhat smaller than in the previous case. As the magnetic field is in-

[4] J. L. Melchor, W. P. Ayres, and P. H. Vartanian, "Microwave Frequency Doubling from 9 to 18 KMC in Ferrites," Electronic Defense Lab., Tech. Memo. No. EDL-M79; September 20, 1956.

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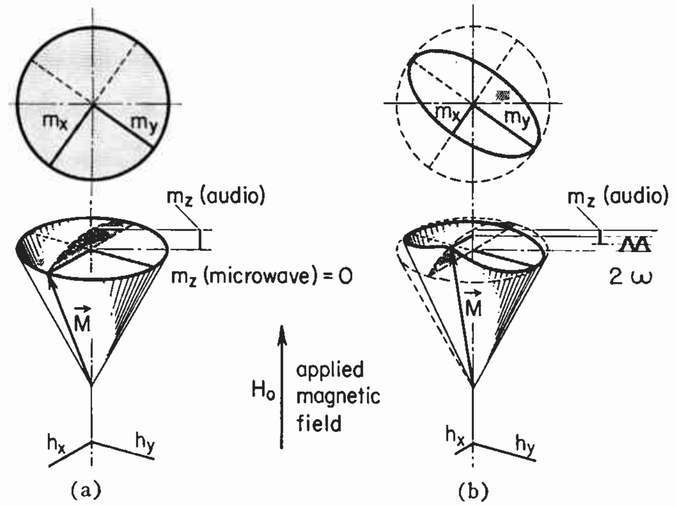

Fig. 2—(a) Precession of the magnetization vector for a circularly polarized driving field (h) or for a linear driving field at resonance (lossless case). (b) Precession of the magnetization vector for a noncircularly polarized driving field (h).

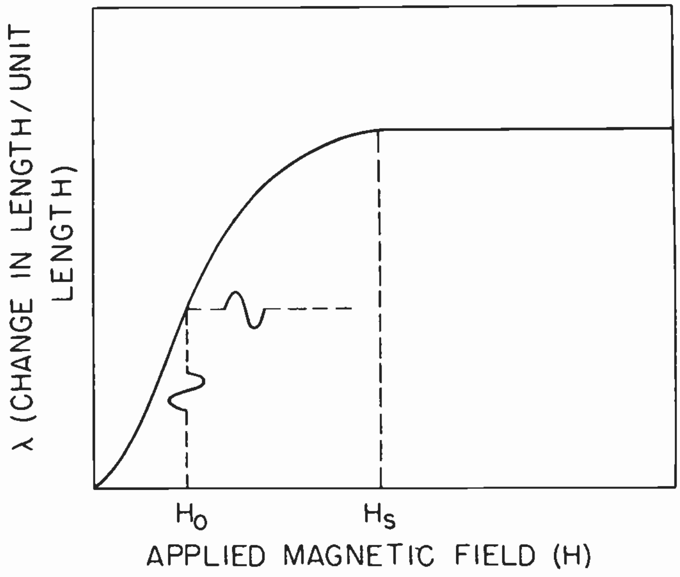

Fig. 3—Typical changes in length per unit length as a function of magnetizing field for a ferrite rod.

creased through ferromagnetic resonance, the double frequency component has a dispersion-type variation while the audio term has an absorption-type variation.

Magnetostriction

The experimental technique for observing the detected signal is based on the magnetostrictive properties of the ferrites used and will be described more fully later. The audio variation of the magnetization vector produces a strain in the ferrite crystal lattice.[5] It is important to realize that the magnitude of the strain will depend on the magnitude of the variation in $\vec{M}$, other factors being held constant.

The dynamic phenomena may be best interpreted by first examining the static effects. The macroscopic effect produced by applying a dc magnetic field along the length of a polycrystalline ferrite rod can be determined by integrating the microscopic effects. As the magnetic field is increased, the magnetization vector in each crystallite will deviate from the easy axis and rotate toward the applied field. In general, the greater this deviation, the greater will be the strain. The integrated effect will be a net change in the length of the rod in the direction of the applied field. Depending on the magnitude of the anisotropy energy, there will be a value of the applied field above which there will be no further significant rotation. At this point the sample is said to be saturated. During this magnetization and until saturation is reached, the length of the rod will change.

If a small ac signal is superimposed parallel to the static bias (H_0) the length of the rod will vary sinusoidally to a first-order approximation according to the static curve as shown in Fig. 3. It is clear that above saturation ($H > H_s$) there will be no variation in the length when a small ac signal is applied.

The mechanism for producing an audio change in $\vec{M}$ by a modulated microwave signal is significantly different from superimposing an audio signal on the bias field. In the latter situation the amplitude of mechanical vibration of a rod driven by a constant-amplitude audio signal will be equal to the derivative of the λ vs H curve. However, it is the microwave field transverse to the magnetization vector which produces a change in the projection of $\vec{M}$ along the bias field. As the bias field is increased and resonance is approached, there will be a relatively large precession of the magnetization and a large associated change in its projection. Therefore, as was pointed out previously, the largest output will occur at ferromagnetic resonance, even though an audio drive produces no magnetostriction at such high bias fields.

There is another significant difference in the magnetostriction results produced by the two processes. In a demagnetized state the rod will experience no change in length when driven directly by a small audio signal. However, due to the anisotropy field in the individual crystallites, there is still precession and consequently detection of a microwave by a demagnetized rod. Using the same argument offered by Polder and Smit,[6] by which they explain zero field loss, it is further evident that at zero field the detection will be somewhat larger than when the sample is slightly magnetized. The argument involves the anisotropy field, and the demagnetizing effects due to the presence of domain walls, and shows, using Kittel's[7] resonance conditions, that some ideally shaped domains will be at resonance even at zero applied field.

[5] C. Kittel, "Physical theory of ferromagnetic domains," *Rev. Mod. Phys.*, vol. 21, pp. 541–583; October, 1949.

[6] D. Polder and J. Smit, "Resonance phenomena in ferrites," *Rev. Mod. Phys.*, vol. 25, pp. 89–90; January, 1953.

[7] C. Kittel, "On the theory of ferromagnetic resonance absorption," *Phys. Rev.*, vol. 73, pp. 155–161; January 15, 1948.

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Description

A schematic diagram of the experimental apparatus is shown in Fig. 4. Fig. 5 shows a detecting ferrite device assembled in the form of a long thin rod which is inserted along the axis of a section of round X-band waveguide. The microwave signal is amplitude modulated by a square wave. Power and rf frequency are monitored. The transition from rectangular to round waveguide is followed by a dielectric quarter-wave plate so that a circularly polarized wave is incident on the ferrite rod. The adjustable shorting plug through which the rod enters the guide terminates the microwave system. A laboratory magnet is used to apply dc fields parallel to the length of the rod.

The component of magnetization alternating at the modulation frequency lies along the length of the rod and causes the rod to magnetostrict in a longitudinal mode. The modulation frequency is adjusted to correspond to the fundamental resonant frequency of the rod vibrating in this mode.

In order to observe the effect of tuning out reflections which vary as the permeability changes with bias field, the rod is inserted perpendicularly through the narrow walls of a section of rectangular guide. This section is preceded by a transformer and followed by a termination or an adjustable short.

The motion of the ferrite is observed by bonding to it a rod of polarized barium titanate ceramic. This $BaTiO_3$ crystal exhibits across its electrodes a potential proportional to its state of strain at any instant. The output of the crystal is amplified, filtered, measured, and observed on an oscilloscope.

The ferrite rod is of small diameter so as to minimize the dc demagnetizing factor and to confine the rod to the region of maximum circular polarization. In addition, the thin rod causes minimum distortion of the rf field in the guide. The most significant results were obtained using a ferrite rod about 4 inches long and 0.100 inch in diameter and of the following approximate composition: $Ni_{0.95}\ Co_{0.05}\ Fe_2O_4$. This was fired at about 1300°C.

The $BaTiO_3$ rod is 1.25 inches long and 0.125 inch in diameter. It is of normal ceramic composition (*i.e.*, 4 per cent lead titanate) of commercial origin. Bonding to the ferrite is accomplished by means of a thin layer of a conducting epoxy resin cured for several hours at about 65°C. Fine wire leads are attached to the crystal electrodes by an air-drying silver cement. The minimum amounts of cement and the fine leads are necessary to reduce mechanical loading.

Vibrating in its fundamental mode, the free-free rod has a nodal point only at its center. Since clamping at this point is cumbersome, horizontal suspension on fine threads is preferred. With the threads several inches long, damping of the longitudinal motion is negligible.

All sections of the measuring system have responses linear with amplitude and frequency within the relevant ranges. Also, the magnetostrictive strain at constant applied magnetic bias is linear with the amplitude of audio drive.

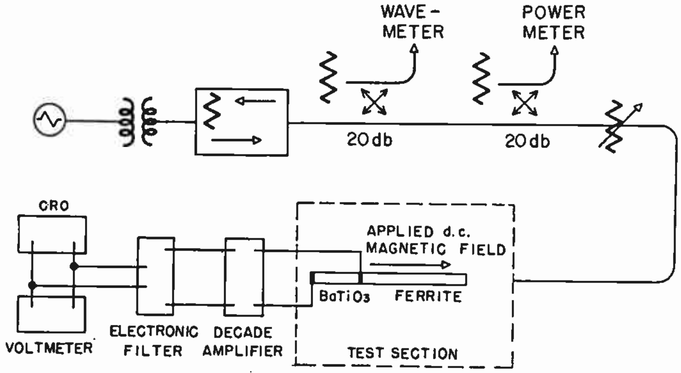

Fig. 4—Diagram of the experimental apparatus.

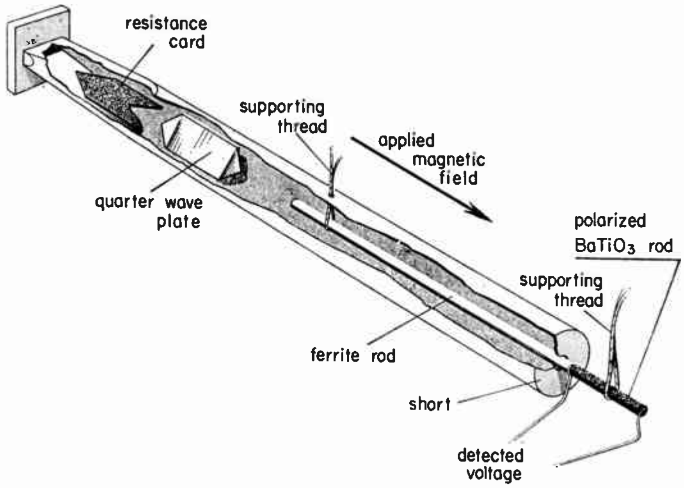

Fig. 5—Ferrite detector for circularly polarized microwaves in cylindrical waveguide.

Results and Discussion

Circular Polarization

The ferrite rod 0.100 inch in diameter was suspended in the circular waveguide and the crystal output was measured as the applied dc magnetic field was varied up to 1500 oersteds. The microwave power was, in all cases discussed, 280 milliwatts. Typical data for such a run is plotted in Fig. 6. The output was observed with decreasing as well as reversed bias fields. The modulation frequency was adjusted for maximum output with each setting of magnet current to compensate for dc changes in the length of the rod.

For one direction of applied field maximum output of 8 millivolts occurred at about 1250 oersteds. As predicted by (15) and the subsequent discussion, the reverse direction of field yielded no corresponding peak. That is, the sense of the circular polarity was in effect reversed by reversing the direction of applied field. Output at zero input power was approximately 0.02 milli-

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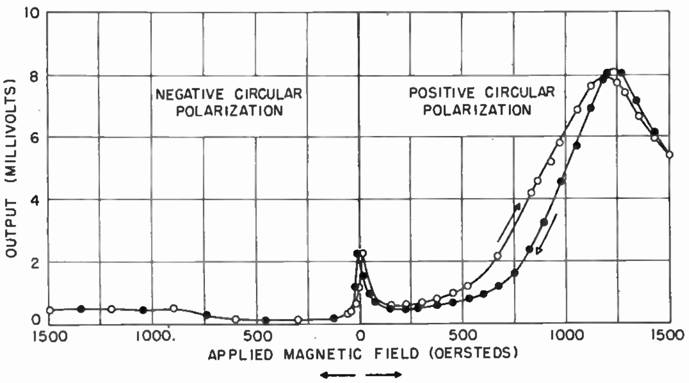

Fig. 6—Detector output as a function of applied dc magnetic field for circularly polarized microwaves (0.100-inch-diameter rod).

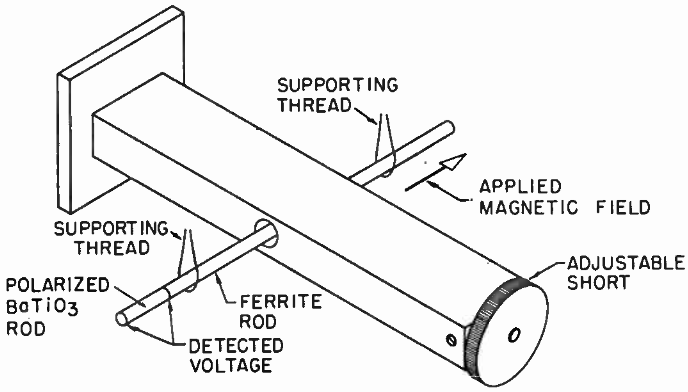

Fig. 8—Ferrite detector in rectangular waveguide with tunable short.

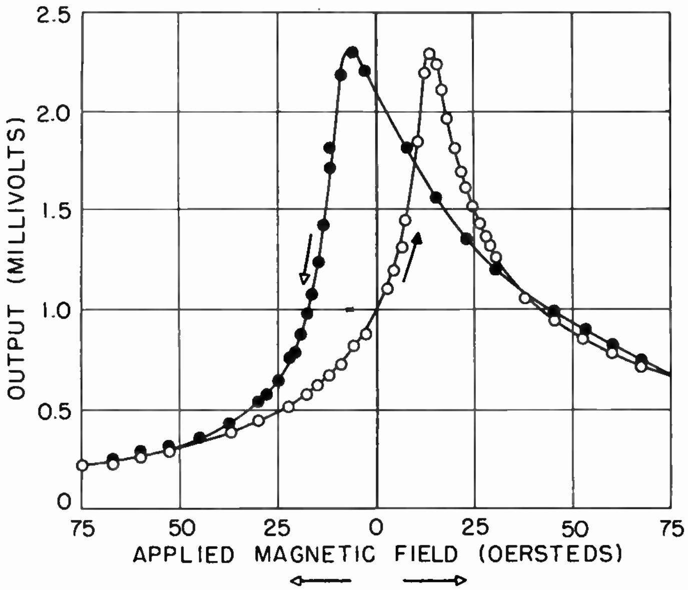

Fig. 7—Detector output as a function of a small applied dc magnetic field for circularly polarized microwaves (0.100-inch-diameter rod).

volt. This defines the noise level referred to below.

The peak which was observed at zero applied field, shown in detail in Fig. 7, was caused by the contributions to the effective field by the anisotropy and demagnetizing fields as already discussed in detail. The hysteresis at these low fields is the ordinary effect observed in ferromagnetic material below saturation.

The hysteresis observed at higher bias fields can possibly be attributed to the nonuniform magnetization of the rod. That is, even at maximum fields part of the rod is not saturated and contributes to the observed hysteresis.

The voltage output was measured as a function of input power, the latter being adjusted by means of the variable attenuator. This relationship was linear over the entire range of powers available.

Linear Polarization

Measurements were made with the microwave signal linearly polarized because higher dc applied fields were available with this type of setup and impedance matching was thereby also made possible.

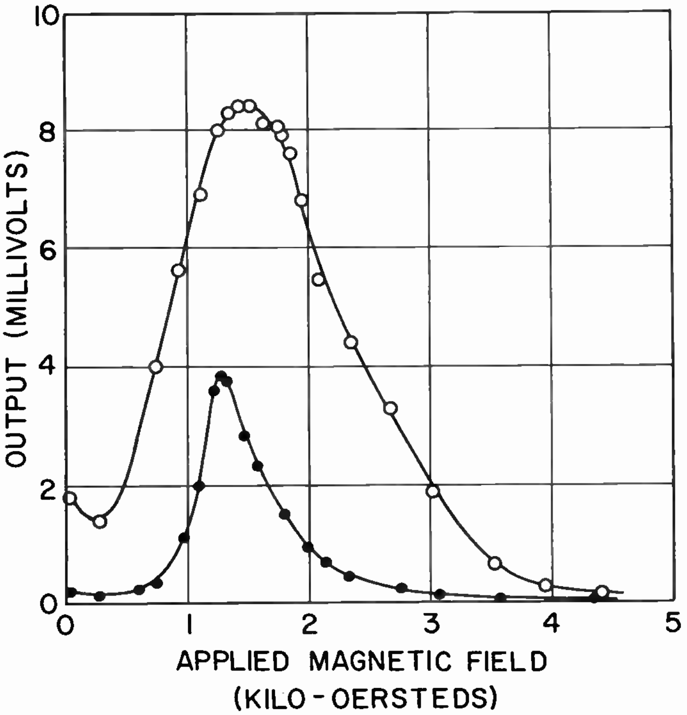

Fig. 9—Detector output as a function of applied dc magnetic field for linearly polarized microwaves matched with transformer (lower curve); matched with transformer and short (upper curve) (0.100-inch-diameter rod).

The same rod, 0.100 inch in diameter and 4 inches long, was inserted through a rectangular waveguide as described above and shown in Fig. 8. With the waveguide terminated by a matched load, the peak output of 4 millivolts occurred at an applied field of 1300 oersteds. These results are given in the lower curve of Fig. 9. The waveguide transformer and modulation frequency were adjusted for maximum output with each setting of magnet current.

An adjustable short was substituted for the load and was used to couple more microwave energy into the ferrite. Because of the better match the peak output was about twice that obtained with the termination. The upper curve of Fig. 9 shows that the maximum was here obtained at a bias field of 1500 oersteds. The Kittel resonance condition for a long thin rod is given by

$$H_{\text{effective}} = \frac{\omega}{\gamma} = H_{\text{applied}} + \frac{4\pi M_s}{2}.$$

$4\pi M_s$ was measured by an independent dc method and found to have a value of 3020 oersteds. This number was used in the above expression (with $\omega/\gamma = 9000/2.8$ oersteds) which gave, for the applied field at ferromagnetic resonance, a value of 1700 oersteds. It may be concluded that the observed peaks are a ferromagnetic resonance effect and that the prediction of the theory has been verified.

The above equation is not expected to hold exactly as applied to this experiment because the rod used is not ideally thin. Further, because the rf field is not uniform along the length of the rod, there is an rf demagnetizing field in this direction which is not taken into account in the equation.

Impedance Matching

The vswr is high with the ferrite rod inserted through the narrow walls of the waveguide, especially about field values corresponding to ferromagnetic resonance. This may be attributed to the large variation of the complex permeability near resonance. The impedance, which is a function of the complex permeability, therefore, is large at bias fields about resonance, and a mismatch results. The consequent reflections are reduced by tuning the transformer and short.

As actually observed, the technique of impedance matching is of some consequence, the greatest output having been achieved for tuning with both the transformer and the short.

Audio Drive

In order to demonstrate the differences between the phenomena of microwave detection and ordinary ac magnetostriction, the ferrite rod 0.100 inch in diameter was demagnetized and driven directly by a field alternating at the audio rate necessary for longitudinal mechanical resonance. Fig. 10 shows these results. It should be noted that the output is in the noise level at zero applied field, passes through a maximum value, and decreases as the sample becomes saturated. This output is found to be in the noise level at bias fields corresponding to ferromagnetic resonance.

Conclusion

It has been demonstrated in the experiments described that the second-order effects in ferrites can be used to demodulate a microwave signal. The experiments have qualitatively verified the theory. Although the voltage outputs derived from the preliminary ferrite-$BaTiO_3$ systems shown here have been on the order of millivolts, the phenomena are sufficiently interesting and promising to merit further investigation.

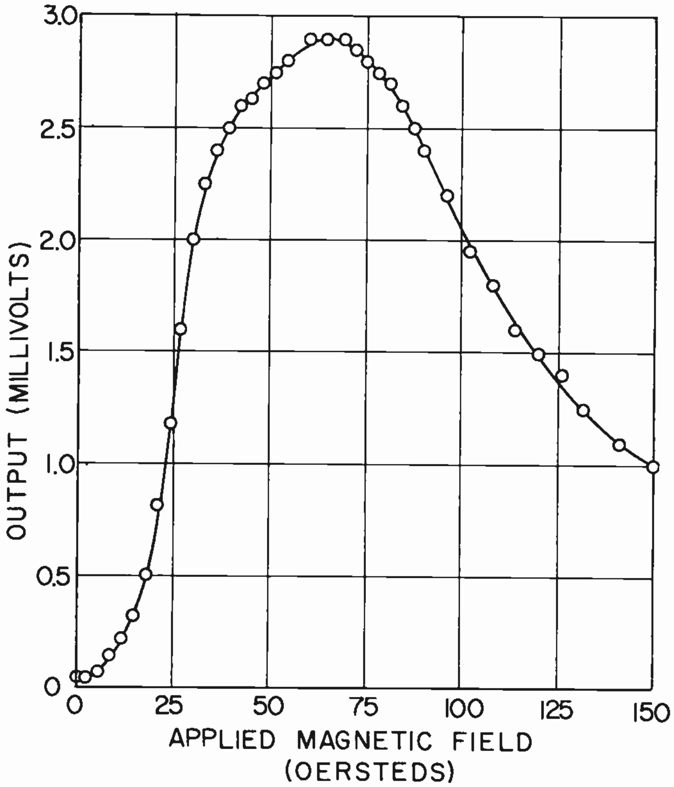

Fig. 10—Magnetostriction of a demagnetized 0.100-inch-diameter ferrite rod as a function of applied dc magnetic field with superimposed audio signal.

There are several factors which can be optimized to improve the performance of the device:

1) A better mechanical impedance match between the ferrite and the crystal will more efficiently transfer the vibrational energy.

2) A more dense ferrite material will have a larger amplitude of vibration for a given drive. The relative density of the ferrite used is 4.6 whereas a value of 5.4 is possible.

3) Materials with larger magnetoelastic coefficients will produce larger oscillations for a given change in the magnetization.

4) Materials exhibiting a narrower resonance line width will, in general, tend to be better detectors since they will show greater "fanning-out" of the magnetization vector at or near resonance.

5) The point of drive has not been optimized. That is, it is possible that if the microwave energy were concentrated at the vibrational node (*i.e.*, the center of the ferrite-$BaTiO_3$ rod) of the system, the output would be increased. Such a setup is easily effected by placing the rod through a hole in the broad face of a reduced height rectangular waveguide.

Even with the above mentioned factors optimized, it seems at present that the greatest value of a ferrite detector will be in the millimeter wavelength region where

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present crystal detectors are not satisfactory. The crystal devices operate on a surface effect while the ferrite detector, being a volume effect device, is not limited by capacitive shunting. The ferrite detector should prove valuable at low temperatures where crystal detectors tend to fail due to separation at the point of contact. It is expected that ferrite detectors can be used with powers up to the order of kilowatts, which is another distinct advantage.

Ferrite of other compositions, as well as new ferromagnetic materials such as garnets, may prove useful in this device and are being investigated. The questions of increased saturation magnetization for improved output at ferromagnetic resonance, variations in anisotropy fields for control of the zero field output, and narrow resonance line width for sharper rf discrimination are under study.

ACKNOWLEDGMENT

The authors wish to thank C. Morrison of DOFL for his invaluable assistance in the refinement of the theory and Dr. C. L. Hogan of Harvard University for his helpful criticism and suggestions. They are also indebted to L. Maxwell and T. Kilduff of DOFL for the polarization and bonding respectively of the $BaTiO_3$ rods, and to R. E. Mundy and I. L. Cooter of the National Bureau of Standards for saturation magnetization measurements.

The Effects of Short Duration Neutron Radiation on Semiconductor Devices*

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Summary—**Transistors, semiconductor diodes, and solid-electrolyte batteries were exposed to short duration, high-intensity neutron radiation from a U-235 critical assembly which was primarily a neutron source. The effect on these components was ascertained by comparing their principal parameters before and after exposure, and in several cases units of operating equipment utilizing these components were monitored during irradiation.**

Of the transistor electrical parameters altered by irradiation, most significant were the decrease in common emitter forward current gain (Beta) and the increase in collector diode reverse leakage current (I_{co}). The higher frequency units were less affected by neutron fluxes than were the audio transistors. High-frequency surface barrier transistors remained virtually undamaged by irradiation up to 10^{13} total neutrons per square centimeter, an exposure which rendered the audio units nearly useless.

Semiconductor diodes suffered an increase in forward resistance and a decrease in back resistance in reasonable accord with the degree of neutron irradiation, while the solid-electrolyte batteries were not permanently affected.

The results obtained are in reasonable agreement with those obtained by others who used pile-type reactors with much longer exposure times, thus indicating that the integrated neutron dosage is of primary significance rather than the rate of exposure.

The degradation of performance of the operating equipment during irradiation was almost entirely attributable to changes in the semiconductor devices utilized therein.

INTRODUCTION

THE USE of semiconductor devices in conjunction with atomic energy sources makes necessary the obtaining of more information concerning effects on these devices from exposure to the various radiations occurring from nuclear reactions.

This report covers experiments with commercially available transistors, diodes, and solid-electrolyte batteries subjected to fission neutron bombardment with a duration of less than one millisecond. A total of about 200 transistors and diodes were exposed to radiation while in both operating and passive conditions; however, the eleven solid-electrolyte batteries were irradiated only when in the passive state.

The neutron source used was the Los Alamos Scientific Laboratories' "Godiva."[1,2] Neutron fluxes to which the devices were subjected were considerably in excess of those tolerable to living tissue. The relatively minor effects of other types of radiation present were not studied in this experiment.

TEST METHODS

Aluminum sheet metal boxes containing units and

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† Diamond Ordnance Fuze Labs., Washington 25, D. C.

[1] H. C. Paxton, "Critical assemblies at Los Alamos," *Nucleonics*, vol. 13, pp. 48–50; October, 1955.

[2] R. E. Peterson and G. A. Newby, "An unreflected U-235 critical assembly," *Nuclear Sci. and Eng.*, vol. 1, pp. 112–125; May, 1956.

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components for test were placed at increasing distances from the critical assembly in positions such that one each would be exposed to nominal integrated neutron fluxes of 10^{13}, 10^{12}, 10^{11}, and 10^{10} total neutrons per square centimeter (NVT) from each burst. Since the aluminum boxes had negligible shielding effect, their contents were considered to have been exposed to the external neutron flux. Both operating transistorized units of equipment and passive unconnected components were included in the boxes so exposed.

The transistorized units which were in operation when irradiated consisted of:

1) Sine-wave oscillators, using type 301 transistors.

2) Single-stage, class "A," amplifiers, using type 2N138 and 2N35 transistors.

3) DC-to-dc (oscillator-rectifier type) power supplies, using type 2N105 transistors.

4) Transistor trigger circuits (two types), using type 302 and 2N35 transistors.

Audio transistors were utilized in these units because of their ready availability, and because the greater resistance to radiation damage of high-frequency transistors was not generally known at this time.

A set of the above units was placed in each box for exposure to the integrated neutron flux at nominal levels indicated. Sine-wave oscillators were not placed in the boxes for exposure at 10^{13} NVT. Instead, they were located at a remote position and their outputs fed through cable connections to the amplifiers in the boxes. This insured that these oscillators would not fail from neutron radiation, and thus the amplifiers would be operationally tested at 10^{13} NVT.

The sine-wave outputs of the oscillators and amplifiers were monitored visually at a remote location by cathode-ray oscilloscopes and also by recording the rectified sine waves (dc) with a strip-chart recorder.

The dc outputs from the power supplies were reduced by voltage dividers and monitored remotely by a vacuum tube voltmeter.

The high-current trigger circuits, either blocking oscillators or two-transistor feedback circuits, were monitored only by being connected to indicator squibs within the boxes.

Components in the passive state during exposure are indicated in Table I.

Equipment and devices tested were subjected to the total neutron irradiation in a single short burst; however, inspection was delayed for about 45 minutes in order to reduce the exposure to personnel from residual radiation. The entire test procedure was repeated three times, using separate sets of unexposed equipment for each test.

TABLE I

Code	Kind	Type	Application
Transistors			
2N77	germanium	*p-n-p* alloy junction	audio, low power
2N132	germanium	*p-n-p* alloy junction	audio, low power
2N35	germanium	*n-p-n* alloy junction	audio, low power
202	germanium	*n-p-n* grown junction	audio, low power
2N139	germanium	*p-n-p* alloy junction	intermediate frequency, low power
SBT	germanium	surface barrier, *p-n-p*	high frequency, low power
904	silicon	*n-p-n* grown junction	medium frequency, low power, high temperature
Diodes			
1N91	germanium	junction, high conductance	general purpose
1N297	germanium	point contact	general purpose
624C	silicon	junction	high voltage
Solid-Ion Batteries			
T-8	solid-ion	95 v at 10^{-9} amperes	special

Test Results and Discussion

Although the sine-wave oscillators showed no decrease in output upon exposure to integrated fluxes up to about 10^{13} NVT, the ability of the amplifiers and power supplies to perform after such irradiation was seriously impaired.

No transient irregularities were observed by the relatively slow-speed monitoring and recording equipment at the time of the neutron burst.

Table II summarizes the results of measurements made on operating units during exposure.

TABLE II

	Percentage output after exposure to nominal NVT of: 10^{13}	10^{12}	10^{11}	10^{10}
Sine wave oscillators, type 301	100*	100	100	100
Amplifiers, type 2N138, 2N35	5–60	40–90	100	100
Dc-to-dc power supplies, type 2N105	15–20	30	85–95	100

* Maximum level here was about 0.34×10^{13} NVT.

Later investigation of the irradiated units showed that damages were confined almost entirely to the semiconductor devices. The transistors suffered a large decrease in common emitter forward current gain (Beta) and an increase in collector diode reverse leakage current (I_{co}), whereas the semiconductor diodes suffered an increase in forward resistance and a decrease in back resistance.

The constant output of the oscillators can be attributed to their design which entailed the providing of a large amount of positive feedback. Although the transistors used in the oscillator circuits suffered a loss of forward current gain (and I_{co} increase) upon irradia-

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tion, because of heavy feedback the oscillator outputs remained unaffected.

The fact that amplifier performance suffered considerably at irradiation of 10^{13} NVT, and was impaired at 10^{12} NVT can be attributed directly to a decrease in current gain of the transistors, as the amplifiers were purposely designed to reflect changes in the transistors' forward current gain. The amplifier design entailed a high impedance ac coupling between driving oscillator and amplifier with a low dc impedance between base and emitter to provide bias stabilization. The performance of simple amplifiers of typical design would not be as adversely affected by loss in current gain of the transistor; if conventional negative feedback were provided only a slight decrease in performance might be expected in most cases.

Because the dc-to-dc power supplies (300 volts, 1 microampere output) were designed to utilize fully the high-forward current gain of an audio type transistor, and because the latter suffered a comparatively high degree of damage from the radiation, considerable degradation of the power supplies was expected. Experimental results shown in Table II bear this out.

It had been anticipated that the high-current trigger circuits (1 a for 1 millisecond) might trigger when irradiated. However, only those exposed to the lower levels of 10^{11} and 10^{10} NVT did so; the indicator squibs connected to the trigger circuits exposed to the 10^{13} and 10^{12} levels were not actuated. This unexpected behavior may be explained by assuming that all circuits were subjected to triggering influences, such as electrical disturbances picked up by the circuit or internal transients caused by the radiation. Thus, the units at the lower levels actually triggered, while those exposed to the higher levels may have been rendered inoperable in less than the triggering time of the circuit. The inoperability could have been caused during neutron build-up by a gradual increase in reverse current leakage through the transistor which would discharge the capacitor power source, or it could have been caused during the neutron burst by a more abrupt radiation damage. Subsequent testing of the trigger circuits showed that those exposed at the lower two levels functioned as before, while those exposed at the 10^{12} and 10^{13} levels were operable only after their transistors were replaced. Measurements showed that the forward current gains of the transistors in the circuits were reduced to an average of 13 per cent of normal from irradiation by 10^{13} NVT, while those exposed to levels of 10^{12} NVT were reduced to about 44 per cent of former values.

A check of the transistors and diodes exposed to radiation when in passive condition showed that most suffered extensive damage upon exposure to neutron radiation of 10^{13} NVT and that some damage occurred upon exposure to 10^{12} NVT.

Fig. 1 and Fig. 2 show the decrease in forward current

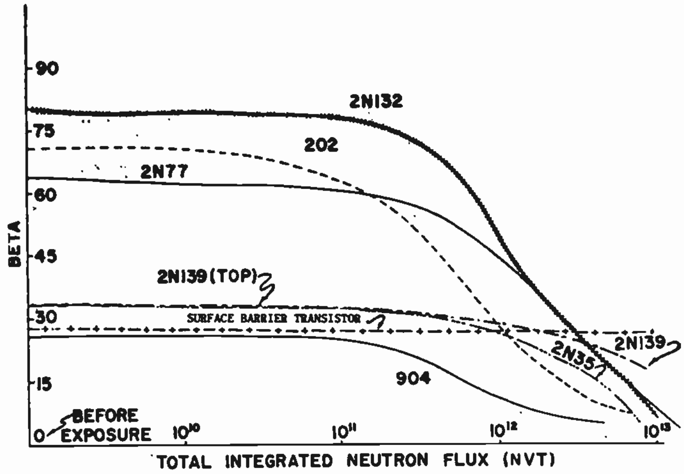

Fig. 1—Decrease in Beta with neutron bombardment.

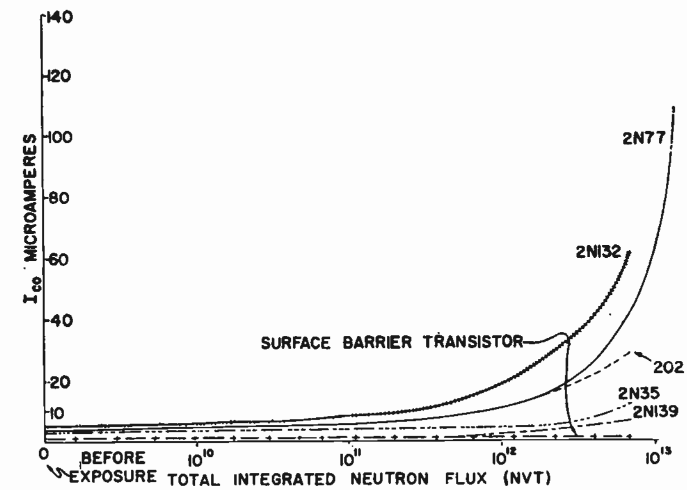

Fig. 2—Increase in I_{co} with neutron bombardment.

gain (Beta) and the increase in backward collector leakage current (I_{co}) of several types of transistors as a result of exposure to different levels of neutron bombardment with the transistors in passive condition.

In analyzing the graphs of Fig. 1 and Fig. 2, it is seen that the higher-frequency 2N139 transistors were less damaged by neutron radiation than were the audio frequency units and that the high-frequency surface barrier units (types SB-100, 2N128, and 2N129) showed no perceptible damage or change of characteristics from exposures as intense as 0.7×10^{13} NVT. Although the decrease in Beta with irradiation was evident to a nearly equal degree in audio *p-n-p* and *n-p-n* transistors, the increase in I_{co} was less for the *n-p-n* units. Silicon transistors (No. 904) showed a loss of Beta (as expected) similar to the germanium audio units. However, the final values of their collector diode reverse leakage currents remained less than 1.0 microampere, a value too low to be plotted on Fig. 2.

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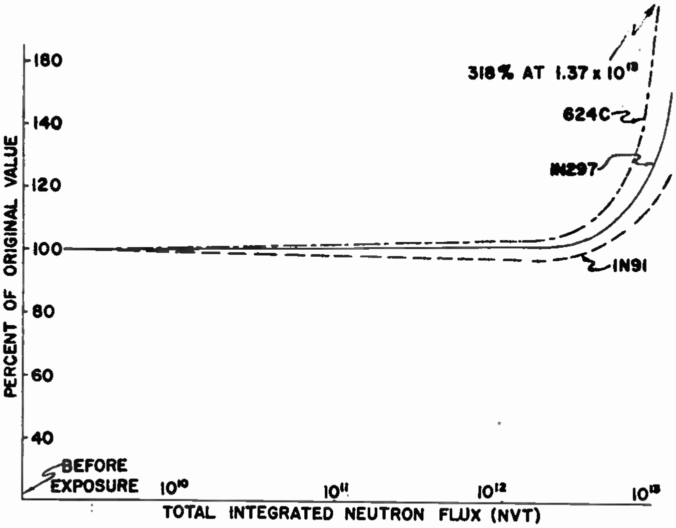

Fig. 3—Increase in forward resistance with neutron bombardment, semiconductor diodes.

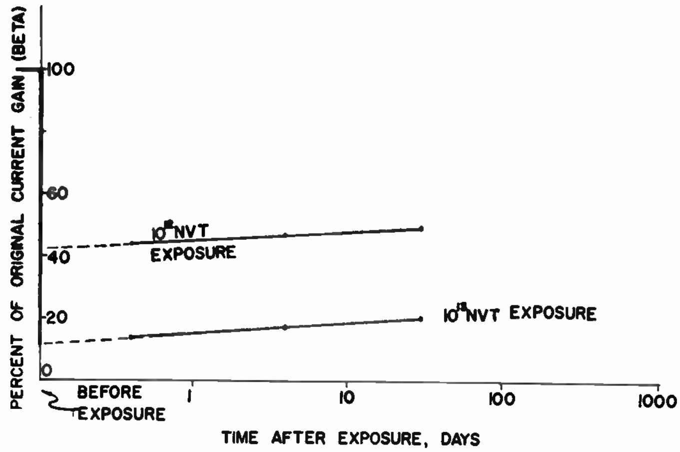

Fig. 5—Recovery of Beta with time, average of 16 audio transistors exposed to approximately 10^{12} and 10^{13} NVT levels.

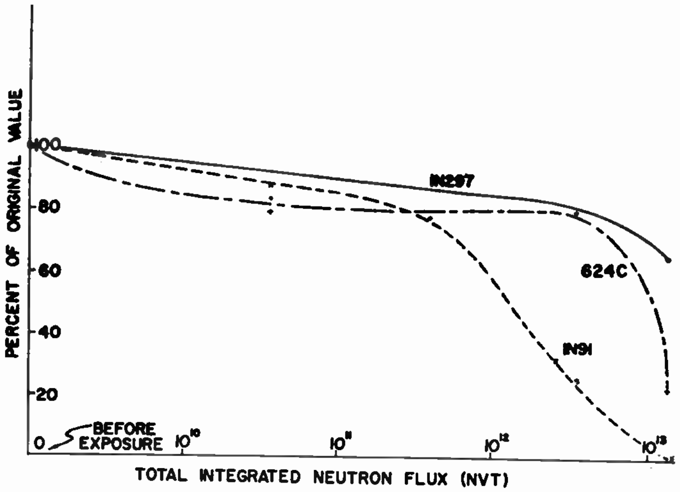

Fig. 4—Decrease in back resistance with neutron bombardment, semiconductor diodes.

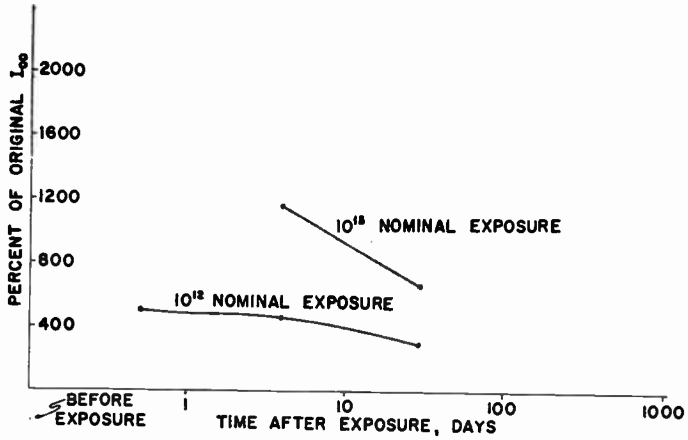

Fig. 6—Decrease in I_{co} with time, average of 16 audio transistors exposed to approximately 10^{13} and 10^{12} NVT levels.

The graphs, Fig. 3 and Fig. 4, show the increase in forward resistance and decrease in back resistance of several types of semiconductor diodes exposed to the neutron bombardment. The Zener voltage of the silicon diodes tested remained almost unchanged by radiation.

Neutron radiation appeared to have negligible permanent effect on the output voltage of solid-electrolyte batteries exposed at levels up to 1.3×10^{13} NVT. Monitoring of the batteries during irradiation indicated a maximum drop in their output voltage of about 15 per cent immediately following the burst. Precise evaluation of the short-term performance of the batteries was difficult because of their high internal resistance and the shunt capacitance of the connecting cable.

Some degree of recovery of performance characteristics was observed after the irradiated transistors were stored at room temperature for two months, the most noticeable change being the decrease in I_{co} with aging. Fig. 5 and Fig. 6 show the Beta recovery and the I_{co} decrease of neutron-damaged transistors with time.

Fig. 7 shows the change of h parameters of one of the audio types of transistors tested. Results are the average of six transistors exposed at each level.

These over-all results are in reasonable agreement both with those obtained by other investigators[3,4] who utilized lower level continuously operated radiation sources, and with the theoretically expected behavior predicted for semiconductor devices.

Decreases in the forward current gains (Betas) of transistors as a result of neutron irradiation have been experienced by others[3,4] and are compatible with the theory of transistor action. The decreases are attributed to a reduction in minority carrier lifetime, which results from bombardment-induced displacements in the crystal lattice.[5,6] The fact that the thin base high-frequency

[3] J. C. Pigg and C. C. Robinson, "Effect of radiation on semiconductors," *Elec. Mfg.*, vol. 59, pp. 116–124; April, 1957.

[4] G. L. Keister and H. V. Stewart, "Preliminary Report of an Investigation of the Effects of Nuclear Radiation on Selected Transistors and Diodes," Contract AF-33 (038) 19589.

[5] K. Lark-Horovitz, "Conductivity in semiconductors," *Elec. Eng.*, vol. 68, pp. 1047–1056; December, 1949.

[6] A. D. Kurtz, S. A. Kulin, and B. L. Averbach, "Effect of dislocations on the minority carrier lifetime in semiconductors," *Phys. Rev.* vol. 101, pp. 1285–1291; February 15, 1956.

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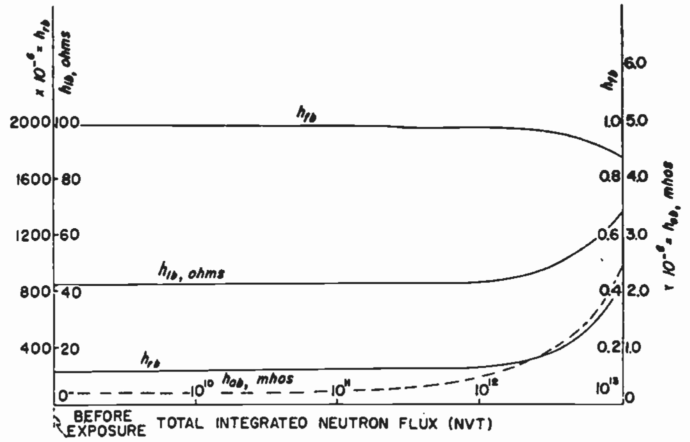

Fig. 7—Change in h parameters with neutron bombardment, average of 24 2N77 transistors.

transistors suffered less decrease in Beta than the audio units may be attributable to the radiation caused reduction of minority carrier lifetime which is of less proportional importance when the carriers diffuse a short distance through a narrow base region.

The possibility of a decrease in transistors' forward current gain as a result of changes in bulk conductivity from neutron irradiation has been considered. This is believed to be an effect of lesser importance, since the loss of transistor current gain experienced was much greater than that which would result from this process.

The increase in collector diode backward leakage current (I_{co}) of irradiated transistors has also been experienced by investigators using other type reactors.[3,4] This degradation is attributed to a combination of surface effects at the collector-base junction together with changes in bulk characteristics of the semiconductor which tend to destroy the barrier. The former effects, of conductances shunting the collector-base barrier, have been considered[3,4,7] to result primarily from exposure to gamma radiation, and were reported to vary considerably with the type of grease or filling compounds surrounding the germanium; whereas changes of bulk characteristics have been attributed principally to neutron irradiation.

The slight recovery of both Beta and I_{co} after two months of storage at room temperature was expected, since it is known that recovery will occur from both surface and bulk damage by annealing. Bulk damage recovery by annealing at room temperature is reported[7] to be quite slow.

In view of the observed increases in I_{co} from neutron irradiation and of the subsequent partial recovery, it would appear either that neutron radiation is of more consequence in causing surface effects than formerly assumed, or that the bulk damage annealing rate at 25–30°C is appreciable.

The discussion covering the increase in I_{co} of transistors is applicable to the loss of back resistance in diodes as indicated in Fig. 4. The increase in forward resistance of diodes as shown in Fig. 3 is believed due to bulk effects which reduce the conductivity of the n-type material.

Conclusion

Audio frequency transistors suffer a considerable decrease in forward current gain (Beta) and a large increase in backward collector current (I_{co}) upon irradiation of 10^{13} total neutrons per square centimeter (NVT), and are damaged to a lesser extent by exposure to 10^{12} NVT. Loss of Beta is attributed primarily to a decrease in minority carrier lifetime, while I_{co} increase is considered to be due to both surface and bulk effects.

Higher frequency transistors are damaged to a lesser extent by neutron radiation at these levels, while the high-frequency surface barrier transistors appear to suffer no damage or change of characteristics under identical conditions of exposure.

The irradiation of semiconductor diodes at the 10^{13} NVT level resulted in an increase in the forward resistance of the diode and a decrease in back resistance.

The results obtained are in reasonable agreement with those obtained by others who used pile-type reactors with longer exposure times, thus indicating that the integrated neutron dosage is of primary significance, rather than the rate of exposure.

Since the neutron radiation damage to operating equipment was manifest chiefly in the semiconductor devices, it is believed that by using special circuit design (employing negative feedback, for example) to compensate for the changes in semiconductor parameters many of the units could be made to operate, even though the performance capabilities of the semiconductor devices would be reduced.

The neutron radiation damage to the semiconductor devices appears to be relatively permanent. After two months of storage at room temperature only a slight recovery of the devices was noted.

[7] D. B. Kret, "Analysis of radiation effects on transistors," *Electronic Design*, vol. 5, pp. 28–31; July 15, 1957.

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Optimum Filters with Monotonic Response*

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Summary—A class of filters is developed whose amplitude characteristic has no ripple in the pass band and a high rate of attenuation in the stop band; thus it combines the desirable features of the Butterworth and Tchebycheff response. Among all filters of a given order this new class has the maximum cutoff rate under the condition of a monotonically decreasing response.

THE AMPLITUDE characteristic of a filter can be written in the form

$$A(\omega) = \frac{A_0}{\sqrt{1 + f(\omega^2)}}, \tag{1}$$

where $f(\omega^2)$ is a positive rational function of ω^2; if the network function has no finite zeros then $f(\omega^2)$ is a polynomial and the resulting filter is easily realizable by an LC ladder with a resistance at one or both ends. Two classes of such filters have been discussed in the literature: the Butterworth (B) class given by

$$f(\omega^2) = \omega^{2n} \tag{2}$$

and the Tchebycheff (T) class given by

$$f(\omega^2) = \tfrac{1}{2}C_n(2\omega^2 - 1) + \tfrac{1}{2} = C_n{}^2(\omega), \tag{3}$$

where $C_n(\omega)$ is the Tchebycheff cosine polynomial. If the only requirement on $A(\omega)$ is to have the maximum possible attenuation for a given variation in the pass band, then the T filter is the optimum.[1] However, in many applications; *i.e.*, when the transient response is also considered, a high ripple in the pass band is not tolerated;[2] one then uses as a simple compromise the B filter whose cutoff properties are not too good. It is natural, therefore, to search for a filter that has the desirable features of the B filter with a faster rate of cutoff. In this paper we shall determine the class L of filters whose amplitude characteristic decreases monotonically with ω and has the greatest possible cutoff rate.

Denoting by $L_n(\omega^2)$, the polynomial generating the L filter, we have

$$\frac{dL_n(\omega^2)}{d\omega} > 0. \tag{4}$$

If we further assume

$$L_n(1) = 1 \tag{5}$$

our problem is to find among all positive polynomials of degree n satisfying (4) and (5) the one whose slope

$$\frac{dL_n}{d\omega}$$

at the point $\omega = 1$, is maximum; the discussion will be limited to odd values of n

$$n = 2k + 1. \tag{6}$$

In the Appendix it is shown[3] that the polynomials L_n are given by

$$L_n(\omega^2) = \int_{-1}^{2\omega^2-1} v^2(x)dx, \tag{7}$$

where

$$v(x) = a_0 + a_1P_1(x) + \cdots + a_kP_k(x) \tag{8}$$

$$a_0 = \frac{a_1}{3} = \frac{a_2}{5} = \cdots = \frac{a_k}{2k+1} = \frac{1}{\sqrt{2}(k+1)} \tag{9}$$

and the P_k's are the tabulated[4] Legendre polynomials of the first kind. The slope at the point $\omega = 1$ is given by

$$\left.\frac{dL_n(\omega^2)}{d\omega}\right|_{\omega=1} = 2(k+1)^2. \tag{10}$$

Since

$$P_1(x) = x, \qquad P_2(x) = \tfrac{1}{2}(3x^2 - 1)$$

we can readily obtain from (7)–(9),

$$L_3(\omega^2) = 3\omega^6 - 3\omega^4 + \omega^2$$

and

$$L_5(\omega^2) = 20\omega^{10} - 40\omega^8 + 28\omega^6 - 8\omega^4 + \omega^2.$$

The polynomials L_3 and L_5 are plotted in Fig. 1 for $0 \le \omega \le 2$ and the B polynomials ω^6 and ω^{10} are shown for comparison; in Fig. 2 the corresponding amplitude characteristics are shown.

We shall next realize the L filters for $n=3$ and 5.

$n=3$: To determine the network function $H(p)$ whose amplitude equals $A(\omega)$ we form

$$h(p^2) = H(p)H(-p)$$

* Original manuscript received by the IRE, August 29, 1957.
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[1] A. Papoulis, "On the approximation problem in filter design," 1957 IRE NATIONAL CONVENTION RECORD, pt. 2, pp. 175–185.

[2] K. Küpfmüller, "Die Systemtheorie der Electrischen Nachrichtenübertragung," S. Hirzel, Zürich, Germany; 1949.

[3] S. Bernstein, "Leçons sur les Propriétés Extrémales et la Meilleure Approximation des Fonctions d'une Variable Réelle," Gauthier-Villars, Paris, France; 1926.

[4] E. Jahnke and F. Emde, "Tables of Functions," Dover Publications, New York, N. Y.; 1945.

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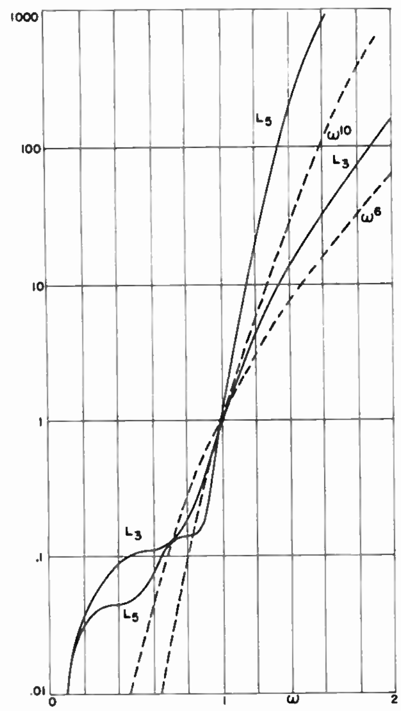

Fig. 1.

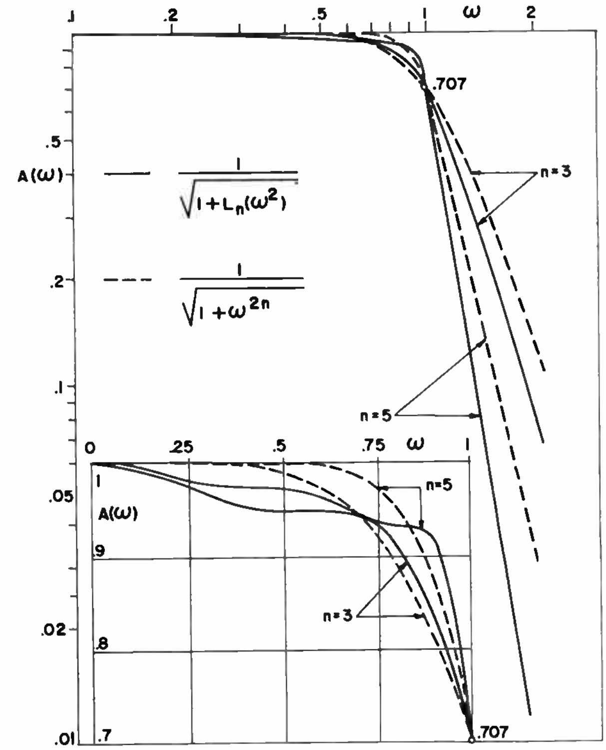

Fig. 2.

then

$$h(-\omega^2) = H(j\omega)\,H(-j\omega) = A^2(\omega)$$

therefore,[5]

$$h(p^2) = \frac{1}{1 + L_3(-p^2)} = \frac{1}{1 - 3p^6 - 3p^4 - p^2}.$$

Factoring $h(p^2)$ and retaining the left-hand plane poles we obtain

$$H(p) = \frac{0.577}{p^3 + 1.310p^2 + 1.359p + 0.577}$$

whose poles are given by

$$p_0 = -0.620 \qquad \begin{matrix} p_1 \\ \bar{p}_1 \end{matrix} = -0.345 \pm j0.901$$

and are shown in Fig. 3.

It might be of interest to show that $H(p)$ can be obtained from the quadratic factors of $h(p^2)$ without the actual evaluation of its roots; this is useful since in the usual methods of numerical evaluation of complex roots one finds first the quadratic factors. Indeed, if

$$p^2 + Ap + B$$

is the Hurwitz factor of

$$p^4 + ap^2 + b,$$

[5] A. Papoulis, "Frequency transformations in filter design," IRE Trans., vol. CT-3, pp. 140–144; June, 1956.

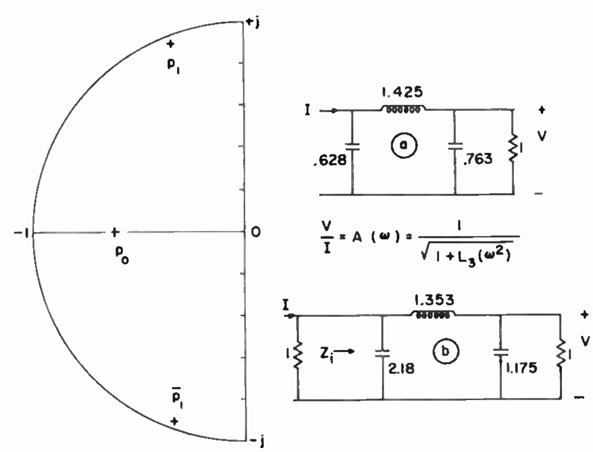

Fig. 3.

then it can easily be seen that

$$B = \sqrt{b} \qquad A = \sqrt{2B - a}.$$

To realize $H(p)$ as a ladder terminated into one ohm resistance we have (see Fig. 3)

$$H(p) = \frac{V}{I} = \frac{Z_{12}}{1 + Z_{22}}$$

hence,

$$Z_{22} = \frac{1.310p^2 + 0.577}{p^3 + 1.359p}$$

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expanding Z_{22} into a continuous fraction we obtain the network of Fig. 3(a).

To realize $H(p)$ as a ladder with a one ohm resistance at both ends, we first determine the reflexion coefficient defined by

$$r(p) = \frac{1 - Z_i(p)}{1 + Z_i(p)},$$

where $Z_i(p)$ is shown in Fig. 3. It can be seen that

$$|r(j\omega)|^2 = 1 - \frac{A^2(\omega)}{A_0^2} = \frac{f(\omega^2)}{1 + f(\omega^2)}$$

therefore,

$$r(p)r(-p) = \frac{-3p^6 - 3p^4 - p^2}{1 - 3p^6 - 3p^4 - p^2}$$

from which we obtain $r(p)$ retaining the Hurwitz factors

$$r(p) = \frac{p^3 + 0.392p + 0.577}{p^3 + 1.310p^2 + 1.359p + 0.577}$$

knowing $r(p)$ we can determine $Z_i(p)$ from

$$Z_i(p) = \frac{1 - r(p)}{1 + r(p)}$$

thus,

$$Z_i(p) = \frac{918p^2 + 782p + 577}{2000p^3 + 1702p^2 + 1936p + 577}.$$

$Z_i(p)$ now can be realized as a reactive network terminated into a resistance; however, since $A(\omega)$ has no finite zeros the same is true for the real part of $Z_i(p)$, therefore, $Z_i(p)$ can be expanded into a continuous fraction and the network of Fig. 3(b) results.

$n=5$: We have

$$h(p^2) = \frac{1}{1 + L_5(-p^2)}$$

$$= \frac{1}{1 - 20p^{10} - 40p^8 - 28p^6 - 8p^4 - p^2}$$

hence,

$$H(p) = \frac{0.224}{p^5 + 1.551p^4 + 2.203p^3 + 1.693p^2 + 0.898p + 0.224}$$

and its poles are given by

$$p_0 = -0.468, \frac{p_1}{\bar{p}_1} = -0.388 \pm j0.589, \frac{p_2}{\bar{p}_2} = -0.154 \pm j0.968.$$

From $H(p)$ we obtain

$$Z_{22}(p) = \frac{1.551p^4 + 1.693p^2 + 0.224}{p^5 + 2.203p^3 + 0.898p}.$$

From the continuous fraction expansion of $Z_{22}(p)$ the network of Fig. 4 results.

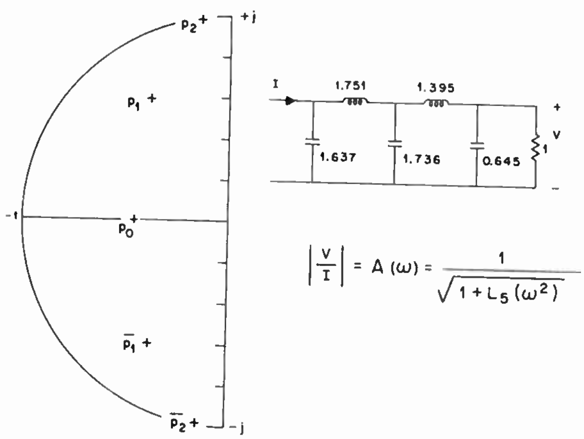

Fig. 4.

To conclude we shall list the important properties of the B, T, and L polynomials.

B polynomials: maximally flat, monotonic, cutoff slope equals $2(2k+1)$.

T polynomials: equal ripple in band-pass, in the class of all polynomials they have the maximum cutoff slope given by[1] $2(2k+1)^2$.

L polynomials: monotonic, in the class of all monotonic polynomials they have the maximum cutoff slope given by $2(k+1)^2$.

Appendix

To prove (7) to (10), we shall first solve the following problem.[6] Consider the class C_n of polynomials $f(x)$ of degree $n=2k+1$ such that

$$f(-1) = 0, \quad f(1) = 1, \quad \frac{df}{dx} \geq 0, \quad -1 \leq x \leq 1. \tag{11}$$

Determine in C_n a polynomial $F(x)$ such that

$$\left.\frac{dF}{dx}\right|_{x=1} = M \tag{12}$$

is maximum.

With

$$\frac{dF}{dx} = \phi(x)$$

we have

$$\phi(1) = M, \quad \int_{-1}^{+1} \phi(x)dx = 1, \quad F(x) = \int_{-1}^{x} \phi(\xi)d\xi,$$

$$\phi(x) \geq 0. \tag{13}$$

We shall first show that the polynomial $\phi(x)$, whose degree is $2k$, is a perfect square; clearly since $\phi(x)$ is

[6] This problem and its proof is a modified version of a theorem proved by Bernstein, *loc cit.*

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nonnegative, all its roots in the $(-1, +1)$ interval are double hence it can be written in the form

$$\phi(x) = u^2(x)q(x), \tag{14}$$

where $q(x)$ is an even polynomial without interior roots, such that

$$q(x) > 0 \quad \text{for} \quad -1 < x \leq 1. \tag{15}$$

At $x = -1$, $q(x)$ might equal zero; in this case

$$\left.\frac{dq}{dx}\right|_{x=-1} > 0. \tag{16}$$

If $q(x)$ were not a constant, then its degree would be at least two, therefore, the polynomial

$$q(x) - \epsilon(1 - x^2)$$

would be of the same degree as $q(x)$, and for ϵ, sufficiently small, it would remain positive; this is obvious [see (15)] if $q(-1) \neq 0$ and it can easily be established [see (16)] if $q(-1) = 0$. Therefore, if the constant A is so chosen that

$$A \int_{-1}^{+1} u^2(x)[q(x) - \epsilon(1 - x^2)]dx = \int_{-1}^{+1} u^2(x)q(x)dx = 1$$

then,

$$A > 1 \tag{17}$$

and the polynomial

$$A \int_{-1}^{x} u^2(\xi)[q(\xi) - \epsilon(1 - \xi^2)]d\xi$$

will be in C_n and its derivative at the point $x = +1$ will be given by

$$Au^2(1)q(1) = AM,$$

which is impossible [see (17)] since M was assumed to be maximum. Therefore, $q(x)$ is a constant which can be taken as one. Thus

$$\phi(x) = u^2(x), \qquad u(1) = \sqrt{M} \tag{18}$$

and

$$\int_{-1}^{+1} u^2(x)dx = 1 \tag{19}$$

To determine the polynomial $u(x)$ whose degree is k, we first expand it into a series[7]

$$u(x) = a_0 + a_1P_1(x) + \cdots + a_kP_k(x) \tag{20}$$

of Legendre polynomials; these polynomials satisfy

$$\int_{-1}^{+1} P_k(x)P_h(x)dx = \begin{cases} 0 \text{ for } h \neq k \\ \dfrac{2}{2k+1} \text{ for } h = k. \end{cases} \tag{21}$$

From (18) and (20) we obtain

$$a_0 + a_1 + \cdots + a_k = \sqrt{M}, \tag{22}$$

since $P_k(1) = 1$, and from (19)–(21)

$$2a_0^2 + \frac{2}{3}a_1^2 + \cdots + \frac{2}{2k+1}a_k^2 = 1 \tag{23}$$

follows. Thus our problem is to determine the constants $a_0, a_1, \cdots, a_k$ such as to maximize (22) under the constraint (23); this can be done easily and the result is given by

$$a_0 = \frac{a_1}{3} = \cdots = \frac{a_k}{2k+1} = \frac{1}{\sqrt{2}(k+1)}. \tag{24}$$

The constant M can readily be evaluated from (22) and (24)

$$M = \frac{(k+1)^2}{2}. \tag{25}$$

$L_n(x)$ can readily be obtained from $F(x)$ by an interval transformation

$$L_n(x) = F(2x - 1) = \int_{-1}^{2x-1} u^2(\xi)d\xi \tag{26}$$

and its slope at $x = +1$ is given by

$$\left.\frac{dL_n}{dx}\right|_{x=1} = 2M = (k+1)^2 \tag{27}$$

with $\omega^2 = x$ (7) to (10) follow readily; the assumption $L_n(0) = 0$ is not restrictive.

[7] R. Courant and D. Hilbert, "Methoden der Mathematischen Physik—I," Springer, Berlin, Germany; 1937.

World Radio History

Correspondence

Unusual Propagation at 40 MC from the USSR Satellite*

Recordings of signals from the first Soviet satellite were made at the Derwood Experimental Laboratory, Carnegie Institution of Washington, Department of Terrestrial Magnetism, from October 4 to October 25, 1957. The recording instrument was a conventional phase-switching interferometer, as used in radio astronomy, operating at 40.002 mc. Receiving antennas were folded dipoles spaced 11 wavelengths apart along an east-west line. These were changed on October 9 to a spacing of approximately 15 wavelengths along the axis of the satellite's orbit, that is, inclined 65° to the equatorial plane.

The instrument was originally designed for use at 38 mc with a bandwidth of 150 kc but was hastily converted to monitor 40.002 mc shortly after the first announcement of the launching. Subsequent reductions in bandwidth of the receiver to about 3 kc were helpful in minimizing interference. Normal sensitivity was such that signals at the receiver of approximately 10^{-16} watts would saturate or drive the recording pen off-scale. The satellite signals were normally saturating except in a few isolated cases which may have been caused by receiver detuning.

A *movement* of any signal source through the receiver's antenna pattern produces the well-known interference pattern with maxima and minima resulting from the addition or cancellation of the combined signals from the two antennas. As used in radio astronomy, the earth's rotation provides the movement as the interference pattern sweeps across a radio star. However, for recording satellite signals, the radio source moves rapidly from horizon to horizon in a few minutes producing interferometer traces which are characteristic and free from confusion with other signals.

The apparatus was operated continuously, thereby providing background data on radio stars, solar activity, and interference, in addition to the satellite signals. The central time—or time of closest approach—for each transit is plotted in Fig. 1. Each dot represents a complete recording. The lines connecting families of dots are separated by time intervals of approximately 96 minutes as established by the satellite's orbital period.

It was normal to record about three successive transits of the morning series but only two for the evening passes. This is reasonable in view of the known difference in altitude of the satellite at these times. Between October 10 and 14, however, several additional transits were recorded, extending the morning group. These appear to fall into normal transit families as joined by the light lines. No extended ranges were observed in the evening group.

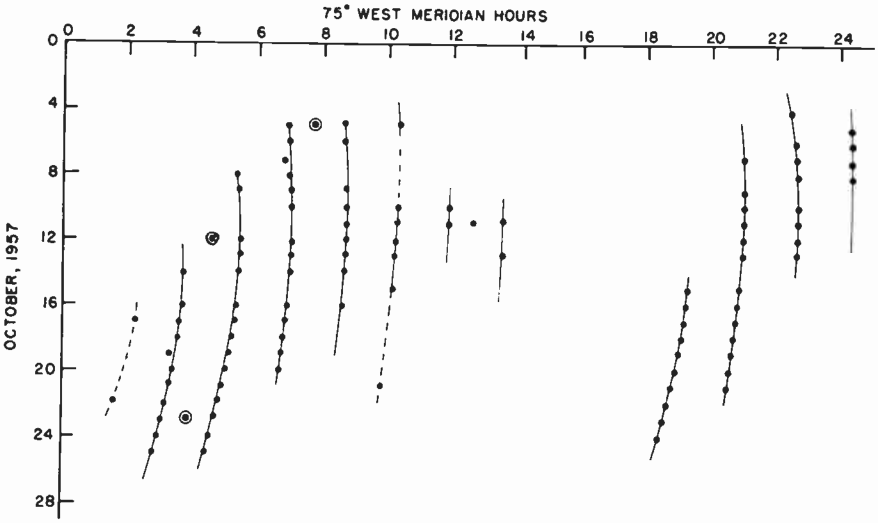

Fig. 1—Reception of Sputnik I at 40.002 mc, Derwood, Md.

But there were also a few distinct recordings which produced the satellite's characteristic interference pattern at times which were centered between normal transits. Occurrences are evident in Fig. 1 on October 5, 12, and 23. *At these times the satellite was on the opposite side of the earth near an antipodal point.* Apparent movement of the satellite's image was normal and the signals were strong enough on one occasion (October 5) to saturate the receiver. In all respects, the image signals appeared to originate in a "point" source similar to the actual satellite.

The recordings between 10 and 14 hours of Fig. 1 are "beyond the horizon." Some of the signals may have originated even west of the Hawaiian Islands. Undoubtedly the mode of propagation was via normal F-layer reflection with perhaps an occasional contribution from sporadic E-region ionization. Long-distance communication at these frequencies is not uncommon at this phase of the sunspot cycle. However, the fact that the characteristic interference pattern was preserved after the signals had bounced back and forth between earth and ionosphere is evidence of very limited dispersion over the propagation path. Since the angular separation between lobes is about 4°, the satellite signals originated from a source much smaller than 4°.

Although "images" are well established in optical and electrical theory, it is difficult to offer a very satisfactory explanation of the "image" or "ghost" satellite recordings. There may have been a few cases in these recordings of extremely stable ionospheric conditions which permitted a focusing of energy from all directions into a small "hot spot" on the opposite side of the sphere. Energy from this concentrated area, presumably in the ionosphere, was then radiated or scattered to the receiving antennas so that the "hot spot" in motion was recorded with the same basic properties as the direct signals from the satellite.

The 40-mc radio signals from the second satellite were monitored in a similar fashion. No unusual events were noted during a period of somewhat less than a week in early November.

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* Received by the IRE, January 9, 1958. Results given here were reported at a meeting of the Washington Academy of Sciences, December 19, 1957

A Note on Some Signal Characteristics of Sputnik I*

Soon after the launching of Sputnik I a simple interferometer antenna was erected at the Ohio State University Radio Observatory. The interferometer consisted of two horizontal half-wave dipoles separated by about 350 feet on an east-west base line so that the minimum lobe separation was at the meridian. Knowing the angle of the orbit and satellite velocity from other sources, the slant distance to the satellite as it crossed the meridian could then be obtained by noting the minimum time between nulls in the interferometer pattern, as given by

$$d = \frac{vt \sin 34°}{\tan \alpha}$$

* Received by the IRE, December 10, 1957.

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where

d = slant distance (miles),
v = velocity of satellite (miles per second),
$34°$ = angle of satellite with respect to meridian at Columbus,
α = minimum angle between nulls of interferometer pattern,
t = minimum time observed for satellite to travel between nulls (seconds).

By noting also the time of nearest approach of the satellite (field strength maximum or rate of frequency change maximum), the distance between the point of nearest approach and the point of meridian crossing could be determined. From this the horizontal (north or south) distance to the point of meridian crossing could be obtained and with the slant distance as determined above the vertical height of the satellite at meridian crossing could then be calculated. Obviously this simple method is accurate only for relatively close passes of the satellite.

Using this procedure the heights of Sputnik I were measured from its 20-mc signal. During the first two weeks after launching the heights measured during morning passes (path NNW to SSE) averaged 355 statute miles and during the evening passes (path SSW to NNE) averaged 175 miles.

While listening to the 20-mc signals during the morning observations, it was noted that as the satellite was approaching, the signal level was very unsteady and subject to rapid fluctuations in strength. However, shortly before the satellite reached its point of nearest approach the signal suddenly (within 1 or 2 seconds) became very steady and, except for a gradual modulation due to its spin, remained so until the signal faded out. The transition from the unsteady or rough signal to the steady or smooth signal is illustrated by the oscillographs in Fig. 1. The upper trace was taken a few seconds before the transition while the lower was taken a few seconds after the transition. The upper trace shows numerous large, abrupt decreases in signal level of the order of 20-milliseconds duration or less. These oscillographs are from the magnetic sound tape of the pass of Sputnik I on October 13, 1957, with the time of nearest approach at 6:56 A.M., EST.

A frequency vs time graph of this same pass is shown in Fig. 2. The ordinate is the frequency difference between the local crystal-controlled beat-frequency oscillator (bfo) and Sputnik I's frequency. Since the local bfo frequency was *higher* than Sputnik I, the Doppler decrease in frequency of Sputnik I as the satellite passed resulted in an *increase* in the frequency difference as noted in Fig. 2. The point (A) at which the signal characteristics change from rough to smooth is seen to occur near the lower knee of the curve, being 69 seconds before the point (B) of nearest approach. The frequency curve was measured using an events-per-unit time counter which sampled one-second intervals every 6 seconds. The low counts as noted near the top of the curve were due to occasional slow deep fading of the signal, the resultant signal strength being insufficient to actuate the counter for a full tally.

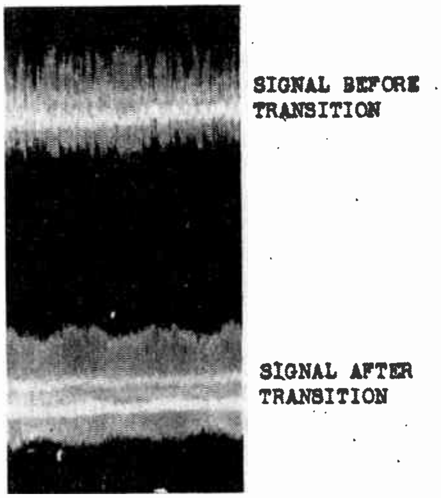

Fig. 1—Cathode-ray oscillographs of the signal of Sputnik I during its close approach to Columbus, Ohio, on the morning of October 13, 1957. The oscillographs show an abrupt change in the signal characteristics occurring at the point (A) of Fig. 2, about one minute before the point of nearest approach. The length of the time axis is 2.7 seconds.

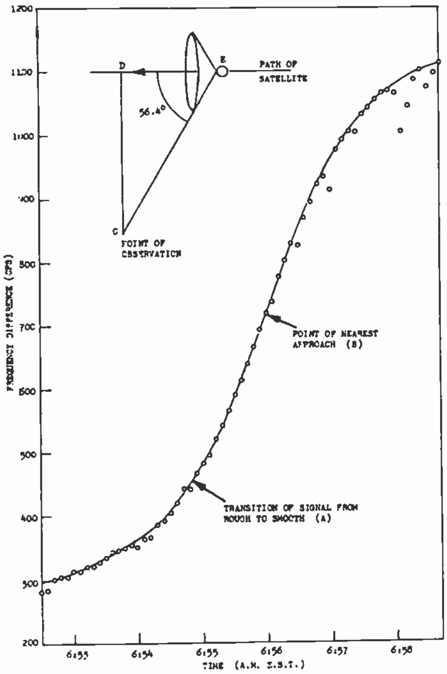

Fig. 2—Graph of (beat) frequency difference during pass of Sputnik I on the morning of October 13, 1957, with arrows showing transition point (A) from rough and unsteady signal to smooth steady signal and point (B) when satellite was at its point of closest approach.

The rough to smooth transition described above was noted repeatedly on other passes of the satellite with the transition always occurring near the lower knee of the frequency-time curve as in Fig. 2. No later second transition from smooth back to rough was noted at any time. Accordingly, it appears that while the satellite is approaching (at point E in insert in Fig. 2) with the angle between the line of observation (line CE) and the satellite path (line ED) less than about 60°, the signal traverses an unstable medium. However, as the satellite moves further to the left and the angle increases to more than about 60°, the signal traverses a more stable medium with the received field strength varying only in a smooth gradual manner. It is as though there is a cone of about 60° half angle preceding the satellite (see insert in Fig. 2) through which transmission is unstable. For the specific case of Fig. 2 the time between points (A) and (B) on the curve is 69 seconds. Taking the satellite velocity as 4.65 statute miles per second this makes the distance DE equal to 321 miles.

The maximum rate of change of frequency (at B) is 4.81 cps so that the slant distance CD of nearest approach[1,2] is 482 miles. Thus, the cone half angle, CED, is 56.4° as indicated in the figure.

Since the phenomenon is unsymmetrical, occurring only near nose-on aspects of the satellite, a possible explanation is that as the satellite encounters gas molecules it bounces them ahead while at the same time ionizing them. Due to various factors, the ions spread out into a cone ahead of the satellite, and because of inhomogeneities of the ion density there may be refraction and rapidly varying multipath transmission through this region resulting in rapid signal strength fluctuations. A more detailed analysis of the phenomenon is in progress.

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[1] R. R. Brown, P. E. Green, Jr., B. Howland, R. M. Lerner, R. Manasse, and G. Pettingill, "Observations of the Russian earth satellite," PROC. IRE, vol. 45, pp. 1552–1553; November, 1957.
[2] A. M. Peterson, "Radio and radar tracking of the Russian earth satellite," PROC. IRE, vol. 45, pp. 1553–1555; November, 1957.

Detection of Sputniks I and II by CW Reflection*

The purpose of this note is to describe a simple experiment which indicates the feasibility of detecting artificial satellites by reflection of cw signals from a high-frequency radio station (in this case WWV). This experiment gives additional evidence of the extensive ion concentrations produced by an artificial satellite at heights of 100 to 500 miles, since the large and sustained field strengths observed could not be produced by reflection from the satellite proper.

During observations of the radio transmissions from Sputnik I it was noted that, while the satellite was approaching, the 20-mc signal was very rough and unsteady. However, about one minute before the nearest approach of the satellite the signal suddenly became very smooth and steady and remained so during the remainder of the pass. This change was attributed to the presence of an ionized region which is pushed

* Received by the IRE, December 17, 1957.

ahead of the satellite.[1] Owing to the large mean free path, the gas molecules are bounced far ahead of the satellite and if not previously ionized may become ionized by the action of the satellite.

Since this ionized region affected the forward radio transmission from the satellite it seemed reasonable that it should be possible to detect this ionization by other means. Accordingly, after the radio transmitter of Sputnik II ceased to transmit, the method of Wylie and Castillo[2] was tried with success. Using this method the receiver is tuned to the 20-mc transmission of WWV, Washington, D. C., at a distance of about 330 miles from Columbus, Ohio. During the day ionospheric ionization is sufficient to reflect a strong signal from Washington to Columbus on this frequency, but at night the ionization decreases until the signal becomes extremely weak. Under these conditions any temporary localized increase in ionospheric ionization may result in a detectable burst in signal. Meteors produce brief bursts, the signal being reflected by the temporary ion trails. A meteor-induced burst lasts only for a few seconds. In the present experiment more sustained bursts were also detected, often lasting for several minutes. These came at times corresponding closely with the predicted times for Sputnik I (spherical part) and Sputnik II.

A sample record taken on the morning of December 12, 1957, is shown in Fig. 1. Over most of the interval the signal level is low except for brief bursts produced by meteors.

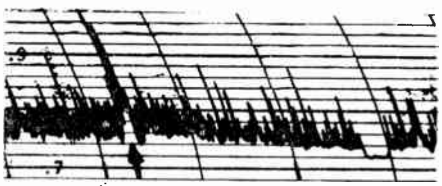

Fig. 1—Record showing burst of 20-mc WWV signal (indicated by arrow) caused by reflection from the ionized region associated with Sputnik I.

The WWV transmitter is off for several minutes at 0545 which establishes the background noise level. At 0625 there is a large burst lasting nearly 3 minutes. The time predicted by the Smithsonian Astrophysical Observatory, Cambridge, Mass., for the latitude crossing of Sputnik I was 0627. The satellite was at a height of about 150 miles on this pass.

Both Sputniks I and II have been detected regularly in this manner, it often being possible to record several passes of each on one night. The time of observed burst depends on the geometry of the satellite path with respect to the ground stations but was usually found to lie between the true latitude and meridian crossing times.

Many of the passes stand out more clearly than the sample in Fig. 1. For example, on the morning of December 14, 1957, between 0300 and 0800 A.M. there are two large bursts both being at least twice the height of the next highest peaks on the record (including all meteor spikes). The first burst occurred at 0442 and corresponded to Sputnik I which was predicted to make its meridian crossing at 0445. The second burst occurred at 0709 and corresponded to Sputnik II. The predicted latitude crossing was 0715. However, Sputnik II was observed visually on the preceding pass to make its latitude crossing at 0527 which was 7 minutes ahead of the predicted time. Applying this correction would make the true latitude crossing at 0708 or 1 minute ahead of the center of the observed burst.

At the time of the visually observed pass at 0527 there is a definite peak on the record but it is not sufficiently large to be convincing evidence that it was produced by Sputnik II. The satellite was at a height of about 600 miles on both passes (0527 and 0709) but perhaps due to differing gas densities at this altitude the burst on the first pass was small, if present at all.

It is apparent from Fig. 1 and the other observations that the satellite-induced ionization is much more effective than meteor ion trails for producing practical night-time transmission via ionospheric reflection on 20 mc. Looking ahead to the time when the upper regions of the atmosphere may be filled with artificial satellites, the ionization they produce may be intense and sustained enough to permit greatly increased night-time use of very high frequencies for long-distance communication.

JOHN D. KRAUS

[1] J. D. Kraus and J. S. Albus, "A note on some signal characteristics of Sputnik I," this issue, p. 610.
[2] L. R. Wylie and H. T. Castillo, "Clustering of meteors as detected by the use of radio technique," *Ohio J. Science*, vol. 56, pp. 339–347; November, 1956.

The Last Days of Sputnik I*

In another communication a simple experiment was described which demonstrated that it was possible to detect artificial earth satellites by means of cw reflection.[1] In this experiment 20-mc signals from WWV, Washington, D. C., were recorded at the Ohio State University Radio Observatory, Columbus, Ohio, at a distance of 330 miles from Washington. At night on this frequency WWV is usually extremely weak but may have occasional brief bursts only a few seconds long when meteors traverse the upper atmosphere somewhere between Washington and Columbus.[2] More sustained bursts lasting a minute or more were also observed at times coinciding with the passages of Sputniks I and II. This method is valuable since it permits the detection of a satellite after its radio transmitter has ceased to function.

An example of a satellite burst is shown in Fig. 1. This burst, recorded at approximately 0445 A.M. on December 14, 1957, occurred during the passage of Sputnik I. It is of particular interest that about 40 seconds after the start of the burst the signal cut off abruptly owing to the fact that the WWV transmitter was turned off. This provides evidence that the signal in the burst was from WWV. Magnetic sound tapes are also employed for the purpose of identification. For four minutes every hour (from the 45th to the 49th minute) WWV has a silent period. Although some bursts might be missed during this period, this silence is useful for establishing the background level and for indicating that the signal bursts on the record are from WWV.

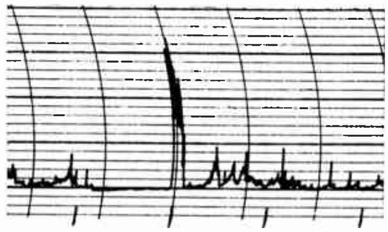

Fig. 1—Record of burst of WWV signal reflected from Sputnik I and associated ionization region at 4:45 A.M. EST on December 14, 1957. The burst ended abruptly when WWV's transmitter was turned off for its hourly 4-minute silent period, a few seconds after 4:45 A.M.

Since a satellite burst might be obtained for a passage of the satellite anywhere over the northeastern United States, it might appear that the times of passage would not be accurate enough to serve a useful purpose. However, it turns out that this is not the case. For example, the times of bursts agree with the expected latitude crossing times of Sputniks I and II to within about one minute (average value) or within two minutes in all but a few per cent of the cases. As explained in the following paragraphs, an average accuracy of one minute was entirely adequate to follow the demise of Sputnik I with a completeness of detail that is apparently not possible with any other method, radio or visual, applied to date.

The story of the last days of Sputnik I is presented by Fig. 2. Date is plotted as ordinate and time of day as abscissa. The small circles indicate the time of occurrence of large sustained bursts on a given date. The area of the circle is approximately proportional to the area under the burst deflection, this area being a convenient criterion for the burst intensity. Only bursts of about 15-second duration or longer are shown on the chart. The circles indicate every large sustained burst observed during the time covered by the graph, that is, between 0330 and 0610 from December 28 to January 8. The inserts in Fig. 2 show two actual records covering portions of the chart. The upper record, taken on December 30, 1957, has three large bursts which appear on the chart as the three circles opposite December 30. The time scale of the record is 15 minutes per 4 divisions (12 inches per hour on the actual record). The lower record, taken January 4, has three large bursts and two smaller ones, the corresponding circles on the chart being indicated by the arrows.

* Received by the IRE, January 13, 1958.
[1] J. D. Kraus, "Detection of Sputnik I and II by cw reflection," this issue, p. 611.
[2] L. R. Wylie and H. T. Castillo, "Clustering of meteors as detected by the use of radio technique," *Ohio J. Science*, vol. 56, pp. 339–347; November, 1956.

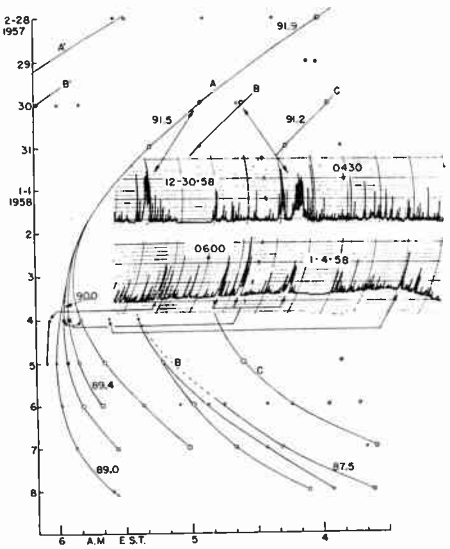

Fig. 2—Diagram showing the final break-up of Sputnik I during its last days aloft as obtained from radio reflection records, at the Ohio State University Radio Observatory. The circles indicate the time of occurrence (horizontal scale) of sustained bursts of WWV signal due to the passage of satellite objects as a function of date (vertical scale). Two samples of actual records are shown by the inserts with circles on the chart corresponding to large sustained bursts indicated by arrows. Many of the brief or spike-like bursts are presumably due to meteors.

Note the 4-minute period on each record during which WWV was off the air.

Referring to the curve A at December 30, the passage of a satellite occurred at 0452. On December 31 one occurred about 24 minutes later or at 0516. The time between the two passes equals 1464 minutes or the number of minutes in one solar day (1440) plus 24 minutes. Assuming that both passes were made by the same object, it must have made an integral number of revolutions (in this case 16) during the 1464-minute interval. Hence, its average period, or time for one revolution, is 91.5 minutes ($=1464/16$). Extending the curve to circles corresponding to passages on following days, it is possible to keep track of the progress of a satellite in the absence of any other information. Of course, to identify the curve with a particular earth satellite, such as Sputnik I, it is necessary to know at some point of the study the approximate period and orbital position of the satellite. On dates in December, earlier than those shown in Fig. 2, curve A (and curve A' for the next following passage) were identified with Sputnik I.

As a satellite falls closer to the earth, its period decreases. This decrease in period appears on the chart of Fig. 2 as a change in the slope of the curve. In the case of Fig. 2 a positive slope corresponds to a period of more than 90 minutes, a negative slope to a period of less than 90 minutes, and a vertical slope to a period of exactly 90 minutes. (Note that $16\times 90=1440$ minutes).

By extending curve A from day to day the progress of Sputnik I was followed in this manner. Connecting the other burst circles on the chart yielded two other curves B and C which follow a trend very similar to that of curve A suggesting that they also correspond to Sputnik I or parts thereof. Whether the three curves all correspond to parts of the 184-pound, 23-inch diameter, antenna-equipped, spherical Sputnik I it is not possible to determine from the radio data. If they do, the curves suggest, by extrapolation to earlier dates, that the sphere with whip antennas may have broken into at least three parts as early as December 22. A collision with a meteor is a possible cause of the break-up, the Ursid meteor shower occurring on this date. However, one of the curves (B or C) might correspond to the nose cone of the last stage rocket which put the sphere in its orbit. The last stage of the rocket itself was apparently down early in December.

The 3-digit figures adjacent to the curves in Fig. 2 indicate the period in minutes at that point of the curve. Curve B' (upper left) corresponds to the next later pass after curve B. The absence of data on the chart for the first three days of January resulted from the fact that the WWV observations were interrupted on these dates in order to determine the feasibility of detecting the satellites using a local cw transmitter on 21 mc. Although airplane reflections had been anticipated, they proved to be much too strong and numerous to apply the method successfully. Accordingly, on January 4 the receiver was retuned to WWV on 20 mc.

On this same date three large bursts were noted at the anticipated position of curve A whereas only one burst had been expected (see middle burst group of lower record in Fig. 2). On January 5 and 6 the three bursts again appeared but showed a significant tendency to spread out in time. The early burst moved farther ahead while the late burst dropped farther behind. This is precisely what would be expected to happen to three satellite bodies with slightly different initial velocities or densities, the one ahead getting farther ahead each day and the last one getting farther behind. Thus, on January 6 the first and last objects were 27 minutes apart or separated about 8000 miles along the satellite's orbit. On January 7 the middle object had disappeared and on January 8 there was no sign of the other two on the record. On January 5 a fourth object was noted trailing the other three. This object was farther behind on succeeding days and still observable on January 8. The longer period of this object could mean that it was in a higher orbit than the others and hence should come down last. Extrapolating the group A curves back to dates prior to January 4, it appears that the original A object began to break into fragments on January 2 or 3.

Referring to curves B and C these follow a similar trend in slope as curve A but show a more rapid decrease in period as would be expected for objects running further ahead in the orbit. Also between January 3 and January 5, curve B branches into three curves suggesting a break-up of the curve-B object into three distinct fragments. On January 6 there appeared to be a total of at least eight distinct fragments of Sputnik I still orbiting. By January 8 all but one of the A group had disappeared, presumably indicating that they had fallen to the earth or completely disintegrated. The fact that the fragments in the B group were still in the orbit on January 8, in spite of their shorter periods, suggests that they were of substantially higher density than the fragments of the A group, or made of a more heat-resistant material, such as would be present in the nose cone.

The self-consistency of the curves in Fig. 2 makes it appear very unlikely that the bursts observed were due to any other cause than the break-up of Sputnik I as suggested above. In fact, the probability that the pattern of the circles in the A group alone could result from a random distribution instead of such a break-up is less than one in a million.

An important supplementary observation on the demise of Sputnik I was made by three fliers in an airplane and the control tower operator at Port Columbus, Ohio, at 0555 on January 4. At that time the pilot, Marine Capt. Donald Parrill, reported a bright, self-luminous, orange-red object, similar to a fireball, with a short thin tail about 5 degrees long crossing the NE sky in horizontal flight at a maximum elevation angle of 15–25°. It was visible for some 15 seconds and travelled from NW to SE. This course corresponded to that for Sputnik I and the time was within a few seconds of the maximum of the center burst of the three-burst group in the lower record of Fig. 2. The circle in the chart corresponding to this burst is marked with an X. It appears very likely that the object reported corresponded to a group-A fragment. The fact that the center burst of the 3-burst group was observed on January 5 and January 6, or two days after the sighting of the orange-red object, may be explained in two ways. One possibility is that the object seen was a part of the A group not previously detected by the radio reflection method which passed near Columbus at the same time as the middle fragment of the three-object group but at a lower altitude where the denser air produced a visible heating effect. Another possibility is that the object seen was one with sharp metal points or edges which were melting and burning off with visible effect during the satellite's dip into the denser atmosphere near perigee. The main body of the object, however, after this self-streamlining process continued to orbit for two more days before coming down or disintegrating completely.

In conclusion, it appears reasonable to deduce that the end of Sputnik I was a gradual rather than an abrupt affair, starting with an initial break-up in December followed by a more complete fragmentation as the pieces descended lower on the first days of January. However, the radio observations give no certain indication of any fragments down or completely disintegrated until January 7 with more down on January 8. On January 9 the record between 0000 and 0700 was dramatically free of satellite bursts, the only large burst occurring at 0308 corresponding to the slowest object in the B group. Its average period during the preceding 24 hours was about 86.9 minutes, making it unlikely that this last remaining significant fragment of Sputnik I would remain up much longer. Thus, the time between the launching of the earth's first artificial satellite on October 4, 1957, and the descent of

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all but one of its detectable fragments is 97 days.

These observations were assisted by a grant from the National Science Foundation and support from the Fund for Basic Research of The Ohio State University.

It is a pleasure to acknowledge the interest and helpful discussions by Prof. E. E. Dreese, Chairman of the Department of Electrical Engineering, and Prof. M. L. Pool, Department of Physics, the Ohio State University, and also the valuable assistance given on certain of the observations by Prof. Robert C. Higgy, Director of Ohio State University's Radio Station WOSU.

JOHN D. KRAUS

Noise Output of Balanced Frequency Discriminator*

Few calculations involving FM and noise are either simple or straightforward. I was somewhat surprised, then, to find the calculation of the rms output, σ^2, of a balanced frequency discriminator to a Gaussian noise input both simple (relatively) and straightforward.

Since the result obtained is not without some interest, I herewith present a brief outline of the analysis.

A common form of discriminator is shown in Fig. 1. The input is passed through

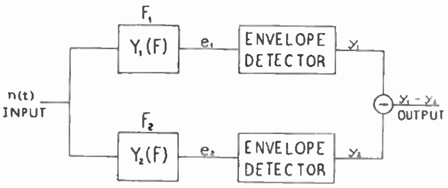

Fig. 1—Frequency discriminator.

two filters, F_1 and F_2, and the outputs are envelope detected and subtracted. Filter F_k has transfer function $Y_k(f)=R_k(f)+iI_k(f)=A_k(f)\,e^{i\phi_k(f)}$, $k=1, 2$. In practice, one chooses the filters so that $A_1(f)-A_2(f)$ is approximately linear in f in the region of the fm carrier, a condition which is independent of the phase characteristics ϕ_1 and ϕ_2.

The analysis below shows that for fixed A_1 and A_2, the rms noise output, σ^2, is minimized if $\phi_1=\phi_2$. In detail,

$$\sigma^2 = \left(2-\frac{\pi}{2}\right)(\alpha_1+\alpha_2) - \pi\sqrt{\alpha_1\alpha_2} \cdot\left[\frac{2}{\pi}(1+h)\mathcal{E}\left(\frac{2\sqrt{h}}{1+h}\right)-1\right] \qquad (1)$$

where

$$\alpha_i = \int_0^\infty |Y_i(f)|^2 w(f)df, \quad i = 1, 2,$$

$$h^2 = \frac{1}{\alpha_1\alpha_2}\left|\int_0^\infty Y_1(f)\bar{Y}_2(f)w(f)df\right|^2. \qquad (2)$$

$$\mathcal{E}(x) = \int_0^{\pi/2}\sqrt{1-x^2\sin^2\theta}\,d\theta$$

is the complete elliptic integral of the second kind, and $w(f)$ is the power spectrum of the input Gaussian noise. The expression in brackets in (1) is plotted in Fig. 2 and is seen to be

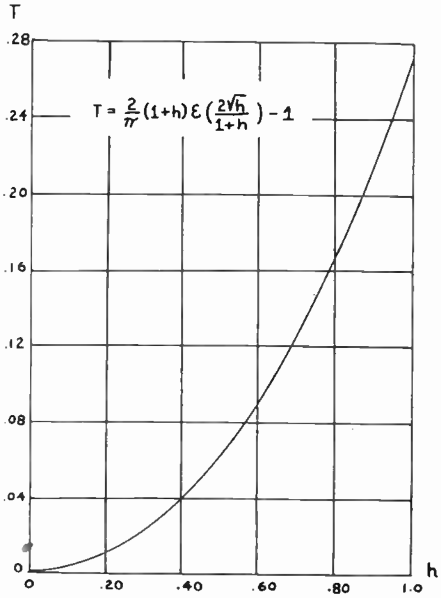

Fig. 2.

monotone in h. For fixed A_1 and A_2, a condition which implies fixed α_1 and α_2, (1) is minimized when h is a maximum. From (2) this occurs for $\phi_1=\phi_2$.

In minimizing the rms noise output by adjusting $\phi_1=\phi_2$, one must keep A_1-A_2 approximately linear in f in the neighborhood of the fm carrier in order to maintain good signal output from the discriminator. With many of the discriminator designs now in use, this condition can be fulfilled only by inserting an additional phase shifting network in either or both arms of the discriminator.

An outline of the derivation of (1) follows. The input noise is represented by

$$n(t) = \sum_{k=1}^{\infty}(a_k\cos 2\pi f_k t + b_k \sin 2\pi f_k t)$$

where $f_k=k/2T$ and where all the a's and b's are independent normal variates with mean zero and variance $E(a_k^2)=E(b_k^2)=w(f_k)/2T$. As $T\to\infty$, $n(t)$ represent a Gaussian noise with power density spectrum $w(f)$. The responses of F_1 and F_2 to this input can be readily computed and put in the form

$$e_i = \xi_i\cos 2\pi f_0 t + \eta_i \sin 2\pi f_0 t, \quad i = 1, 2. \qquad (3)$$

Here

$$\xi_i = \sum\left[\{a_kR_i(f_k)+b_kI_i(f_k)\}\cos 2\pi(f_k-f_0)t + \{b_kR_i(f_k)-a_kI_i(f_k)\}\sin 2\pi(f_k-f_0)t\right], \quad i = 1, 2.$$

The η_i can be obtained from the ξ_i by replacing a_k by b_k and by replacing b_k by $-a_k$.

The ξ's and η's are jointly Gaussian, each with mean zero. Their covariances are readily computed and in the limit $T\to\infty$ become

$$E\xi_i^2 = E\eta_i^2 = \alpha_i, \qquad i = 1, 2,$$

$$E\xi_1\xi_2 = E\eta_1\eta_2 = \mathrm{Re}\int_0^\infty Y_1(f)\bar{Y}_2(f)w(f)df,$$

$$E\xi_1\eta_2 = -E\xi_2\eta_1 = \mathrm{Im}\int_0^\infty Y_1(f)\bar{Y}_2(f)w(f)df.$$

The joint density $p(\xi_1, \xi_2, \eta_1, \eta_2)$ can then be written down in terms of these quantities.

If F_1 and F_2 are rather sharply tuned to a frequency near f_0, then the ξ's and η's will vary slowly in times of order of magnitude f_0^{-1} and the envelopes of the filter outputs are given from (3) by $y_i=(\xi_i^2+\eta_i^2)^{1/2}$, $i=1, 2$. The discriminator output, y_1-y_2 has rms value $\sigma^2=E(y_1-y_2)^2-[E(y_1-y_2)]^2$. In evaluating this expression, the only integral of any difficulty is

$$Ey_1y_2 = \int\int\int\int p(\xi_1, \xi_2, \eta_1, \eta_2) \cdot\prod_1^2(\xi_i^2+\eta_i^2)^{1/2}d\xi_i d\eta_i.$$

Let $\xi_i=\sqrt{\alpha_i(1-h^2)}\,r_i\cos\theta_i$, $\eta_i=\sqrt{\alpha_i(1-h^2)}\,r_i\sin\theta_i$, $i=1, 2$. The θ integration can then be performed to yield a Bessel function $J_0(ihr_1r_2)$. If this factor is expanded in a series, and term by term integration performed, Ey_1y_2 is obtained as a power series in h^2. Formula 6.7111 of the *Smithsonian Mathematical Formulae* (Smithsonian Institution, Washington, 1947) shows the series to be

$$Ey_1y_2 = \sqrt{\alpha_1\alpha_2}(1+h)\mathcal{E}\left(\frac{2\sqrt{h}}{1+h}\right).$$

It is to be noted carefully that the results obtained are for Gaussian noise input to the discriminator. No limiter action has been assumed.

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* Received by the IRE, September 24, 1957.

An Improved Operational Amplifier*

As pointed out by Staudhammer,[1] the equations are in error. An interesting sidelight to these errors is that of the five comments I have received, only two were aware of the error.

The two letters by Bossinger[2] and Staudhammer commenting on my letter[3] assume

* Received by the IRE, October 22, 1957.

[1] J. Staudhammer, "An improved operational amplifier," PROC. IRE, vol. 45, p. 1415; October, 1957.
[2] A. Bossinger, "An improved operational amplifier," PROC. IRE, vol. 45, p. 1415; October, 1957.
[3] R. Nitzberg, "An improved operation amplifier," PROC. IRE, vol. 45, p. 880; June, 1957.

that I wish to have all computers redesigned incorporating my recommendation. All I really desired was to demonstrate a technique for building a cheap, fairly accurate operational amplifier. A particular desirable application of this circuit is a Miller sweep generator.[4] In this case, the tolerances on the regenerative feedback loop are much less severe.

The Miller sweep circuit of Fig. 1 is similar to the traditional circuit. A cascode amplifier has been used to facilitate the connection of the positive feedback loop. The similarity between this circuit and an integrating operational amplifier is apparent.

If the feedback resistor is infinite, the circuit will generate a negative going sawtooth at the cathode follower output and the nonlinearity caused by the finite amplifier gain may easily be calculated. As the feedback resistor is decreased from its infinite value, the output waveform will improve in linearity. When the feedback resistor value is approximately equal to, but less than, the value of the cascode plate resistor, the output waveform will be truly linear. Further decrease of the feedback resistor value will cause the output waveform to become nonlinear again. Thus, any value of feedback resistor between infinity and the previously mentioned value will decrease the linearity error.

Bossinger mentioned that for algebraic addition, it is possible to correct the computational error by adjustment of the resistive components in the beta network. If the components of integration and differentiation circuits are intentionally not pure resistances and capacitors, their computational error may also be corrected.

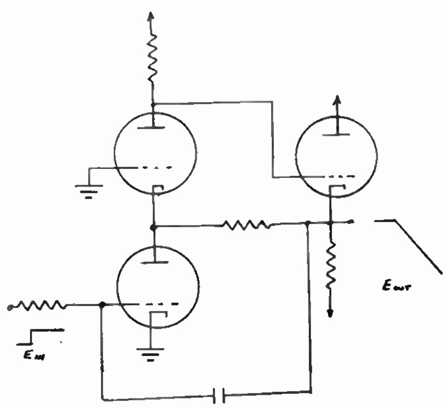

Fig. 1.

For the circuit of Fig. 1[3]

$$G = -\frac{Z_2}{Z_1}\,\frac{1}{1+\dfrac{1+Z_2/Z_1}{K}} \tag{1}$$

and it is desired that

$$G = \pm jA \tag{2}$$

where A is a real number and the sign is determined by whether differentiation or integration is to be performed. If

$$\frac{Z_2}{Z_1} = \mp jkAe^{j\phi}$$

$$G = \pm jA\,\frac{Kke^{j\phi}}{K+1\pm kA\sin\phi\mp jkA\cos\phi}$$

to meet the requirements of (2) above

$$\frac{Kke^{j\phi}}{C_1e^{j\theta}} = 1 \tag{3}$$

$$\frac{Z_2}{Z_1} = \mp j\,\frac{C_1}{K}\,Ae^{j\theta} \tag{4}$$

$$C_1 = \left[(K \pm kA\sin\phi + 1) + (kA\cos\phi)^2\right]^{1/2} \tag{5}$$

$$\theta = \mp \tan^{-1}\frac{kA\cos\phi}{K\pm A\sin\phi+1}. \tag{6}$$

To facilitate the use of (5) and (6), it is desirable to assume that K is large, ϕ is small, and k is approximately unity. Under these assumptions,

$$\frac{Z_2}{Z_1} = \pm\left[1+\left(\frac{A}{K}\right)^2\right]^{1/2} Ae^{\mp j(A/K)}.$$

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[4] R. Nitzberg, "Linear sweep voltage generators and their application to range measuring," Syracuse Univ., Master's thesis; August, 1953.

The Effect of Beam Position on Deflection in Slit Lenses*

The use of a hollow or sheet electron beam in vacuum tubes often requires the beam to pass through an annular or straight slit in the electron gun. These slits have a lens action, somewhat stronger than that of the usual circular aperture, which together with their double-sided character poses a serious focusing problem. The symmetry of beam deflection in these lenses is extremely sensitive to placement of the beam in the lens.

Consider an annular aperture lens of inner radius L_1, outer radius L_2, with a hollow beam passing through it. The beam has inner and outer radii R_1 and R_2 respectively. Glewwe[1] has shown that the deflection angle ϕ in such a lens may be given by $\tan\phi = M \ln (R/\sqrt{L_1L_2})$ where M is a factor including the fields, voltages, and geometrical terms. If we define

$$x = R/\sqrt{L_1L_2}$$

we may write:

$$\tan\phi_a = M\ln x_a \text{ and } \tan\phi_b = M\ln x_b.$$

Let us set $x_ax_b = 1$, or $R_aR_b = L_1L_2$, then $\tan\phi_a = -\tan\phi_b$, or the beam edges are deflected symmetrically. Now suppose the beam thickness is preserved but both edges are shifted by an amount ΔR increasing both x_a and x_b by:

$$\Delta x = \frac{\Delta R}{\sqrt{L_1L_2}}.$$

After making the approximation

$$\ln (x+\Delta x) \approx \ln x + \Delta x/x,$$

we find that

$$-\frac{\tan\phi_a}{\tan\phi_b} = \left(\frac{1-A\Delta x}{1+B\Delta x}\right)$$

where

$$A = x_b/\ln x_b$$
$$B = 1/(x_b \ln x_b).$$

Fig. 1 shows this ratio (or its reciprocal for negative Δx) as a function of Δx with x_b as a parameter reflecting beam thickness. It is evident that a very slight displacement of a beam can give rise to a large asymmetry in the deflection which will show up as a net inward or outward deflection of the entire beam. To avoid this net deflection, which is particularly serious for thin beams, requires extreme precision in design and construction of the electron gun.

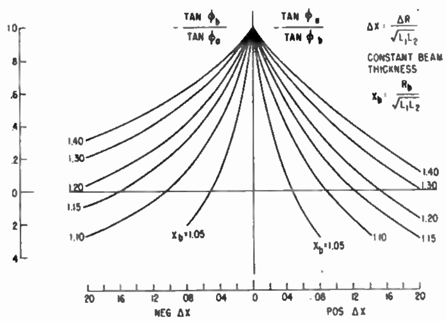

Fig. 1.

As an example of the precision needed, consider a gun in which $\sqrt{L_1L_2} = 0.684$ inch and $R_b = 0.790$ inch giving $x_b = 1.158$. The observed ϕ_a and ϕ_b were $-24.9°$ and $16.4°$; respectively. Application of the above theory shows that to account for this assymmetry an inward shift of only 0.024 inch or 3.5 per cent of $\sqrt{L_1L_2}$ is required.

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* Received by the IRE, October 28, 1957.

[1] C. W. Glewwe, "Some properties of an annular electron lens," M.S. thesis, University of Minnesota, Minneapolis, Minn.; October, 1955.

An Extended General Network Theorem on Rectification*

Reference is made to Gewartowski's[1] timely and highly useful theorem about the need for nonlinear resistance for rectified ac output from one or more ac sources.

* Received by the IRE, October 29, 1957.

[1] J. W. Gewartowski, "A general circuit theorem on rectifications," Proc. IRE, vol. 45, p. 1410; October, 1957.

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The theorem states that it is impossible to obtain a rectified dc power from an ac source or sources unless nonlinear resistance is present in the circuit. This is not altogether true, since this formulation leaves out the important category of networks referred to as *time variant linear resistances*. They are just as linear as the time-independent versions, meaning that a change in voltage causes a proportional change in current, and vice versa, although it is willingly admitted that the effect of the periodic variation is nonlinearity. This, of course, is the secret of the success of the "multiplicative" transistor or tube mixer, used for frequency changing.

The original Fig. 2 is here redrawn as Fig. 1(a) with a loop free from nonlinear

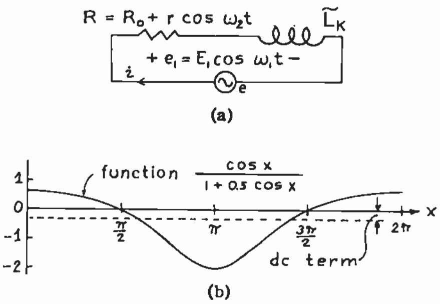

Fig. 1.

capacitance, however, with nonlinear inductance and one generator, and with a periodically varied *linear* resistance R. The variation is caused by one of the other generators allowed in the system. As has already been proven by Gewartowski, $\tilde{L}_K$ and $\tilde{C}_K$ may be disregarded as contributors of a rectified dc component, so that we may write

$$i = \frac{e_1}{R} = \frac{E_1 \cos \omega_1 t}{R_0 + r \cos \omega_2 t} = I_1 \frac{\cos \omega_1 t}{1 + a \cos \omega_2 t}, \quad (1)$$

or, after the application of the binomial theorem,

$$i = I_1 \cos \omega_1 t (1 - a \cos \omega_2 t + a^2 \cos^2 \omega_2 t - \cdots). \quad (2)$$

Then, for synchronous operation of the system, *i.e.*, $\omega_2 = \omega_1$,

$$i = \text{dc term} + I_1 \cos \omega_1 t - 0.5 a I_1 \cos 2\omega_1 t + \cdots, \quad (3)$$

so, evidently, we have produced a dc term without the aid of a resistance possessing a nonlinear characteristic. This is the theory of the *synchronous detector* in a nutshell! Actually, if we add enough harmonics, in accordance with Fourier, the time variant resistance becomes a square-wave switch; a perfectly logical analytic derivation!

That (1) produces a dc term is proven by the graph in Fig. 1(b), which represents the *linear* and stable resistance function $(1+0.5 \cos x)$.

Further, this writer feels that a rectification theorem should state explicitly that the resistance, whether nonlinear or linear-periodic, can be negative, *i.e.*, a stimulance,[2] as long as the system remains stable in accordance with common stability criteria. For $R_K < 0$ for the original network in simplest possible interpretation, oscillations would take place with a pair-pole in the right-hand side of the s plane, and unless we integrate properly for the average, a fictitious dc component due to increasing amplitude may show up. Proper integration would mean $T = \infty$, an academic case, since physical systems are limited by nonlinearity; linear stimulance in the end producing nonlinear resistance. This, by the way, is the theme of the common feedback oscillator!

Rather than saying that the criterion on rectification is nonlinearity, why do we not say that the criterion on rectification is whether or not the introduced network element is a quadratic-term generator! Inside this classification, then, there are two groups: the inherently nonlinear resistance ("additive" mixing), which produces the quadratic term by an inherent embodiment of its current-voltage characteristic, and the linear-periodic resistance, producing the quadratic term when the periodicity is that of the properly applied and phased signal. Nonsynchronous operation of the latter yields modulators and frequency changers. In view of the above discussion, the following suggestion for the formulation of the theorem is forwarded.

Theorem: It is impossible to obtain a rectified dc power output from one or more periodic sources in a stable system unless a nonstorage nonlinear network element, or linear-periodic network element, synchronous with one applied source, is present; being either resistance or stimulance.

This writer suggests that the term "drop" in "voltage drop" be abolished in network theory. The use with reference to $\tilde{L}_K$ and $\tilde{C}_K$ provides additional examples of "drop" (inherently associated with direction) being used when apparently the idea is not to convey direction. "Drop" does not merely introduce duplicate information; much worse, it frequently introduces contradictory information!

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[2] See H. Stockman, "On reciprocal inductance," PROC. IRE, vol. 43, p. 341; March, 1955. Also, "Time-Saving Network Calculations," SER Co., Waltham, Mass.

Variation of Junction Transistor Current Amplification Factor with Emitter Current*

Analyses which have taken into account the influence of the electric field as well as diffusion in the base region of junction transistors have been given by Webster[1] and Rittner.[2]

More recently, Fletcher[3] has attempted to bring the two theories into concord by pointing out errors in the former which arose as a result of Webster's neglecting to replace D_p by his own "effective" diffusion constant

$$D_p\left(\frac{1 + 2p_E/N_d}{1 + p_E/N_d}\right)$$

when modifying the emitter efficiency and volume recombination components of $1/\alpha_{CB}$ carried over from previous theories.[4] It is the purpose of the present note to show that:

1) On the basis of the Webster assumptions (*i.e.*, space-charge neutrality, $I_n \ll I_p$ in base, etc.), it is permissible to retain Fletcher's fall-off factor associated with the emitter efficiency component, but that a discrepancy in his volume recombination term still remains. Thus a further revision in this factor is required other than that necessitated by considering the correct variation of lifetime τ_p with injected level,
2) In consequence of the variation of mobility of carriers with impurity concentration,[5] the expression $\sigma_B W/\sigma_E L_{nE}$, commonly used for the low-level value of I_{En}/I_{Ep}, can be expected to be in error by an order of magnitude; this is in a direction to resolve the disparity which Webster found[1] between the value of $\sigma_E L_{nE}$ chosen to yield best fit and the value of $\sigma_E L_{nE}$ expected from other considerations.

THE VOLUME RECOMBINATION TERM

The steady-state one-dimensional continuity equation to be solved in the base region is[6]

$$\frac{1}{q} \cdot \frac{dI_p}{dx} + \frac{p - p_n}{\tau_p} = 0. \quad (1)$$

When (1) was solved on the basis of diffusion only,[4] with $p_c = 0$ and $W \ll L_p$, the expression

$$\frac{I_{VR}}{I_{Ep}} \doteq \frac{1}{2} \frac{W^2}{L_p^2}$$

was derived. Webster's (16) and Fletcher's (9) are simply obtained by modulating this answer. However, this treatment overlooks the fact that the variations in D_p and τ_p are more correctly taken into account in the original (1) before solving it. Thus, when Webster's (12) for I_p is substituted, (1) becomes

$$\left(\frac{N_d + 2p}{N_d + p}\right) \frac{d^2 p}{dx^2} + \frac{N_d}{(N_d + p)^2} \left(\frac{dp}{dx}\right)^2 - \frac{p - p_n}{L_p^2} = 0. \quad (2)$$

* Received by the IRE, October 13, 1957.

[1] W. M. Webster, "On the variation of junction transistor current amplification factor with emitter current," PROC. IRE, vol. 42, pp. 914–920; June, 1954.

[2] E. S. Rittner, "Extension of the theory of the junction transistor," *Phys. Rev.*, vol. 94, pp. 1161–1171; June, 1954.

[3] N. H. Fletcher, "The variation of junction transistor current amplification factor with emitter current," PROC. IRE, vol. 44, pp. 1475–1476; October, 1956.

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

[5] M. B. Prince, "Drift mobilities in semiconductors," *Phys. Rev.*, vol. 92, pp. 681–687; November, 1953.

[6] W. Shockley, "Electrons and Holes in Semiconductors," D. Van Nostrand, Inc., New York, N. Y.; p. 313; 1950.

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A first integral of this nonlinear equation can be obtained without assuming a specific law of variation of L_p with injected level, by letting $y=dp/dx$ and considering p as the independent variable. Eq. (2) is then of standard Bernoulli type and can be integrated immediately to give

$$\left(\frac{N_d+2p}{N_d+p}\right)^2\cdot\left(\frac{dp}{dx}\right)^2$$
$$=2\int\frac{p-p_n}{L_p^2}\left(\frac{N_d+2p}{N_d+p}\right)dp+\text{const.}\quad(3)$$

It is seen that the left-hand side is proportional to the current density squared. Eq. (3) is therefore equivalent to

$$I_{Ep}^2-I_{cp}^2=2(qD_p)^2$$
$$\cdot\int_{p_c}^{p_E}\frac{p-p_n}{L_p^2}\left(\frac{N_d+2p}{N_d+p}\right)dp.\quad(4)$$

Again, the left-hand side can be factorized to 2 $I_{VR}\cdot I_{Ep}$, provided $I_{VR}\ll I_{Ep}$. Introducing Webster's

$$Z=\frac{W}{N_d}\cdot\frac{I_{Ep}}{qD_p},$$

(4) can then be written, for $p\gg p_n$,

$$\frac{I_{VR}}{I_{Ep}}=\frac{W^2}{Z^2}\cdot\int_{P_c}^{P_E}\frac{P}{L_p^2}\left(\frac{1+2P}{1+P}\right)dP\quad(5)$$

where $P=p/N_d$.

If it is also assumed that $P_c=0$, then the required volume recombination term, obtained by differentiation of (5), is

$$\frac{\partial I_{VR}}{\partial I_{Ep}}=\frac{W^2}{Z}\left[\frac{P_E}{L_p^2(P_E)}\right.$$
$$\left.-\frac{1}{Z}\int_0^{P_E}\frac{P}{L_p^2}\left(\frac{1+2P}{1+P}\right)dP\right]\quad(6)$$

and[1] $Z=2P_E-\ln(1+P_E)$. There is thus a deviation, associated with I_{VR} rather than I_{Ep}, from the form given by Fletcher. *E.g.*, for bimolecular recombination expressed by

$$\frac{1}{L_p^2}=\frac{1}{L_{p0}^2}(1+P),$$

(6) yields

$$\frac{\partial I_{VR}}{\partial I_{Ep}}\doteq\frac{W^2}{2L_{p0}^2}\left(1+\frac{2}{3}Z\right)$$

at low levels, etc., which agrees with (52) of a solution by Hauri;[7] or for constant lifetime,

$$\frac{\partial I_{VR}}{\partial I_{Ep}}\doteq\frac{W^2}{4L^2}$$

at high levels, etc., agreeing with Rittner's (74).

A more general value of $\partial I_{VR}/\partial I_{Ep}$ is readily obtainable from (6) by use of the Shockley-Read formula[8] for variation of lifetime with injected level.

[7] E. R. Hauri, "Zur Frage der Abhängigkeit der Stromverstarkung von Flächentransistoren vom Emitterstrom," *Tech. Mitt. PTT.*, no. 11, pp. 441–451; November, 1956.

[8] W. Shockley and W. T. Read, "Statistics of recombination of holes and electrons," *Phys. Rev.*, vol. 87, pp. 835–842; September, 1952.

THE EMITTER EFFICIENCY TERM

Since conductivity modulation is usually negligible in the emitter, conventional diffusion theory for minority carrier flow is still applicable there. Thus, Fletcher's (5) contains the correct fall-off factor.

Nevertheless, there appears to be a discrepancy in the value of γ usually taken for very low-level injection. The relevant expression derived from the diffusion equations is

$$\frac{1}{\gamma}-1=\frac{I_{En}}{I_{Ep}}\propto\frac{D_n\cdot n_E}{D_p\cdot p_E}\quad(7)$$

where n_E and p_E are minority carrier densities on either side of the junction. Now

$$\frac{n_E}{p_E}\doteq\frac{n_p}{p_n}$$

(at low levels), and since, in thermal equilibrium, minority carrier density is inversely related to majority carrier density, this ratio, in turn, $\doteq(N_d)_B/(N_a)_E$. However, D_n and D_p in (7) are essentially *minority* carrier diffusion constants.

Thus, to equate (7) to σ_B/σ_E as is usually done, is in error by a factor $(\mu_n\mu_p)_E/(\mu_n\mu_p)_B$. *E.g.*, with typical impurity concentrations for germanium of $(N_d)_B=5\times10^{14}$ cm^{-3} and $(N_a)_E=5\times10^{17}$ cm^{-3}, say, Prince's results[6] give

$$\frac{(\mu_n\mu_p)_E}{(\mu_n\mu_p)_B}\doteq\frac{1}{17}.$$

This is in the right direction to resolve the order-of-magnitude disparity which Webster found between $\sigma_{B.W}.I_{Ep}/I_{En}$ obtained from best fit and the value of $\sigma_E L_{nE}$ expected by extrapolating measured diffusion lengths on relatively pure material to the emitter resistivity.

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Efficiency of Large Antennas in O/H Links*

Taking into account the anisotropy of atmospheric turbulence in the mechanism of tropospheric scattering propagation, one arrives at the conclusion that to obtain maximum efficiency from a large antenna, it should be dissymmetrical. The ratio between the two apertures depends on the dimension of the beamwidth and the distance. On the basis of these considerations we have realized and patented a new type of antenna which is undergoing tests. We also noticed that Booker and Gordon[1] arrive at a similar conclusion about stratospheric scattering propagation.

GIUSEPPE DE VITO
Co. El. Complementi Elettronici
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* Received by the IRE, October 24, 1957.

[1] H. G. Booker and W. E. Gordon, "The role of stratospheric scattering in radio communication," PROC. IRE, vol. 45, pp. 1223–1227; September, 1957.

High-Frequency Quartz Filter Crystals*

One of the requirements for high-frequency quartz filter crystals is freedom from disturbing frequencies, also referred to as unwanted or spurious modes, in a specified range in the vicinity of the desired frequency. The elimination of disturbing frequencies is an essential problem in the design and construction of thickness mode resonators.[1–3]

Single response thickness-shear modes, *e.g.*, of the AT type in the frequency range 0.5 to 5 mc, can be provided by using adequately bevelled circular plates of a diameter-thickness ratio less than approximately 30.[4,5] The process of bevelling provides boundary conditions which greatly reduce excitation of the disturbing modes, but it becomes ineffective for larger diameter-thickness ratios at frequencies higher than 10 mc.

For plates with the larger diameter-thickness ratio, the electrode size has a characteristic influence on the disturbing modes. The electrode area must be considerably smaller than the plate area in order to reduce the disturbing frequencies.[6]

Sufficient elimination of the disturbing modes of high-frequency resonators can be achieved for practical purposes by providing particular boundary conditions by means of special plate geometry. A triangular plate shape reduces disturbing frequencies to some extent according to recent investigations. The frequency spectrum for a typical 11.5-mc commercial crystal unit using circular AT quartz oscillators is shown in Fig. 1.

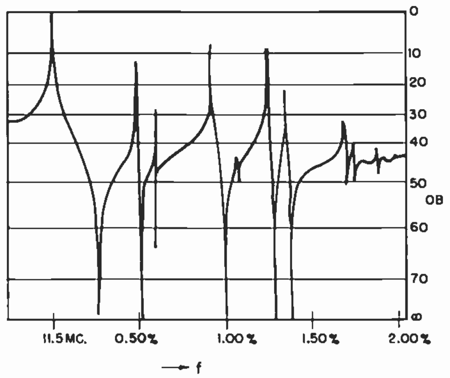

Fig. 1.

Fig. 2 shows the effect of reducing the size of the electrodes on a circular AT disk. Fig. 3 shows the spectrum of a 11.5-mc resonator using a triangular-shaped AT type plate combined with small electrodes. The

* Received by the IRE, November 4, 1957.

[1] R. Bechmann, "Quartz resonators," *Telefunken Ztg.*, vol. 18, pp. 5–15; July, 1937.

[2] R. Bechmann, "Properties of quartz oscillators and resonators in the range from 300 to 5000 kc/s," *Hochfreq. Elektroak.*, vol. 59, pp. 97–105; April, 1942.

[3] W. G. Cady, "Piezoelectricity," McGraw-Hill Book Co., Inc., New York, N. Y.; 1946.

[4] R. Bechmann, "Single response thickness-shear mode resonators using circular bevelled plates," *J. Sci. Instr.*, vol. 29, pp. 73–76; March, 1952.

[5] R. Bechmann, "Piezoelectric crystal," U.S. Patent 2,245,178; June 10, 1941.

[6] R. Bechmann, "Piezoelectric plate," U.S. Patent 2,249,933; July 22, 1941.

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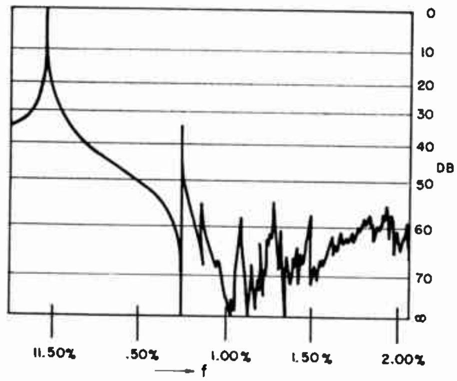

Fig. 2.

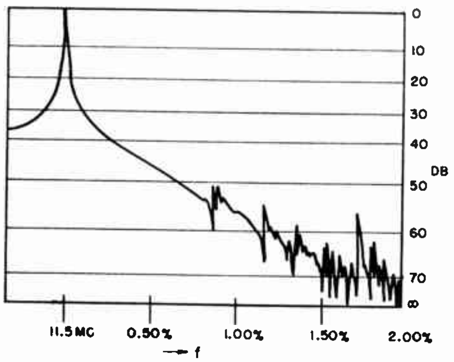

Fig. 3.

disturbing modes were reduced by an amount greater than 50 db below the main resonance. The resonance curves shown were recorded by an equipment which was first described by the author[1] but modified by use of a hybrid coil bridge circuit, a logarithmic amplifier, and an attenuator.

Besides the usual requirements for oscillators; *e.g.*, low temperature coefficient of frequencies and others, there is a requirement for certain values of the motional capacitance or inductance for filter crystals which can be fulfilled.

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Design of a Conical Taper in Circular Waveguide System Supporting H_{01} Mode*

In a circular waveguide system it is often required to connect two circular guides of different cross section. Because of the low attenuation constant the energy propagates in the H_{01} mode. Any other modes are undesirable. Nevertheless, the taper will produce higher-order modes. The aim of the present work is to determine the distribution of the higher-order modes if the taper is conical.

In the calculations the following assumptions are used.

1) The conductivity of the wall is infinite.
2) The dielectric inside the guide is lossless.
3) The taper is very gradual; *i.e.*, θ_0 is small (see Fig. 1).

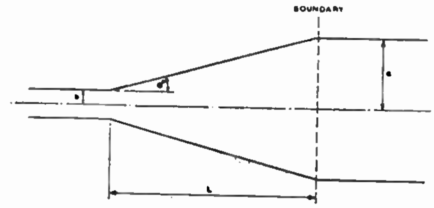

Fig. 1—The conical taper.

4) The higher-order modes in the taper are neglected; *i.e.*, in the taper a single transverse electric spherical, axial symmetrical wave will propagate.
5) In the larger guide the attenuation is small; *i.e.*, the wavelength is much smaller than the critical, $\lambda \ll \lambda_c$.
6) The reflections from the junctions are neglected.
7) In the larger guide only H_{0n} waves will propagate.

Our method of solution is to compute the field intensities in the taper and in the circular guide, respectively, and match them on their boundary.

Solving the Maxwell equations in the taper section subject to the above assumptions, after some mathematical approximations we obtain, for the circumferential component of the electric intensity on the boundary,

$$E_\phi = C \exp\left[-j\sigma t^2\right] J_1(k_{11}t), \qquad (1)$$

where

$\sigma = \pi(a/\lambda)(a-b)/L.$
a = the radius of the larger guide.
b = the radius of the smaller guide.
L = the length of the taper.
k_{11} = the first root of the first order Bessel function $J_1(x)$.
$t = \rho/a$.
ρ = the radial co-ordinate of a cylindrical co-ordinate system coaxial with the guide.
C = a constant factor.

The electric intensity of the H_{0n} mode in the circular guide is given by the well-known formula

$$E_\phi = \frac{h}{k_{1n}} J_1(k_{1n}t), \qquad (2)$$

where

k_{1n} = the nth root of the function $J_1(x)$.
h = a constant factor.

In both the taper and the guide the electric intensity is in the plane of their boundaries and has only circumferential component. Hence it is enough to use the identity of E_ϕ and E_ϕ on the boundary as follows

$$C \exp(-j\sigma t^2) J_1(k_{11}t) \equiv \sum_{n=1}^{\infty} \frac{h_n}{k_{1n}} J_1(k_{1n}t), \qquad (3)$$

where $|h_n|$ is the relative amplitude of the H_{0n} mode.

According to the theory of the Fourier-Bessel series h_n may be obtained from the following expression

$$h_n = \frac{2k_{1n}}{J_0^2(k_{1n})} \cdot \int_0^1 t \exp(-j\sigma t^2) J_1(k_{11}t) J_1(k_{1n}t) dt. \qquad (4)$$

The value of σ is zero, if $a=b$. In this case the amplitude of all the modes, except for the first, is zero because of the orthogonality of the above Bessel functions.

The amplitude of the H_{0n} mode is the function of σ only, which contains the geometrical data of the taper and the wavelength.

Because in a multimode guide the comparison of the amplitudes does not give a clear picture, we show $(W_n/W_1)^{1/2}$, *i.e.*, the square root of the ratio of the power of the nth mode to that of the first mode. This quantity is plotted in Fig. 2 against σ, for $n=2$, 3, and 4.

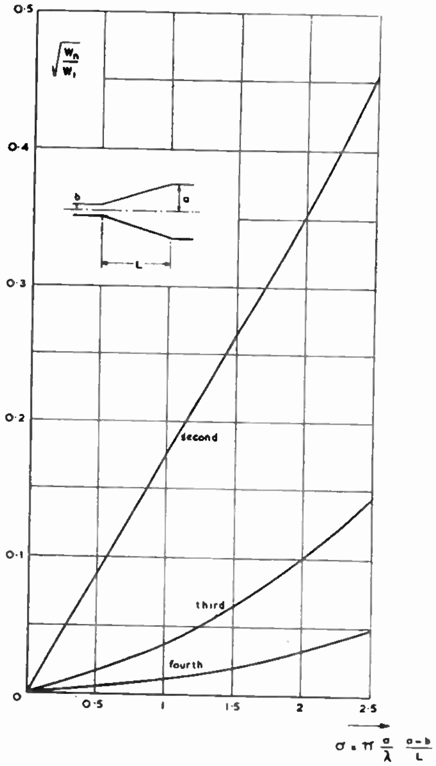

Fig. 2.

It may be seen that it is enough to take into account the power in the second mode. If the permitted ratio of the power in the second mode is given, σ may be obtained from Fig. 2 from which the taper is determined.

If $\sigma < 2$, then a very good approximation is

$$\sqrt{\frac{W_2}{W_1}} = 8 \frac{k_{12}k_{11}}{(k_{11}^2 - k_{12}^2)^2} \sigma = 0.181\sigma.$$

Hence the length of the taper may be obtained from the following formula

$$L \cong \frac{0.57}{\sqrt{p}} \frac{a}{\lambda} (a-b)$$

* Received by the IRE, November 28, 1957.

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where pW_1 is the permitted power in the second mode.

For example, if $a=5$ cm, $b=1.5$ cm, $\lambda=1$ cm,

$$p = \frac{W_2}{W_1} = 0.01,$$

then $L = 1$m.

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Measuring Noise Figures of Transistor Amplifiers*

The accepted method for measuring the noise figure of a tuned amplifier using a noise diode calls for the matching of the noise diode output impedance to the input impedance of the amplifier under test. In measuring the noise figure of transistor amplifiers, it is sometimes necessary to operate the transistor under unmatched conditions. This case requires a modification of the equations used in the matched condition measurements.

Fig. 1 shows the equivalent circuit of the noise diode feeding the amplifier under test. The mean square noise current is:

$$i_n^2 = 2eI_DB \tag{1}$$

where

e = electronic charge,
I_D = dc plate current of the noise diode,

* Received by the IRE, October 3, 1957.

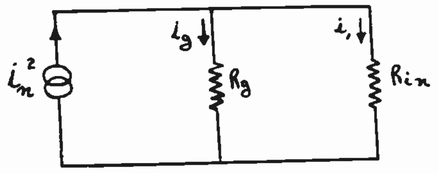

Fig. 1.

B = noise bandwidth,
R_g = source resistance of the noise generator,
R_{in} = input resistance of the amplifier under test.

In the accepted method of measuring noise figure, where R_g is made equal to R_{in}, the noise figure at 290°K is given by

$$\text{N.F.} = \frac{20I_DR_{in}}{M-1} \tag{2}$$

where

$$M = \frac{\text{noise power of the amplifier with diode ON}}{\text{noise power of the amplifier with diode OFF}}.$$

However, for the general case, where R_g is not necessarily equal to R_{in}, the following derivation is given.

The noise power delivered to the amplifier, P_d, is

$$\begin{aligned} P_d &= i_i^2R_{in} \\ &= \left(\frac{R_g}{R_g+R_{in}}i_n\right)^2 \times R_{in} \\ &= 2eI_DB \times \left(\frac{R_g^2R_{in}}{(R_g+R_{in})^2}\right) \\ &= 2eI_DBR_F \end{aligned} \tag{3}$$

where

$$R_F = \frac{1}{R_{in}}\left(\frac{R_gR_{in}}{R_g+R_{in}}\right)^2.$$

Let the noise power of the amplifier alone at the input be W_n, then the amplifier's noise figure at 290°K is:

$$\begin{aligned} \text{N.F.} &= \frac{W_n}{KTB} = \frac{W_n}{P_d} \times \frac{P_d}{KTB} \\ &= \frac{\dfrac{P_d}{KTB}}{\dfrac{P_d}{W_n}} = \frac{\dfrac{2eI_dBR_F}{KTB}}{\dfrac{W_n+P_d-W_n}{W_n}} \\ &= \frac{80 \times I_d \times R_F}{\dfrac{W_n+P_d}{W_n} - 1} \\ &= \frac{80 \times I_d \times R_F}{M-1} \end{aligned} \tag{4}$$

where

$$M = \frac{W_n+P_d}{W_n}.$$

A close study of (4) will show that if R_g is made equal to R_{in}, then the equation will reduce to (2), which is the expression for the noise figure of the matched condition case.

It can then be concluded that the commonly known (2) is a special case of the more general expression (4).

Also, for the general equation, the definition of the noise figure does not have to be based on the *available* power from the noise source.

D. H. Rice has given valuable suggestions in connection with this problem.

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Contributors

S. V. Chandrashekhar Aiya (A'39–M'40–SM'43) was born in Saklaspur, India on May 17, 1911. He received the B.S. degree in 1931 from Wilson College, Bombay, India and the B.A. degree with first class honors in 1934 from Gonville and Caius College, Cambridge, England. He was professor of radiophysics at S.P. College, Poona, from 1936 to 1942, experimental physicist at the Indian Institute of Science, Bangalore, from 1942 to 1945, and professor of electrical communication at the College of Engineering, Poona, India from 1945 to 1956. He now is Principal of the L.D. College of Engineering, Ahmedabad. In 1948–1949, he was Dean, Faculty of Technology, Bombay University.

S. V. C. Aiya

Mr. Aiya is India's representative on Commission IV of the URSI and a member of the Indian National Committee of the URSI.

He has served on several government committees and on university bodies in Bombay, Poona, and Ahmedabad. He is the author of several papers on atmospheric noise and has specialized in the tropicalization of electronic equipment.

Mr. Aiya is a full member of the Institution of Electrical Engineers.

W. R. G. Baker, for a photograph and biography, please see p. 1458 of the November, 1957 issue of Proceedings. Dr. Baker retired as Vice-President of General Electric Company in November, 1957, and is now Vice-President for Research, Syracuse University, Syracuse, N. Y.

❖

A. G. T. Becking was born on January 31, 1920. In 1952, he received the Doctor's degree, cum laude, for experimental and theoretical work on fluctuation-phenomena in bolometers at the University of Utrecht, The Netherlands. Prior to that time, he already had joined the Philips' Research Laboratories, Eindhoven, The Netherlands, in

1948, where he constructed a vectorcardiograph and did research work on hearing aids. He had to cope with acoustical, electronic, physiological, and psychological problems. He gave much assistance to the work at the school for severely deaf children in St. Michielsgestel.

A. G. T. Becking

The application of the transistor to hearing aids revived his interest in noise problems. In 1956, he attended the Acoustical Congress in Cambridge, Mass., although he already was suffering from a severe illness, which, on June 26, 1957, ended the life of a man who worked hard to help other people and from whose unselfish work many more would have benefited, if he had lived longer.

❖

Warren V. Behrens (M'56) was born on December 2, 1919, in Seattle, Wash. He received the B.S. degree in arts and science, in 1944, and the B.S. degree in electrical engineering in 1948, both from the University of Washington, Seattle, Wash. Following completion of the Navy's radar courses at Harvard University, Cambridge, Mass. and Massachusetts Institute of Technology, Cambridge, Mass., in 1945, he served as a technical electronics officer with the U. S. Naval Air Force. On return to civilian life, in 1947, he was employed by the Bonneville Power Administration, Portland, Ore., as a design engineer in communication and relaying, until June, 1952. From 1952 to 1955, he worked as an electronic engineer for the U. S. Naval Research Laboratory, doing research and development work in the field of infrared application.

W. V. Behrens

In 1955, he joined the Electromechanical Laboratory of the Diamond Ordnance Fuze Laboratories, where he is now employed as an electronic scientist, working principally on transistor applications.

❖

John C. Cacheris (M'48–SM'56) was born on May 18, 1916, in Chicago, Ill. He graduated from the Capitol Radio Engineering Institute in 1941. He received the B.S. degree from Carnegie Institute of Technology, Pittsburgh, Pa., in 1946, and the M.S. degree from Maryland University, College Park, Md., in 1953. From 1941 to 1946, Mr. Cacheris was employed as a radio engineer in the test department of radio division of the Westinghouse Electric Corp., Baltimore, Md. From 1946 until 1949, as an electronic scientist with the Naval Ordnance Laboratory in White Oak, Md., he designed circuits and instruments for ultra-high and microwave-frequency ranges.

J. C. Cacheris

He joined the staff of the Ordnance Development Division of the National Bureau of Standards, Washington, D. C., in 1949, where he engaged in microwave antenna and diffraction studies, and in investigations of the microwave properties of ferrites. He is continuing the latter investigations at the Diamond Ordnance Fuze Laboratories, Department of the Army, to which the functions and staff of the Ordnance Development Division were transferred on September 27, 1953. He is Chief of the Ferrite Research Section of the Supporting Research Laboratory.

Mr. Cacheris is a member of the American Physical Society and Eta Kappa Nu, and is a registered professional engineer in the District of Columbia.

❖

Malcolm R. Currie (S'52–A'55–SM'58) was born in Spokane, Wash., on March 13, 1927. From 1944 to 1947, he served in the U. S. Navy Air Corps. He received the A.B. degree in physics from the University of California, Berkeley, Calif., in 1949, and the M.S. and Ph.D. degrees from the same institution in 1951 and 1954, respectively.

M. R. Currie

Dr. Currie was associated with the Microwave Laboratory at the University of California from 1949 to 1951. From 1952 to 1954, he was a research assistant in the Electronics Research Laboratory. He was an instructor in the Department of Electrical Engineering during 1953–1954. Since 1954, he has been a member of the technical staff in the Electron Tube Laboratory, Hughes Aircraft Company, Culver City, Calif., where he has been concerned with problems in microwave electronics. At present he is co-head of the Microwave Tube Department. He also is a lecturer in engineering at the University of California at Los Angeles.

Dr. Currie is a member of the American Physical Society, Sigma Xi, RESA, and Phi Beta Kappa.

❖

Donald C. Forster (S'55–M'56) was born in Los Angeles, Calif., on September 28, 1928. He served in the U. S. Navy from 1946 to 1949 and from 1950 to 1952. He received the B.S.E.E. degree from the University of Southern California, Los Angeles, Calif., in 1955, and the M.S. degree from the California Institute of Technology, Pasadena, Calif., in 1957.

D. C. Forster

He has been working with problems in microwave electronics as a member of the technical staff in the Electron Tube Laboratory, Hughes Aircraft Company, Culver City, Calif., since 1955. At present he is working towards the Ph.D. degree at the California Institute of Technology under a Howard Hughes Fellowship.

Mr. Forster is a member of Phi Kappa Phi, Tau Beta Pi, and Eta Kappa Nu.

❖

Paul E. Green, Jr. (S'46–A'48–SM'55) was born in Durham, N. C., in 1924. He received the B.A. degree in physics from the University of North Carolina in 1944. After a brief tour of duty as a Navy electronics officer, he entered North Carolina State College, receiving the M.S. E.E. degree in 1948. He then joined the Agricultural Engineering Department of the same institution, to study photoelectric color measurement and grading techniques.

P. E. Green, Jr.

In 1949, Mr. Green became a research assistant in the Research Laboratory of Electronics at the Massachusetts Institute of Technology working on statistical techniques as applied to brain-wave and speech problems. In 1951, he joined Lincoln Laboratory, where he has since been concerned with the application of statistical communication theory to communication system problems. He received the Sc.D. degree from M.I.T. in 1953.

Dr. Green is a member of Tau Beta Pi and Sigma Xi.

❖

David Jaffe was born on June 11, 1929, in New York, N. Y. He received the B.S. degree in physics from Brooklyn College, Brooklyn, N. Y. in January, 1951, and the M.S. degree in June, 1952 from the University of Connecticut, Storrs, Conn., where he served as a physics laboratory instructor Mr. Jaffe was employed as a physicist in the Technical Evaluation Division

D. Jaffe

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of the U. S. Naval Ordnance Laboratory from 1952 to 1953. From 1953 to 1955, he served with the United States Army's Ballistic Research Laboratories, where he was engaged in studies of interior ballistics. Since 1955, he has been employed at the Diamond Ordnance Fuze Laboratories in their Supporting Research Laboratory. He has been engaged in the study of the microwave properties of ferrites.

Mr. Jaffe is a member of the American Physical Society.

❖

F. E. Jaumot, Jr.

Frank E. Jaumot, Jr. was born in Charleston, W. Va. on August 3, 1923. He received the B.S. degree in physics from Western Maryland College in 1947, and the Ph.D. degree in physics from the University of Pennsylvania in 1951.

After graduation he was employed as an instructor in physics at the University of Pennsylvania until he joined the Franklin Institute Laboratories in 1952, where he was chief of the Physics of Metals Section. While at the Institute, he remained on the staff of the University on a part-time basis, holding an appointment as visiting assistant professor of metallurgical engineering.

Dr. Jaumot joined the Delco Radio Division of General Motors Corp. in December, 1956, where he now holds the position of Director, Research and Development-Semiconductors. He has published more than twenty technical papers on various topics in the solid-state physics field.

Dr. Jaumot is a member of the American Physical Society, American Crystallographic Society, American Institute of Mining, Metallurgical and Petroleum Engineers, American Association of Physics Teachers, American Institute of Physics, Sigma Xi, and Scientific Research Society of America.

❖

N. Karayianis

Nicholas Karayianis was born on August 17, 1931, in Washington, D. C. He majored in physics at George Washington University, Washington, D. C., where he received the B.S. degree in February, 1954. He was awarded a Sanders Fellowship for graduate studies and joined the University faculty for a year as a physics laboratory instructor, receiving the M.S. degree in February, 1956. Since 1954, he has been employed by the Supporting Research Laboratory of the Diamond Ordnance Fuze Laboratories, where he has been investigating the microwave properties of ferrites.

Mr. Karayianis is a member of Sigma Xi.

❖

A. Papoulis

A Papoulis (SM'55) was born in Greece in 1921. He studied mechanical and electrical engineering at the Polytechnic Institute of Athens, Athens, Greece, from 1937 to 1942. He came to the United States in 1945 and continued his studies at the University of Pennsylvania, Philadelphia, Pa., where he was awarded the M.S., M.A., and Ph.D. degrees.

He is an associate professor of electrical engineering at the Polytechnic Institute of Brooklyn, Brooklyn, N. Y., where he has been teaching since 1952. He also has been a consultant with the Burroughs Corporation since 1951. Previously, he taught at the University of Pennsylvania and at Union College.

❖

R. Price

Robert Price (S'48–A'54) was born in West Chester, Pa., on July 7, 1929. He received the A.B. degree in physics from Princeton University, Princeton, N. J., in 1950, and the degree of Sc.D. in electrical engineering from the Massachusetts Institute of Technology, Cambridge, Mass., in 1953.

At the Massachusetts Institute of Technology, he held an Industrial Fellowship in the Research Laboratory of Electronics and later became a research assistant in the Lincoln Laboratory, where his thesis work on the problems of communicating through multipath disturbances was carried out. Upon completion of this study, he did research in radio-astronomy under a Fulbright award at the Commonwealth Scientific and Industrial Research Organization in Sydney, Australia.

Since 1954, Dr. Price has been a staff member of the Lincoln Laboratory and is currently concerned with various topics in statistical communication theory, decision theory and noise theory, in addition to communications system design and evaluation.

He is a member of Phi Beta Kappa, Sigma Xi, and the Franklin Institute.

J. M. Shaull

John M. Shaull (SM'52) was born in Hagerstown, Md., on August 31, 1910. He attended public schools in Charles Town, W. Va. He was employed as a radio and electrical serviceman in Charles Town and has held amateur radio licenses since 1930 and commercial radiotelephone licenses since 1937. In 1935, he came to Washington, D. C. and worked in several Government departments, during which time he graduated from Capitol Radio Engineering Institute and attended George Washington University. In 1938, he was employed in the engineering department of Bendix Radio Corp., Baltimore, Md., calibrating and adjusting precise frequency measurement equipment.

In 1939, Mr. Shaull joined the staff of the National Bureau of Standards as operator at the Bureau's standard-frequency radio station WWV. He later designed and supervised construction and installation of the frequency and time interval control equipment for the new WWV station, and the Bureau's microwave frequency standard. In 1946, he was placed in charge of monitoring the accuracy of the WWV standard frequency and time transmissions and improving the standards and methods associated with and constituting the primary standard of frequency.

In 1953, Mr. Shaull transferred to the Electromechanical Laboratory of the Diamond Ordnance Fuze Laboratories, where he is presently in charge of an engineering research section.

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A. van der Ziel

A. van der Ziel (SM'49–F'56) was born at Zandeweer, The Netherlands, on December 12, 1910. From 1928 to 1934, he studied physics at the University of Groningen, The Netherlands, where he received the Ph.D. degree in 1934.

He was a member of the research staff of the Physics Laboratory of N. V. Philips' Gloeilampenfabrieken, Eindhoven, The Netherlands, from 1934 to 1947. From 1947 to 1950, he was an associate professor at the University of British Columbia, Vancouver, Canada, and has been professor of electrical engineering at the University of Minnesota since 1950.

Dr. van der Ziel is a member of the American Physical Society and Sigma Xi.

World Radio History

1958 IRE National Convention Program

Waldorf-Astoria Hotel, New York Coliseum, March 24-27, 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 convenient location, the New York Coliseum at 59th St. and 8th Ave. A list of the 850 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 24.
Place: Grand Ballroom, Waldorf-Astoria Hotel.
Speaker: Ernst Weber, President, Polytechnic Institute of Brooklyn and IRE Director, "The Broad Spectrum."

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 26.
Place: Grand Ballroom, Waldorf-Astoria Hotel.
Guest Speaker: Robert C. Sprague, Chairman of the Board, Sprague Electric Co. and Federal Reserve Bank of Boston, "The Federal Reserve and the Electronics Industry."
Presentation of IRE Awards: Donald G. Fink, IRE President.
Spokesman for IRE Fellows: Patrick E. Haggerty, Executive Vice-President, Texas Instruments, Inc.
Toastmaster: G. F. Haller, General Manager, Defense Electronics Div., General Electric Co.

Cocktail Party

Time: 5:30–7:30 P.M., Monday, March 24.
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.

World Radio History

SCHEDULE OF TECHNICAL SESSIONS

* Sessions terminate at 12:00 Noon

	WALDORF-ASTORIA HOTEL					NEW YORK COLISEUM		
	Starlight Roof	**Astor Gallery**	**Jade Room**	**Sert Room**	**Grand Ballroom**	**Morse Hall**	**Marconi Hall**	**Faraday Hall**
Monday March 24 2:30 P.M.–5:00 P.M.	*Session 1* TUTORIAL SESSION ON DETECTION THEORY AND ITS APPLICATIONS	*Session 2* VEHICULAR COMMUNICATIONS	*Session 3* TELEMETRY AND REMOTE CONTROL	*Session 4* TECHNIQUES AND CRITERIA CONSIDERATIONS IN ELECTRONIC ENGINEERING	*Session 5* PANEL: EDUCATIONAL NEEDS IN SYSTEMS ENGINEERING	*Session 6* ENGINEERING WRITING AND SPEECH	*Session 7* RADIO FREQUENCY INTERFERENCE	*Session 8* ADVANCES IN PRODUCTION ENGINEERING
Tuesday March 25 10:00 A.M.–12:30 P.M.	*Session 9* AUTOMATIC CONTROL—General	*Session 10* CONTROLLED THERMONUCLEAR POWER	*Session 11* BROADCAST TRANSMISSION SYSTEMS	*Session 12* STEREOPHONIC DISC RECORDINGS	*Session 13** PLANNING AGAINST TIME	*Session 14* AERONAUTICAL AND NAVIGATIONAL ELECTRONICS	*Session 15* MEDICAL ELECTRONICS	*Session 16* GENERAL COMMUNICATIONS SYSTEMS
Tuesday March 25 2:30 P.M.–5:00 P.M.	*Session 17* PANEL: CHANGING DEMANDS ON THE BREADTH OF ELECTRICAL ENGINEERING EDUCATION	*Session 18* ATOMIC CLOCKS AND MASERS	*Session 19* BROADCAST TRANSMISSION SYSTEMS AND COMMUNICATIONS SYSTEMS	*Session 20* AUDIO, AMPLIFIER AND RECEIVER DEVELOPMENTS		*Session 21* BEAM AND DISPLAY TUBES	*Session 22* PANEL: BIOLOGICAL TRANSDUCERS	*Session 23* RELIABILITY THROUGH COMPONENTS
Tuesday March 25 8:00 P.M.–10:30 P.M.	*Session 24* PANEL: ELECTRONICS IN SPACE							*Session 25* PANEL: ELECTRONICS SYSTEMS IN INDUSTRY
Wednesday March 26 10:00 A.M.–12:30 P.M.	*Session 26* AERONAUTICAL AND NAVIGATIONAL ELECTRONICS	*Session 27* STATISTICAL APPLICATIONS	*Session 28* ELECTRONIC COMPONENT PARTS	*Session 29* CIRCUIT THEORY I AND ULTRASONICS I: SYMPOSIUM ON 'MODERN ASPECTS OF DELAY LINES'	*Session 30** THE CANADIAN AUTOMATION SYSTEM OF POSTAL OPERATIONS	*Session 31* RADAR IN MILITARY ELECTRONICS	*Session 32* MICROWAVE MEASUREMENTS	*Session 33* SEMICONDUCTOR DEVICES
Wednesday March 26 2:30 P.M.–5:00 P.M.	*Session 34* RELIABILITY THROUGH SYSTEMS	*Session 35* INFORMATION THEORY: CODING AND DETECTION	*Session 36* ELECTRONIC COMPONENT PARTS	*Session 37* COMPUTERS AND CONTROL		*Session 38* INSTRUMENTATION SYSTEMS	*Session 39* MICROWAVE COMPONENTS	*Session 40* PROPAGATION AND ANTENNAS I—General
Thursday March 27 10:00 A.M.–12:30 P.M.	*Session 41* MAGNETICS AND COMPUTERS	*Session 42* CIRCUIT THEORY II—Unusual Aspects of Filter Design	*Session 43* ULTRASONICS II—Delay Line Measurements	*Session 44* INDUSTRIAL ELECTRONICS	*Session 45** ASPECTS OF RF INTERFERENCE IN MILITARY ELECTRONIC AND COMMUNICATIONS SYSTEMS	*Session 46* DATA REDUCTION AND RECORDING	*Session 47* ANTENNAS II—General	*Session 48* MICROWAVE TUBES
Thursday March 27 2:30 P.M.–5:00 P.M.	*Session 49* GENERAL SYSTEMS	*Session 50* CIRCUIT THEORY III—Application of Topological and Group Concepts	*Session 51* ULTRASONICS III—Measurement of Radiated Acoustic Power	*Session 52* LONG DISTANCE COMMUNICATIONS		*Session 53* HIGH ACCURACY INSTRUMENTS, MEASUREMENT AND CALIBRATION	*Session 54* ANTENNAS III—Microwave Antennas	*Session 55* RADIO AND TELEVISION

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ABSTRACTS OF TECHNICAL PAPERS

SESSION 1*

Mon. 2:30–5:00 P.M.

Waldorf-Astoria
Starlight Roof

TUTORIAL SESSION ON DETECTION THEORY AND ITS APPLICATIONS

Chairman: D. SLEPIAN, *Bell Telephone Labs., Inc., Murray Hill, N. J.*

1.1. Detection as a Statistical Decision Problem

D. VAN METER, *Melpar, Inc., Boston, Mass.*

The detection of signals in noise has for a number of years now been recognized as an operation analogous to hypothesis testing in statistics. This paper presents a review of the theoretical approach that has proved useful in treating detection from this point of view with particular attention given to Wald's Statistical Decision Theory. The variety of decision situations encountered in detection and the importance of careful problem formulation are emphasized. General results in the theory are illustrated with simple examples.

1.2. Some Communications Applications of Detection Theory

W. B. DAVENPORT, JR., *Lincoln Lab., Massachusetts Institute of Technology, Lexington, Mass.*

Two specific systems are discussed which illustrate the application of statistical detection theory to practical communication problems. The first system, called RAKE, is designed to provide optimum reception of a wide-band signal which has been transmitted over a noisy multipath channel. The second system, called MAUDE, is designed to automatize the reception of hand sent Morse code signals. The statistical principles underlying the design of both systems are presented.

1.3. Some Applications of Detection Theory to Radar

W. M. SIEBERT, *Elec. Eng. Dept., Massachusetts Institute of Technology, Cambridge, Mass.*

Various ways are considered in which the radar detection problem can be stated and organized. Particular emphasis is placed on the role of Decision Theory and the concept of sufficient statistics. Simple examples of the application of the theory are presented. A discussion of the practical implications and limitations of the simple theory suggests several extensions, one of which is treated in some detail to illustrate more sophisticated applications of the theory.

1.4. Human Factors in Detection and Speech Communications

J. P. EGAN, *Psych. Dept., Indiana University, Bloomington, Ind.*

Detection Theory has led to reformulation of those psychophysical methods which are used to investigate the detection and recognition of signals in noise. The operating characteristic of the human observer is defined, and the role of the criterion in the yes-no judgment is thereby made explicit. Results of forced-choice experiments are related in the model to the more realistic yes-no situation. Application of these concepts to speech communication in noise is discussed. The decision by the communicators to repeat a message or to move on to the next one may also be described by an operating characteristic.

* Sponsored by the Professional Group on Information Theory. To be published in Part 4 of the 1958 IRE NATIONAL CONVENTION RECORD.

SESSION 2*

Mon. 2:30–5:00 P.M.

Waldorf-Astoria
Astor Gallery

VEHICULAR COMMUNICATIONS

Chairman: A. A. MACDONALD, *Motorola, Inc., Chicago, Ill.*

2.1. Direct Despatch Service

A. J. DINNIN, *The Bell Telephone Company of Canada, Montreal, P. Q., Can.*

A new mobile radio service has been introduced in the Bell Telephone Company of Canada. Its purpose is to provide more efficient use of the frequency spectrum by permitting up to twenty dispatchers to share common radio system. Commercially available radio equipment, combined with the specially designed tone signaling system, provides a high grade of transmission and an equitable sharing of air time. Installation has been completed by the Bell Company in five cities in Ontario and Quebec.

2.2. A Unique Radio System Designed for Flood Forecasting

W. C. WRAY, *Motorola, Inc., Chicago, Ill.*

The Commonwealth of Pennsylvania, Department of Forests and Waters, has recently contracted for a communications and alarm system throughout the Susquehanna River Basin, to provide instant and accurate information on the condition of water levels in the various branches of the river. This information is relayed daily, or as often as required, from the remote sites to a central control point located at the U. S. Weather Bureau Office at the Harrisburg Airport.

A variety of facilities are utilized for this purpose and include automatic equipment for transmission of coded signals, as well as manually operated stations which report by voice.

The main backbone system is comprised of 960-mc equipment. Over this baseband are carried the various multiplex speech circuits, teletypewriter, telemetering, etc. At most of the repeater sites, a 170-mc or 400-mc base station is utilized for communication with the many outlying reporting stations. The system is equipped with failure alarm protection, so that outages can be corrected in short order.

2.3. A New Approach to Broad-Band Vehicular Antennas

H. BRUECKMANN, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

The standard communication antenna for military ground vehicles today is a sectionalized metal whip insulated and fed at its base in such a way that the metal body of the vehicle acts as a counterpoise. It is not a satisfactory solution for a modern army. USASEL has taken a new approach based on previous research which had revealed the existence of distinct modes of electrical resonance of the vehicle body and the dependence of the coupling between these modes and the antenna on the position of the antenna and its electrical length. The new antenna minimizes the coupling to the vehicle body by eliminating its use as a counterpoise. In addition, several new techniques are applied, such as: methods of producing extremely strong hollow fiberglass rods, plastic insulators and ferrite materials of high permeability and Q, and simple methods of synthesizing broad-band matching networks.

2.4. Mobilization of Transistors

R. E. HANSEN, *General Electric Co., Syracuse, N. Y.*

The advent of semiconductors points to a revolution in the electronic industry. The field of mobile communication is not left untouched by this flurry of activity. Like many of the parameters that the engineers deal with, transistorization of any product is a complex expression having real and reactive components. The j components are the entourage of magical and wondrous claims that serve to cloud the "real" component—the solid advantage of transistorization. The mobilization of transistors is examined in terms of recently developed products and laboratory proposals.

2.5. Vehicular Noise Problems in Modern Land Mobile Systems

S. F. MEYER, *Allen B. DuMont Labs., Inc., Clifton, N. J.*

With the advent of higher compression ratios in automotive engines, the vehicular communications industry finds itself in the midst of a performance degradation problem. The situation is further aggravated by the ever increasing steeper selectivity curves of the associated mobile receiver as a result of squeezing the precious channel separation.

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This paper discusses the basic problem in modern vehicular installations, together with a number of suppression techniques. Also covered is the joint effort of the individual committees, representing the Society of Automotive Engineers, the Electronics Industry Association, and the Institute of Radio Engineers, currently working on the problem.

SESSION 3*

Mon. **2:30–5:00 P.M.**

Waldorf-Astoria
Jade Room

TELEMETRY AND REMOTE CONTROL

Chairman: K. T. LARKIN, *Lockheed Missile Systems Div., Palo Alto, Calif.*

3.1. The RCA Flight Data System

C. N. BATSEL, JR., R. E. MONTIJO, JR., AND E. J. SMUCKLER, *Radio Corp. of America, Los Angeles, Calif.*

An integrated electronic system for in-flight monitoring and recording of transducer data is described. The system is designed to meet the need of modern high-performance flight vehicles for recording the outputs of a large number of transducers at high accuracy and for wide ranges of data frequency.

The system embodies the latest techniques in digital data-handling. Analog transducer data are converted in the aircraft to digital form with a resolution of 10-binary bits plus sign. Inputs from digital transducers also may be accepted by the system. The digital data may be recorded on magnetic tape in the aircraft. Simultaneously with recording, the data are transmitted to the ground at the full rate by pulse code modulation techniques.

At the ground collection station, the transmitted data may be recorded on magnetic tape. Visual monitoring identical to that in the aircraft is provided, together with a direct writing presentation. At the computer site, the data collected in the ground station or in the flight vehicle may be edited either on-line or off-line for conversion into a form suitable for input to a digital computer.

3.2. A Pulse Position Telemetry System

L. WEISMAN AND E. S. TELTSCHER, *Ford Instrument Co., Long Island City, N. Y.*

This paper essentially describes a miniaturized pulse position telemetry system. Reference is made to a paper by J. W. Poliseo, contained in the National Telemetering Conference, Chicago, 1955. The whole system contains an airborne radar beacon transponder, telemetering pulse position modulated signals to the ground. The system capacity is 9 data channels plus one reference and synchronizing channel. The ground equipment has been highly miniaturized, consisting of a decommutator and separate demodulator channels. The demodulators have been transistorized, using a novel compact pulse position demodulator. The introduction of a quasi-feedback type link between the airborne transmitter and the ground equipment demodulator ensures highly stable and accurate operation during flight.

3.3. Sample and Hold Circuits for Time Correlation of Analog Voltage Information

W. T. EDDINS, *Radiation, Inc., Melbourne, Fla.*

This paper gives theoretical consideration and an example of a circuit for taking samples of an analog voltage at well-defined points in time, then storing these samples for readout at later times. Since the voltage being sampled represents information, it is necessary to have a highly accurate system or loss of information will occur.

The example discussed accepts signals in the range from +5 volts to −5 volts, holding the samples for a period of 1000 μsec to an accuracy better than 10 mv. The period required to accomplish taking of the sample is 100 μsec, but the point in time which the sample represents is defined to less than 1 μsec. This makes possible time correlation of analog signals to better than 1 μsec, even though readout of the various signals may vary several hundred microseconds.

3.4. A Transistorized Six-Channel Airborne Digitizer

S. H. MCMILLAN AND W. A. SUTTON, *Strand Eng. Co., Ann Arbor, Mich.*

A six-channel shaft position digitizer with 0.1 per cent accuracy for missile telemetering is described. Transducers having less than 0.2 oz.-in. torque and 0.025 oz.-in.^{2} inertia resolve 3° shaft movements up to 600 rpm. Digitized information is stored and electronically commutated into a standard fm-fm telemeter. A 40-bit word, transmitted every 0.01 second, contains information on all shaft movements since the previous word; absolute position of one shaft from time of preflight calibration; shaft identification; sync; and parity. Six consecutive words contain absolute position of all shafts. The 5500 components required, including 500 silicon transistors, are "potted" in modules of approximately 4 cubic inches, and occupy a space 12" by 12" by 3". Power drain is 1.4 amperes at 27 volts.

3.5. Channel Selection for Multicarrier Telemetry

L. S. TAYLOR AND G. F. BIGELOW, *Range Instrumentation Development Div., White Sands Proving Ground, N. M.*

Complex telemetry systems using a number of carrier frequencies are now in use. Important intercarrier effects resulting in large data errors can occur in this type of system. The problem of reducing intercarrier modulation by the proper selection of carrier frequencies from the uniformly spaced set within the telemetry band is, therefore, of considerable interest. This problem has been studied on an analytical basis and sets of optimum carrier frequencies have been determined for multicarrier systems. These sets were derived graphically by considering the carrier frequencies to be points in a multidimensional lattice space in which the intercarrier effects are restricted to hyperplanes.

3.6. Telemetering Receiving Station Time Pulse Detector

J. STAR, *Appl. Phys. Lab., The Johns Hopkins University, Silver Spring, Md.*

This paper describes a time pulse detector which will recover periodic field timing signals and superimposed real time code from tone bursts recorded on missile flight tape. The unit supplies appropriate timing pulses to receiving station equipment which processes the telemetered information also contained on the tape.

The time pulse detector contains an agc selective amplifier tuned to the carrier frequency of the bursts, allowing an input level of 0.2 to 10 volts peak to peak without operator adjustment. Input repetition rates of 10 to 100 pps are detected and counted down to 1, 5, 10 pps as desired by cold cathode glow transfer tubes (Dekatrons).

SESSION 4*

Mon. **2:30–5:00 P.M.**

Waldorf-Astoria
Sert Room

TECHNIQUES AND CRITERIA CONSIDERATIONS IN ELECTRONIC ENGINEERING

Chairman: J. M. BRIDGES, *Office of Asst. Secretary of Defense (Research and Eng.), Washington, D. C.*

4.1. Use of Kros-Term System for Quick Retrieval of the Technical Details from Large Pools of Information

A. P. VIGLIOTTA, *U. S. Navy Training Device Center, Port Washington, N. Y.*, AND K. D. SWARTZEL, *Engleman & Co., Washington, D. C.*

Because of numerous extensive research programs, the frontiers of scientific knowledge are rapidly being expanded, and vast pools of technical information are being accumulated by various libraries in the form of reports, theses, articles, books, and other publications. Earlier methods of indexing these data have not been adequate to retrieve ***valuable detail*** once these publications have been reviewed and filed, and as a result much time is lost in searching for data and, in many instances,

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* Sponsored by the Professional Group on Military Electronics. To be published in Part 8 of the 1958 IRE NATIONAL CONVENTION RECORD.

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much money is spent in duplicating work already accomplished.

The use of the Kros-Terming system in the U. S. Navy Training Device Center is described to illustrate the rapid method being used for indexing, storing, and retrieving the valuable detail from the large pool of engineering data developed over the last decade at the Center. The relationship of the use of such systems to the conservation of time, manpower, and resources and the meeting of critical schedules is emphasized.

4.2. Techniques for the Presentation of Three-Dimensional Information

E. J. KENNEDY AND E. F. LAFORGE, *Rome Air Development Center, Griffiss Air Force Base, N. Y.*

The design of any display is dependent upon the unique characteristics of the situation which requires representation to a human observer. We may employ several methods of presenting three-dimensional data on objects located in a space-volume. Scale models have inherent limitations at the present time, which tend to limit their utility compared to a two-dimensional display with symbolic coding of the third dimension.

4.3. Transistorized Airborne Military Television Techniques

J. J. KELLY, *Norden-Ketay Corp., Stamford, Conn.*

A closed circuit television system for airborne military use is described. The system is completely transistorized, except for camera pickup device and monitor display tube.

Techniques involved in the design of this system are discussed. Particular circuits are treated in some detail. Growth potential of the system in the light of new component developments is discussed.

4.4. Design Criteria for Missile Automatic Test Equipment

W. O. CAMPBELL, *The Martin Co., Baltimore, Md.*

How is it possible to produce automatic test equipment with a performance reliability higher than the system it is designed to test? This problem has defied solution for several years. Although specification writing, component selection, circuit derating, and environmental testing are all necessary, it is pointed out that none of these procedures can be counted on to produce a test equipment *more* reliable than the missile.

Why is this so? The reason is that the missile designers also are writing specifications, selecting components, derating circuits, and making environmental tests.

A new approach is necessary. The evolution of this approach leads to a new logic and yields certain rules of thumb—or design criteria.

A performance reliability greater than that of the missile can be realized by judicious use of the following design criteria incorporated in a simple, general method of testing in the design of automatic test equipment: 1) self-verification on every step; 2) integral malfunction localization; and 3) automatic stand-by verification.

Using these criteria, results approaching 100 per cent already have been achieved.

4.5. Active Space-Frequency Correlation Systems

W. E. KOCK AND J. L. STONE, *Bendix Systems Div., Ann Arbor, Mich.*

The principle of space-frequency equivalence, which permits the spatial complexity of an antenna to be reduced by increasing the frequency complexity of the detection system, is considered for use in active systems. In such systems, a group of frequencies spaced over several octaves is transmitted and received by two or three widely spaced antennas having low directivity, but the received signals are correlated so as to achieve a highly directional receiving beam. Scanning is accomplished by introducing relative delays between the individual antennas. It is concluded that in situations requiring a broad transmitting radiation pattern and a sharp receiving pattern, such systems offer a significant reduction in antenna complexity.

SESSION 5*

Mon. 2:30–5:00 P.M.

Waldorf-Astoria Grand Ballroom

EDUCATIONAL NEEDS IN SYSTEMS ENGINEERING

Chairman: R. P. JOHNSON, *The Ramo-Wooldridge Corp., Los Angeles, Calif.*

Panel Members: H. CHESTNUT, *General Electric Co., Schenectady, N. Y.*

H. H. GOODE, *Dept. Elec. Eng., University of Michigan, Ann Arbor, Mich.*

S. HERWALD, *Westinghouse Electric Corp., Baltimore, Md.*

R. J. KOCHENBURGER, *Dept. Elec. Eng., University of Connecticut, Storrs, Conn.*

W. K. LINVILL, *Dept. of Defense, Washington, D. C.*

J. MOORE, *North American Aviation, Inc., Downey, Calif.*

Present trends toward large systems emphasize the need for a broader and more theoretical background for the engineer. The panel will discuss the role of systems engineering in the industrial and military fields and its impact on engineering education. Questions the panel will consider are: What new demands are placed on our engineering curriculum? Are present courses and laboratories appropriate? Are our educational institutions changing to meet the need of systems engineering? What new approaches are being used? What are the obstacles to progress? How does education in secondary and elementary schools match the trends in college engineering?

* Sponsored by the Professional Group on Automatic Control. To be published in Part 4 of the 1958 IRE NATIONAL CONVENTION RECORD.

SESSION 6*

Mon. 2:30–5:00 P.M.

New York Coliseum Morse Hall

ENGINEERING WRITING AND SPEECH

Chairman: A. A. MCKENZIE, *McGraw-Hill Book Co., Inc., New York, N. Y.*

6.1. Roadblocks in Technical Writing

T. GRIGGS, *Bendix Aviation Corp., Teterboro, N. J.*

Technical writing is, and must remain, more an art than a science because communication is dynamic, not static. Five pitfalls in writing are cited: rigorous word selection, congested phrasing, rigid form, situational notation, and subjective perspective. To counteract them, positive suggestions are given to writers: 1) start with an over-all view; 2) look for topics and headings; 3) use plenty of words in drafts; 4) correct drafts by reading rapidly; and 5) visualize a particular reader. In the management of writing, collaboration between engineers and technicians and nontechnical persons skilled in writing is recommended as being both efficient and effective, and as saving engineering time.

6.2. Writing for a Technical Journal

E. T. EBERSOL, JR., *Electronic Design, New York, N. Y.*

This paper discusses important factors which should be considered carefully by engineers planning publication of an article in a technical journal. Some of the factors often overlooked include determination of the reader audience desired, selection of the appropriate publication, investigation of the writing style of the magazine, and method of obtaining manuscript acceptance with minimum of changes. The elements discussed include survey of reader audience and magazine circulation, contact with editors to determine style and content for the manuscript, writing of the article and means of obtaining necessary approvals and clearance, and payment considerations. Factors which generally make for high readership interest are given, as well as many reasons for nonacceptance of the manuscript by a technical journal.

6.3. Nontechnical Help for Engineer-Writers

R. B. MACPHERSON, *Daystrom, Inc., Murray Hill, N. J.*

The assistance that public relations or other nonengineering groups can give to engineers in the preparation and presentation of technical papers is described. Technical articles and papers are of great value to the public relations and sales efforts of a company. Assistance

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which can be given before a paper is written is described, including discussion to assure the proper emphasis from the company's and the magazine's standpoint, contact with magazine editors, and planning of press coverage. After the first draft is prepared, additional help can be given, including editorial assistance, aids for oral presentation, and training in public speaking. When the paper is completed, the public relations department or similar group can arrange for the proper audience, technical and general press coverage, reprints, and recognition for the author within the company. The need for such assistance to engineer-writers is discussed, and its value to the company is described.

6.4. We Are What We Say

A. Henesian, *Lockheed Corp., Sunnyvale, Calif.*

This paper describes basic principles which will help the technical man to develop or improve his public-speaking ability. Items discussed include: 1) the elements of a talk designed to capture and hold listener interest; 2) methods of preparing a talk; 3) methods of opening and closing the speech; and 4) "secrets" of successful delivery. Technical personnel can become proficient in public speaking and thus promote better and more interesting talks in local IRE Professional Groups.

6.5. The Automatic Creation of Literature Abstracts (Auto-Abstracts)

H. P. Luhn, *IBM Corp., Yorktown Heights, N. Y.*

A process will be described by which it has become feasible to create excerpts from technical papers and magazine articles automatically to serve the purposes of conventional abstracts. The complete text of an article in machine readable form is scanned by a data-processing machine, such as the IBM 704, and analyzed in accordance with a standard program. Statistical information is derived, involving frequency and distribution of words in the text. The machine then computes a relative measure of significance, first of the words and then of the sentences. Sentences scoring highest in significance are extracted from the text and printed out to become the "Auto-Abstracts."

SESSION 7*

Mon. 2:30–5:00 P.M.

New York Coliseum
Marconi Hall

RADIO FREQUENCY INTERFERENCE

Chairman: E. J. Isbister, *Sperry Gyroscope Co., Great Neck, N. Y.*

7.1. Bandwidth Conservation in Pulse Modulated Radars

R. A. Rosien and R. Shavlach, *The Moore School of Elec. Eng., University of Pennsylvania, Philadelphia, Pa.*

Bandwidth can be conserved in radar systems by proper choice of the modulating pulse. Although the rectangular pulse is widely used, half cosine, cosine squared, and Gaussian waveforms are found to be more economical in generated bandwidth. Two pulses were developed by different means with the ideal of reducing the magnitude of the transmitted sidebands. Quasi-Gaussian waveshapes were formed by successive integrations of an input rectangular pulse. Cosine squared pulses were obtained by adding unity to a half cosine waveform and gating. The spectra produced by these waveshapes modulating an rf carrier demonstrates total power being concentrated nearer the carrier frequency.

7.2. Measurement of Spurious Radiation from Missileborne Electronic Equipments

A. L. Albin and C. B. Pearlston, *Filtron Co., Inc., Flushing, N. Y.*

Efficient utilization of the radio-frequency spectrum and the reduction of mutual interference require the imposition of limitations on radiation of harmonic and spurious emissions from missileborne electronic equipments. Present interference specifications used for control of radiated and conducted interference from electronics equipment have been restricted to an upper limit of 1000 mc. Since interference measurements are generally made in shielded enclosures, reflections from the walls and anomolies of measurements in the near-field have imposed great difficulties in extending the upper frequency limit. Recent availability of field intensity measuring equipment in the range of 1000–10,000 mc and the urgent need for establishing interference limits applicable to airborne or missileborne microwave equipment prompted an experimental program to establish techniques for radiated interference measurements and to determine interference limits in this frequency range.

Propagation tests made in a shielded enclosure in accordance with the prescribed techniques are shown to yield reproducible data. The correlation to measurements made on actual equipments is demonstrated. Parameters for guidance of microwave equipment designers in establishing tolerable interference limits are presented.

7.3. Small, Lightweight, RF Interference Suppressors Using Transistors

W. Pecota, *Sperry Gyroscope Co., Div. of Sperry-Rand Corp., Great Neck, N. Y.*

A conventional filter designed to suppress relay coil and heater thermostat switches often presents unsurmountable weight and packaging difficulties. RFI filters using only a transistor and a resistor may be used to replace conventional types in direct current circuits, thus reducing filter weight to less than five per cent and eliminating shielding, special wire routing, and filter mounting and space problems.

Successful design of a transistor filter requires matching of load, switch, and transistor characteristics heretofore not considered by the filter design engineer. A complete picture of the effects of alpha-cutoff, carrier diffusion, contact potential, voltage, current, and load impedance on the amount of RFI suppression obtained with a transistor filter is presented.

7.4. Transmission Interference in Low-Level Instrumentation Systems

J. C. Crosby, *Consolidated Electrodynamics Corp., Pasadena, Calif.*

In the field, instrumentation engineers often are troubled by line frequency signals interfering with desired transducer signals. Such difficulties are summarily dismissed as "ground loops" but frequently their mechanism is not understood. Some of the common reasons for this interference are discussed in connection with strain gauges, thermocouples, low-level amplifiers, and galvanometers. Simple working definitions are presented which, it is hoped, will eliminate the confusion about common mode signals.

7.5. Spurious Frequency Measurement in Waveguide

M. Morelli, *Rome Air Development Center, Griffiss Air Force Base, N. Y.*

This paper describes two techniques which are being developed for measuring within the waveguide spurious frequencies and their modes which may be generated and propagated. Also described in this report are free-space measuring techniques for obtaining spurious frequencies in the far field and Fresnel field of a radar set.

SESSION 8*

Mon. 2:30–5:00 P.M.

New York Coliseum
Faraday Hall

ADVANCES IN PRODUCTION ENGINEERING

Chairman: J. Davis, *Motorola, Inc., Chicago, Ill.*

8.1. Automatic Transistor Classifier

F. J. Morcerf, *General Electric Co., Syracuse, N. Y.*, and L. F. Roehm, *General Electric Co., Schenectady, N. Y.*

Transistors must be handled, tested, and classified in the final stages of their manufacture. A machine which performs these functions has been designed and built. DC low current and audio tests are now being performed, while high-frequency testing is being planned for the future. Individual test stations read out in terms of go-no-go or one of four possible levels. Units are separated into as many as ten classifications by means of a novel mechanical memory system.

* Sponsored by the Professional Group on Radio Frequency Interference. To be published in Part 8 of the 1958 IRE National Convention Record.

* Sponsored by the Professional Group on Production Techniques. To be published in Part 6 of the 1958 IRE National Convention Record.

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8.2. Circuit Packaging and Integration of Transistor Assemblies

H. H. HAGENS, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

The increasing use of transistors and the application of printed wiring and automatic assembly required new developments to establish practical and effective systems of packaging and integrating Auto-Sembled transistorized printed circuits.

Providing this type of information has been the objective of a Signal Corps research and development contract with P. R. Mallory and Company, Incorporated, Indianapolis, Ind., entitled, "Packaging and Integration of Transistor Assemblies." The technical requirements of the contract encompass two general circuitry categories. One category consisted of development with a transistorized computer circuit. The second involved development with a transistorized receiver circuit.

This paper will present pertinent data and design criteria developed from the contract.

8.3. Automatic Soldering Machine for Printed Circuit Assembly Boards

W. L. OATES, *Radio Corp. of America, Camden, N. J.*

The design and development of the new conveyor-type production machine for automatically soldering various printed circuit assembly boards was handled by the Equipment Development Section and presently is operating in the Automation Department, Building 5-6, Camden, N. J.

One operator loads and unloads the assemblies in adjustable pallets which constantly move around the track. The machine applies liquid flux, pre-heats the board, solders the assembly, and cleans off the excess flux before returning the pallet to the unloading position. The solder applicator is unique, being the subject of an RCA patent. Solder is pumped from a bulk supply at regulated temperature and flows down an inclined plane, across a number of ridge strips attached to the sloping surface, forming the solder into a number of smooth crests or waves. The board surface to be soldered is moved parallel to the slope of the plane causing any point on the board to come into contact with each solder wave in succession. The arrangement is basically simple, is extremely stable, not sensitive to changes in solder level, or requiring any dross removal. It will operate at a production rate of 7 to 15 feet per minute, and gives 280 to 600 finished assemblies per hour.

8.4 Some Economic Aspects of Wire Processing for Low Volume Production of Electronic Gear

J. BENSION AND R. D. PETERS, *General Electric Co., Utica, N. Y.*

The LMEE Department of General Electric has conceived a series of versatile, programmed machines for processing hookup wires economically in the relatively low quantities required in military electronic manufacturing. The color-coding machine already developed meets specifications for 90 per cent of the Department's wire requirements. Reduction in expensive setup time, elimination of large in-stock inventories of color-coded wire, and shortened lead line on wire orders are prominent advantages already realized by this program. The goal is to integrate wire stock, wire selection, color coding, induction bonding, wire measuring, cutting and stripping, and wire bundling, all into a single system controlled and operated from a single console, and programmed by punched cards. Programming will compensate in efficient handling of materials for the absence of large volume in the military electronics business.

8.5. "Case" History

T. C. COMBS, *Zero Manufacturing Co., Burbank, Calif.*

Pressure-tight instrument, combination and transit cases from aluminum deep drawn boxes are discussed in detail. Their increasing use is based on over-all suitability and trim appearance.

Instrument cases, both functional and industrial design versions, are described. For limited runs, the availability of stock deep drawn aluminum boxes provides savings in tooling costs and production time. For all sizes of run, the availability of specially designed bezels and accessory fabrication is essential.

Among the newest of products is the Modular Case, a recent development for packaging and shipment of large gear such as missiles, missile components, and instrumentation.

A number of practical examples are given.

8.6. Tension in Coil and Tape Winding

E. J. SAXL, *Tensitron, Inc., Harvard, Mass.*

Tension is important during winding of coils, potentiometers, and other wirewound products. Tension also influences tape for audio recording and data handling.

Some of the facts are discussed that limit tension with reference to material constants of wires and tapes.

A theory for Safe Operating Tensions for wires is presented (and the expression S.O.T. suggested as a new parameter for coil winding operations). It is extended for tapes, condenser foils, and films. A nomogram based upon wire size vs Safe Operating Tension is shown.

The moment of inertia is discussed as it influences tensions of creels during unwinding and rewinding. Devices for checking the winding tension of wire and similar materials are shown.

SESSION 9*

Tues. 10:00 A.M.–12:30 P.M.

Waldorf-Astoria Starlight Roof

AUTOMATIC CONTROL—GENERAL

Chairman: J. M. SALZER, *The Magnavox Co., Los Angeles, Calif.*

* Sponsored by the Professional Group on Automatic Control. To be published in Part 4 of the 1958 IRE NATIONAL CONVENTION RECORD.

9.1. A Servopressure Control System for the Iron Lung

G. A. BIERNSON, *Sylvania Electric Products, Inc., Waltham, Mass.*, AND J. E. WARD, *Servomechanisms Lab., Massachusetts Institute of Technology, Cambridge, Mass.*

Ordinary body respirators for care of patients with respiratory difficulty have rudimentary pressure control systems—usually a large bellows actuated by a crank or cam. Pressure patterns cannot be controlled precisely, because inevitable leaks in the tank at the neck opening and elsewhere make it impossible to hold any steady pressure different from atmospheric. Also, pattern shapes and amplitudes cannot be adjusted easily because of the mechanical linkage. To study how pressure patterns affect patient comfort and therapeutic effectiveness and to experiment with operation in which the patient is part of a closed-loop, a servocontrolled pressure system has been desgined as a research tool for the Harvard Medical School of Public Health.

The paper describes the design of the pressure-control system in detail and includes both the mathematical background and the hardware design. It also includes representative data obtained during patient testing.

9.2. Gain-Phase Relations of Nonlinear Circuits

E. LEVINSON, *Sperry-Gyroscope Co., Div. of Sperry Rand Corp., Great Neck, N. Y.*

This paper describes a novel property of certain nonlinear circuits and its application to the stabilization of feedback control systems. This property is one whereby the steady-state "describing" frequency response exhibits gain-phase relationships that are proven to be impossible for linear circuits. Any conceivable gain-phase combination can be obtained with appropriate nonlinear circuits. The desirable relations for servo stabilization are gain reduction and/or phase lead with increasing frequency. Several examples of nonlinear circuits exhibiting gain reduction without phase change are described and analyzed. The operation of one particular circuit with minimized dependence upon input magnitude is verified experimentally. This circuit is then shown experimentally to aid in the stabilization of a servomechanism.

9.3. On the Design of Adaptive Systems

H. L. GROGINSKY, *Electronics Research Labs., Columbia University, New York, N. Y.*

Parameter controlled adaptive mechanisms are designed using the method of dynamic programming. The method enables the optimal design of the control element for continuous or sampled data systems to be determined using criteria heretofore largely ignored. Experience gained with the unadapted system is used to aid the design. The design of a second order system with control exerted through a variable gain element is illustrated.

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9.4. The Organization of Digital Computers for Process Control

G. Post and E. L. Braun, *Litton Industries, Beverly Hills, Calif.*

The problems of process control sharply indicate the need for digital control and computing methods to provide adequately and economically the complexity of responses required. The prime requisite is a super-reliability which seems somewhat inconsistent with the state of the art of the general purpose digital computer. An incremental memory-based digital computer seems to have definite advantages in achieving reliable open or closed loop operation of 4000 hours without major equipment failure. Initial investigations reveal that even with vacuum tube circuitry, and eight hour overhaul periods occurring every 4000 hours, a useful and adequate control computer for process control can be produced.

Versatility and flexibility are maintained by a large memory capacity. A typical control function for a fractionating tower can be studied and a comparison made of control by an analog computer, a general purpose digital computer, and an incremental digital computer to show the power of memory-based techniques for process control applications.

9.5. A Self-Adjusting System for Optimum Dynamic Performance

G. W. Anderson, J. A. Aseltine, A. R. Mancini, and C. W. Sarture, *Aeronutronic Systems, Inc., Glendale, Calif.*

This paper describes the development of a technique which allows a system to automatically adjust its parameters for optimum dynamic performance. The method is applicable to all linear systems and has been verified experimentally in the laboratory.

A system which will adjust for changes in its environment must perform the following functions: 1) provide means for continuous, noninterfering measurement of system performance; 2) generate figure of merit from measured performance suitable for adjustment of system parameters; and 3) control parameters in accordance with changes in figure of merit.

Based on these criteria, a system has been designed and mechanized on an analog computer. The results of the simulation together with descriptions of special equipment are presented.

SESSION 10*

Tues. **10:00 A.M.–12:30 P.M.**

Waldorf-Astoria
Astor Gallery

* Sponsored by the Professional Group on Nuclear Science. To be published in Part 9 of the 1958 IRE National Convention Record.

CONTROLLED THERMONUCLEAR POWER

Chairman: E. W. Herold, *C. Stellarator Associates, Princeton, N. J.*

Important research is being conducted all over the world aimed at utilizing nuclear fusion reactions for power generation. One measure of the significance of this research is that success would permit us to tap a reservoir of energy sufficient to supply power at 1000 times the present rate for a billion years. A measure of the difficulty is that one must heat a fusionable gas such as deuterium to 100 million degrees centigrade and confine it by magnetic fields alone.

The program arranged for this occasion starts with an over-all review of the problems involved in controlling fusion reactions and goes on to consider the instabilities that are the greatest present concern. Then there are papers on two of the many diagnostic techniques used to study high-temperature plasmas: microwaves and emitted neutrons. Next, there is a paper on the generation of intense (~200,000 Gauss) magnetic fields such as may be needed to contain plasmas. Finally, there is a forward-looking paper on the use of plasmas for propulsion.

The important role of electronic engineers in fusion research has been pointed out recently. This session is designed to provide IRE members with an introduction to controlled fusion power and some of the research techniques required for understanding the behavior of high-temperature plasmas.

10.1. Controlled Thermonuclear Fusion and Its Meaning for the Radio Engineer

E. W. Herold, *C. Stellarator Associates, Princeton, N. J.*

10.2. Hydromagnetic Instabilities—A Pictorial Approach

I. Bernstein, *Project Matterhorn, Princeton University, Princeton, N. J.*

10.3. Microwave Measurements in Controlled Fusion Research

M. Heald, *Project Matterhorn, Princeton University, Princeton, N. J.*

10.4. Measurements of Neutron Production in a Dynamic Pinch

O. Anderson, *Radiation Lab., University of California, Berkeley, Calif.*

10.5. Production of Intense Magnetic Fields and Their Relation to Fusion Reactors

M. Levine, *Cambridge Research Center, Cambridge, Mass.*

10.6. Plasmas for Propulsion

W. Bostick, *Physics Dept., Stevens Institute of Technology, Hoboken, N. J.*

SESSION 11*

Tues. **10:00 A.M.–12:30 P.M.**

Waldorf-Astoria
Jade Room

BROADCAST TRANSMISSION SYSTEMS

Chairman: G. E. Hagerty, *Westinghouse Broadcasting Co., New York, N. Y.*

11.1. Video Modulation Limiter

L. S. Sadler, *Television Station WMTV, Madison, Wis.*

A simple video modulation limiter is described which can be built at low cost by making use of equipment already in use at most tv stations. Modulation can easily be held at 95 per cent to 100 per cent with inputs varying over a range greater than 20 db. Transient effects are negligible, and the unit is suitable for use with network color. The work of the video operator is greatly simplified while a brighter, more uniform picture is put on the air.

11.2. Color TV Recording on Magnetic Tape

J. L. Grever, *Radio Corp. of America, Camden, N. J.*

It has been established that a practical method of recording monochrome tv signals on magnetic tape is the technique of high speed transverse scanning of a 2-inch-wide tape moving at 15 inches per second. Using this same recording technique for color tv signals presents some interesting and challenging problems. This paper describes some of the precise signal handling techniques and special equipment that had to be developed in order to build a broadcast quality video tape recorder good for both color and monochrome.

11.3. An Automatic TV Level Control Using Vertical Interval Test Signals

J. R. Popkin-Clurman and F. Davidoff, *Telechrome Manufacturing Corp., Amityville, N. Y.*

The introduction of a peak white reference bar in the vertical interval of television signals by three major networks has pointed out the

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importance of this heretofore unused time. The transmission of such information provides all concerned with an easily visible amplitude reference, which may be viewed on either horizontal or vertical oscilloscope presentations. Such information in the signal permits manual adjustments to be made at various control points to assure proper levels throughout the system.

A new device is described, which automatically responds to any amplitude variation in this reference white signal. A servosystem is used to automatically compensate for these variations. Such a device provides alc (automatic level control) which eliminates the necessity for manual adjustment, thereby reducing brightness variations and corresponding distortions.

11.4. Report on Remote Control of a Directive Antenna System

H. E. Rhea, *Triangle Publications, Philadelphia, Pa.*

For a period of several months WFIL experimentally operated a 5-kw AM transmitter and a three tower directional antenna system by remote control. All pertinent meter readings and all operating functions were read and controlled from the remote location. Phase angle readings also were duplicated at the control point. In addition, an automatic meter reading system was field tested. The results of this experimental program will be discussed.

11.5. A Novel System for Feeding a Single Tower AM-FM and TV Signals

A. C. Goodnow, *WBC, Inc., New York, N. Y.*

A novel system of feeding AM, fm, and tv signals to a single tower is described. The tower, a full wavelength long for KYW-AM on 1100 kc, is insulated at its base and sectionalized at the center. The tv antenna is a six-bay super turnstile mounted at the top. The fm antenna is side mounted multi V. Both tv and fm antennas are fed normally with 3½-inch coaxial lines which are specially mounted inside of the tower and brought out of the tower at the bottom and grounded. The tower is fed AM signals above base insulator currents in upper half-wave sections, and lower half-wave sections are controllable for maximum performance conditions.

SESSION 12*

Tues. 10:00 A.M.–12:30 P.M.

Waldorf-Astoria
Sert Room

STEREOPHONIC DISK RECORDINGS

Chairman: H. E. Roys, *RCA Victor Record Div., Indianapolis, Ind.*

12.1. RIAA Engineering Committee Activities with Respect to Stereophonic Disk Records

W. S. Bachman, *Columbia Records, Inc., New York, N. Y.*

Many engineering problems arise with the introduction of stereophonic disk phonograph records. Considerations of the type and depth of modulation, groove shape, compatibility, etc., should be resolved to standards, so that an optimum system will be offered to the public. This paper outlines the efforts of the RIAA Engineering Committee in this direction.

12.2. The Westrex Stereodisk System

C. C. Davis and J. G. Frayne, *Westrex Corp., Hollywood, Calif.*

The paper describes the Westrex Stereodisk recorder, which records two stereophonic channels in a single groove with a single stylus. The axes of the two recordings are at 90 degrees to one another, each being at 45 degrees with the horizontal plane of the record. By using appropriate input circuits, the vertical-lateral type of stereo record also may be recorded. The recorder utilizes the electrodynamic feedback features of the Westrex lateral recorder. Design features of the recorder are described and the performance is discussed. Data on channel cross talk, intermodulation distortion, and frequency characteristics are given.

The design features and performance of the complementary stereodisk reproducer are described, including the desirability of maintaining the same vertical tracking angle in both recorder and reproducer. The reproducer employs dual d'Arsonval movements resulting in exceptionally faithful reproduction.

12.3. Tracing Distortion in Stereophonic Disk Recording

M. S. Corrington and T. Murakami, *RCA Victor Television Div., Camden, N. J.*

Tracing distortion in the 45°–45° and the vertical-lateral systems has been calculated to compare the two systems. Analysis shows that with an ideal pickup there is no cross modulation between the two channels in the 45°–45° system of recording, if the groove angle is 90°. The intermodulation and harmonic distortion within each of the channels in the 45°–45° system is the same as that obtained in the normal vertically-cut record. In the vertical-lateral system there is cross modulation from the lateral channel to the vertical channel and the amount has been calculated. The intermodulation and harmonic distortion in each channel also has been calculated. Curves of the amounts of distortion, cross talk and cross modulation for various groove velocities, stylus radii, and recording velocities are given for both systems.

12.4. Compatibility Problems in Stereophonic Disk Reproduction

B. B. Bauer and R. Snepvangers, *CBS Labs., New York, N. Y.*

The addition of stereophonic information to a record groove introduces a vertical component of groove modulation which generates tracking problems with present day phonograph pickups. A special set of test records was pressed to study the tracking ability of phonograph pickups for vertical modulation. A number of commercial pickups were tested. The results give a measure of the vertical impedance of commercial pickups and provide means for ascertaining the maximum modulation usable in a compatible record.

12.5. Phonograph Pickups for Stereophonic Record Reproduction

W. Bachman, *Columbia Records, Inc., New York, N. Y.*, and B. B. Bauer, *CBS Labs., New York, N. Y.*

The introduction of stereophonic records will require the design of new pickups capable of translating record groove modulation in orthogonal modes. Several schemes are described, together with the advantages and disadvantages of each.

12.6. The Requirements of a Record Changer, Component Parts and Associated Equipment for Stereophonic Record Reproduction

W. Faulkner, *V-M Corp., Benton Harbor, Mich.*

With the advent of stereophonic records, new and more rigid specifications on record changers will be required. The rumble specifications will have to be improved considerably because of the vertical vibration of the changer which will be introduced into the system. On monaural recording, the vertical vibration of the changer had little effect on the cartridge or system. A smaller stylus requiring lower needle pressure necessitates lower horizontal friction and less trip pressure to activate the changer mechanism. Vertical friction likewise must be reduced.

Better cabinet construction and less air coupling between speaker and changer will be required.

SESSION 13*

Tues. 10:00 A.M.–12:00 Noon

Waldorf-Astoria
Grand Ballroom

PLANNING AGAINST TIME

Chairman: C. R. Burrows, *Ford Instrument Co., Long Island City, N. Y.*

The accelerating pace of world events and of scientific progress has put increasing pressure on reducing the time required in translating technical concepts into operational hardware. Proper planning is essential to assure that all the required ingredients are brought together to produce maximum performance at the earliest possible time.

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* Sponsored by the Professional Group on Engineering Management. To be published in Part 10 of the 1958 IRE National Convention Record.

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The speakers will discuss three aspects of this problem, with specific relationship to military weapons, competitive commercial products, and the need for scientists.

13.1. Weapons Systems Development

B. A. Schriever, *Commander, Air Force Ballistic Missile Div., Inglewood, Calif.*

13.2. Commercial Product Development

R. Thalner, *Sylvania Radio and Television Set Div., Batavia, N. Y.*

13.3. Scientific Manpower

H. A. Meyerhoff, *Scientific Manpower Commission, Washington, D. C.*

SESSION 14*

Tues. 10:00 A.M.–12:30 P.M.

New York Coliseum
Morse Hall

AERONAUTICAL AND NAVIGATIONAL ELECTRONICS

Chairman: E. G. Fubini, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

14.1. A VORTAC Traffic Control System

P. E. Ricketts, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

This paper details a traffic control system based on common navigation facilities, operating in conjunction with other available equipments and proven techniques.

Data concerning the identity and position in space of any aircraft are continuously reported and converted into displays for traffic controller analysis. Projected traffic patterns, based on the continuation of current data, also can be displayed at selected intervals of time. Alarms are incorporated to permit manual or automatic correctives for collision or near-miss courses. A pilot display of the traffic environment is provided.

Equipments availability, plus integration, expansible and economic characteristics, suggest its application as an answer to present traffic control problems.

* Sponsored by the Professional Group on Aeronautical and Navigational Electronics. To be published in Part 5 of the 1958 IRE National Convention Record.

14.2. Airborne VORTAC DME for Federal Airways System

S. H. Dodington and B. B. Mahler, *Federal Telecommunication Labs., Nutley, N. J.*

The new VORTAC system to be applied to the Federal Airways provides that distance-measuring service be obtained from a Tacan transponder located at the VOR or ILS site. Existing military airborne Tacan equipment provides Tacan bearing service, which is not needed by all users, and employs some components which are not favored by civil users.

The paper describes a new design of airborne Tacan DME, featuring reduced spurious responses and packaged in accord with civil users' requirements. It is housed in a standard half-ATR chassis and follows the recommendations of the Airlines Electronic Engineering Committee. Provision is made for multiple displays and for the subsequent addition of Tacan bearing service without internal modification.

Its chief features are:

1) ½ ATR size, with no "doghouses" (unless multiple displays are required).
2) No subminiature tubes. ARINC types used wherever applicable.
3) Direct crystal control of 126 transmitting and 126 receiving frequencies.
4) Full military accuracy of ±0.2 mile, independent of distance, up to 195 nautical miles.
5) Provision for easy addition of a ¼ ATR Tacan bearing adaptor.

14.3. IDEA—Integrated Defense Early-Warning Air Traffic Control

B. H. Baldridge and E. V. Hogan, *General Electric Co., Utica, N. Y.*

An integrated system is described which consolidates the function of long and short distance signaling, defense, and air traffic control. Collision avoidance, facilities for malfunctions and nonparticipants, and guidance or control or monitoring are provided for all aircraft. System capacity exceeds the 1975 requirements as determined by the Curtis Committee, even when accommodations for free flight and cruise-climb operations are provided. Phasing, rather than abrupt transition to a new system, is inherent in the system.

The system integrates and performs the required functions from available ground information, supplemented with data such as IAS, unique identity, etc., relayed automatically from the aircraft. Two-way data link is the minimum airborne equipment required for operation in the system, but the use of self-contained navigators, or systems, is permitted. Ground-air data link transmissions are synchronized with the ground surveillance radar to provide coincident reception of radar echo and data link messages for automatic correlation and computer processing. The radar antenna or one with similar pattern will be used for hard-to-jam signaling system contact only while in radar contact. Different queries or commands will be sent with each radar pulse for many exchanges during each radar contact. Data such as aircraft heading, air speed, and unique identity will be obtained automatically and continuously from the aircraft. Courses for navigation, if desired, and headings to avoid ground computer-predicted collisions, will be sent by the signaling system. Integration into ADIZ concepts and military situations appears practical with a common system for military jet transports, light aircraft, and helicopter flights.

14.4. The AN/APN-96 Doppler Radar Set

M. W. McKay, *General Precision Lab., Inc., Pleasantville, N. Y.*

The AN/APN-96 is a completely self-contained and automatic Doppler radar navigation set which provides ground speed and drift angle information of extremely high accuracy. The paper explains the techniques employed in the measurement of ground speed and drift angle. The beam pattern is described and major system parameters are given. In block diagram form, the functions of the receiver-transmitter, the frequency tracker, and the converter are explained. Accuracies of measurement and some specific applications of the equipment are given.

14.5. Increasing the Traffic Capacity of Transponder Systems

H. Davis and M. Setrin, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

Present transponder systems include the Mark X System and the Air Traffic Control Radar Safety Beacon. These systems, or any other single-frequency transponder systems, are severely limited in the density of traffic which they can handle without excessive interference or overload. Garbled reply codes, severely cluttered displays, transponder overload, or reduction in the effective range of the system result from such high density traffic without suitable corrective measures. This paper describes several methods which may be used to alleviate the deleterious effects, analyzes each of them, and presents experimental and field test results showing some rather dramatic conclusions.

SESSION 15*

Tues. 10:00 A.M.–12:30 P.M.

New York Coliseum
Marconi Hall

MEDICAL ELECTRONICS

Chairman: W. E. Trolles, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

15.1. A New Nipkow-Disk Scanner for Accurate Cytological Measurements

H. S. Sawyer and R. C. Bostrom, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

* Sponsored by the Professional Group on Medical Electronics. To be published in Part 9 of the 1958 IRE National Convention Record.

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The mechanical scanner, using a rotating Nipkow disk, has recently found application in the field of medical electronics for accurate microphotometric measurements of cells in cytological smears. In the construction of a Nipkow disk, inaccuracies in the angular position, transmission, and size of the holes in the disk result in measurement errors. Inaccuracies in angular positions of the disk holes cause a variable line-repetition rate. Inaccuracies in hole size and transmission cause variations in the amplitude of the video signal.

In the scanner described here, the disk errors are eliminated electronically by a unique circuit that utilizes a constant light level as a reference source to correct the timing and amplitude of the video signal.

15.2. Electrocardiograph Telemetering (Radio)

J. C. Webb, L. E. Campbell, and J. G. Hartsock, *U. S. Dept. of Agriculture, Beltsville, Md.*

A system of radio-telemetering of electrocardiac voltages of animals using subminiature portable battery-powered equipment is described. The cardiac amplifier and frequency modulated transmitter, including batteries, weighs about 40 ounces. The system makes use of commercial components with limited modifications.

Cardiac signals are collected by electrodes and amplified by a 2-channel subminiature amplifier. The amplified signal modulates a subminiature frequency-modulated transmitter. Conventional receiving and detection equipment are connected to monitoring and recording apparatus.

Used on livestock, the system provided useful monitor of heartbeat frequency at varying levels of environmental noise stress. Ambient sound levels varied from normal to 130 db.

15.3. Electronics in Biochemical Spectroscopy

M. Rogoff, *Federal Telecommunication Labs., Nutley, N. J.*, and T. Gallagher, *Sloan Kettering Institute, New York, N. Y.*

Medical research is being accelerated by a combination of electronics and absorption spectroscopy. A special purpose high-speed computer has been applied to the problem of analysis of complex biochemical mixtures, and the results have shown that this new method can provide more accurate answers in much less time than has been possible by older methods. Multicomponent specimens of steroid mixtures are being analyzed by means of their infrared spectra in minutes, where heretofore days have been required for the same result. The method of computation is a combination of synthesis and waveform matching; the mixture spectrogram is the linear combination of the constituent spectra of the mixture. A set of prerecorded reference spectra are withdrawn from a library, and the computer determines the relative amounts of each to be used in a synthesis of the unknown mixture spectrogram. This process is automatic and the resulting analysis is displayed. This method has been used in the analysis of five component steroid mixtures; the accuracy of the results is equal to or better than those obtained by manual methods. The power of the method lies in the fact that the mixtures themselves can be analyzed without intermediate steps of chemical or physical separation. By eliminating these extra steps, large amounts of time can be saved in the analytical procedure. In addition, the computer displays the results in terms of the quality of match between the synthesized and unknown spectra. This "check" facility is not ordinarily provided by other methods of computation and adds a degree of confidence to the results. This application of electronics to medical and biochemical research promises to accelerate important investigations in this field of analysis.

15.4. Patient Data Systems for Hospitals

G. G. Knickerbocker and G. N. Webb, *Johns Hopkins Hospital, Baltimore, Md.*

The primary purpose of this presentation is to stimulate discussion and thinking on data systems for teaching and general hospitals. Single sample periodic and continuous data collecting techniques are described. Flow plans indicating the phases of transduction, transmission, storage, recall, and evaluation are presented. Methods of data deviation detection are described which will provide warning signals and operation of stand-by recorders.

That equipment which is presently available, as well as systems not yet applied to hospital usage and speculation into what new techniques might be usefully included, are discussed.

Discussion of equipment includes: 1) presently used equipment; 2) currently available units and systems not now specifically applied to hospital usage; and 3) speculation on undeveloped techniques, techniques yet to be developed.

15.5. A New Intracardiac Pressure Measuring System for Infants and Adults

A. Warnick, *Ford Motor Co., Dearborn, Mich.*, and E. H. Drake, M.D., *Henry Ford Hospital, Detroit, Mich.*

The Scientific Laboratory's Intracardiac Pressure Measuring System was developed in cooperation with the Henry Ford Hospital for accurate measurement of cardiac pressures in infants and adults. The pressure sensing element utilizes the strain gauge principle and is of hollow construction, to permit blood sampling at the point pressure is measured and in situ calibration by an external manometer. The pressure transducer is assembled on a No. 6 french catheter and has a maximum outside diameter of 0.090 inch, an effective inside diameter of 0.040 inch, and an over-all length including attachment of 0.65 inch. The natural frequency of the pressure transducer is above 2000 cycles per second when measured in a fluid environment. The transducer is used in conjunction with a low-level dc transistor amplifier (described elsewhere) to raise the output voltage of 3 μv per mm Hg pressure for 9-volt bridge excitation to a level suitable for typical hospital recording equipment. The pressure range of the system is -20 to $+300$ mm Hg; the noise of pickup and preamplifier is equivalent to 1 mm Hg; and the linearity and hysteresis together are better than 2 per cent of range.

Laboratory evaluation followed by clinical use has established the safety, stability, and ease of manipulation of the pressure transducer and catheter assembly.

15.6. The Electronic Evaluation of Fetal Distress

E. H. Hon, M.D., *School of Medicine, Yale University, New Haven, Conn.*

Associated with the birth process in the United States each year there are about 160,000 infant deaths and a large number of infants afflicted with cerebral palsy and mental retardation. These problems are possibly related to poor fetal environment.

The present criteria of fetal distress are based on the average fetal heart rate as determined by periodic sampling in the interval between uterine contractions. These data are therefore open to question.

A method of continuous recording of the instantaneous fetal heart rate during labor and throughout contractions is presented. The total system provides for digital and analog presentation of data, storage on magnetic tape, and semi-automatic data reduction.

SESSION 16*

Tues. 10:00 A.M.–12:30 P.M.

New York Coliseum
Faraday Hall

GENERAL COMMUNICATIONS SYSTEMS

Chairman: R. L. Marks, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

16.1. Digital Communication Systems

R. L. Plouffe, *Federal Telecommunication Labs., Nutley, N. J.*

A communication concept is described which provides for universal communication service. A system, according to this concept, employs digital techniques for all functions required of a communication system; namely, those of looping, switching, and trunking. A communication system is described rather than a telephone system. The distinction is that the terminal equipment may consist of a telephone, a digital data translator, a printer, or any other means of exchanging information at terminals between the terminal and the final receiver, be it a human being or a business machine. This concept is put forth to stimulate ideas in this area rather than to present a development which is just around the corner from a hardware standpoint.

16.2. Constant Amplitude Speech

* Sponsored by the Professional Group on Communications Systems. To be published in Part 8 of the 1958 IRE National Convention Record.

P. J. FERRELL, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

Due to increase in their performance, modern aircraft outfly their communication links. Constant Amplitude Speech increases performance of voice circuits, which will help alleviate this problem. The increase in average power given by constant speech more than compensates for the loss of amplitude information.

16.3. Exploitation of Physical Phenomena for Communications

J. L. RYERSON, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

The extensive development of radio communication has overcrowded existing facilities. This paper proposes solutions to relieve this condition through the use of phenomena other than present day radio for communications. The paper investigates communication through natural ducts by the use of low-frequency radio, sound, light, heat, and gamma rays.

Areas, in which experimental work is required, are summarized. Estimates are made of the probable speed rates with which information may be transmitted by use of the cited phenomena.

16.4. Reduction of Intermodulation in Microwave Systems by Using Ferrite Load Isolators

N. P. WEINHOUSE, *Collins Radio Co., Dallas, Texas*

The results of an experimental study of the long line effect on reflex klystron transmitters of a microwave system are presented. All of the factors that enter into klystron distortion such as tube pulling figure, length of line, and vswr of the load are considered in the determination of system intermodulation. Curves are presented showing system intermodulation as a function of line length for a given amount of load isolation and also as a function of load isolation for a given line length.

The tests were performed using random noise as a simulated speech load. Comparison of results with a recent theory is made.

16.5. The Effects of Pulse Shape and Frequency Separation on FSK Transmission through Fading

G. L. TURIN, *Hughes Research Labs., Culver City, Calif.*

Results are presented which enable the computation, as a function of pulse shape and the frequency separation of the mark and space bands, of the probability of error for binary fsk transmission through a channel disturbed by noise and by nonselective, slow, Rayleigh fading; these results are compared with results for a nonfading channel given by Helstrom. It is assumed that the receiver mark and space filters are matched, respectively, to the transmitted mark and space pulses. Four particular cases are discussed in detail: exponential pulses, Gaussian pulses, rectangular pulses, and pulses with rectangular spectra.

16.6. A 45-Channel PPM System

B. MCADAMS AND S. M. SCHREINER, *Federal Telecommunication Labs., Nutley, N. J.*

A new ppm system is described using simplified circuitry. A 23-channel terminal has only 41 vacuum tubes, occupies 3 feet of rack space. Previous equipment had 123 vacuum tubes and occupied 14 feet of rack space. A novel diode modulator and pwm demodulator reduce the total tube complement to less than two tubes per channel. The system is designed for both commercial and military operation.

Forty-five channel operation is obtained by interleaving the output pulses from two 23-channel multiplexers. At the receiving terminal a high speed flip-flop triggered by the common pulse train is used to control two gate circuits. These gates separate the common pulse train into two 23-channel groups, since each gate passes every other pulse. Guard pulses are provided to maintain proper synchronism of the flip-flop in the event of missing channel pulses.

16.7. New Trends in Directional Communications

R. C. BENOIT, JR. AND F. COUGHLIN, JR., *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

This paper will cover the design and development of electronically steerable wide aperture antenna arrays and associated equipment for directional communications. Emphasis will be placed on equipment functioning in the high-frequency spectrum.

SESSION 17*

Tues. 2:30–5:00 P.M.

Waldorf-Astoria Starlight Roof

CHANGING DEMANDS ON THE BREADTH OF ELECTRICAL ENGINEERING EDUCATION—A PANEL DISCUSSION

Chairman: J. D. RYDER, *Dean of Eng., Michigan State University, East Lansing, Mich.*

Panel Members: S. W. HERWALD, *Westinghouse Electric Corp., Baltimore, Md.*

H. POLLAK, *Bell Telephone Labs., Inc., Murray Hill, N. J.*

D. B. SINCLAIR, *General Radio Co., Cambridge, Mass.*

G. K. TEAL, *Texas Instruments, Inc., Dallas, Tex.*

* Sponsored by the Professional Group on Education. To be published in Part 1C of the 1958 IRE NATIONAL CONVENTION RECORD.

The continuing development of new electronic devices based on an understanding of semiconductor principles and modern physics, the expansion of the electronic engineer into automatic control and systems engineering, and the expansion of communications engineering to embrace statistical systems evaluations, coding and modulation schemes, and computer technology have resulted in explosive increases in the required breadth of electrical engineering education—in other words, the emphasis and coverage given not only to purely electrical topics, but also to related fields such as physics, mathematics, thermodynamics, and materials. In this panel discussion, representatives of industry will attempt to evaluate these changing demands on education and to indicate the influence on education of present trends and probable future developments in electrical engineering.

SESSION 18*

Tues. 2:30–5:00 P.M.

Waldorf-Astoria Astor Gallery

ATOMIC CLOCKS AND MASERS

Chairman: A. G. FOX, *Bell Telephone Labs., Inc., Holmdel, N. J.*

18.1. A Gas Cell "Atomic Clock" Using Optical Pumping and Optical Detection

M. ARDITI, *Federal Telecommunication Labs., Nutley, N. J.*, AND T. R. CARVER, *Princeton University, Princeton, N. J.*

The need for a small size "atomic clock" compatible with airborne operation has led to the development of a gas cell simpler in construction than the atomic beam apparatus. The physical principles involved in the utilization, for a frequency standard, of the hyperfine structure of alkali metal vapor will be outlined. The methods of "optical pumping" and "optical detection" which are utilized to increase the signal-to-noise ratio of the microwave transition will be reviewed. Finally, experimental breadboard models of "atomic clocks" using either sodium or cesium as the alkali element will be described.

18.2. The Atomichron®—An Atomic Frequency Standard—Physical Foundations

A. O. MCCOUBREY, *National Co., Inc., Malden, Mass.*

The physical basis for the use of the radio frequency resonance of Cs^{133} to control frequency will be described in terms of familiar classical engineering concepts. While quantum mechanics is required to give a complete theory of the phenomenon, the classical picture con-

* Sponsored by the Professional Groups on Nuclear Science and Microwave Theory and Techniques. To be published in Part 1 of the 1958 IRE NATIONAL CONVENTION RECORD.

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tains all of the important features and serves to explain the ultra-stability which is an inherent property of the resonance. The cesium atomic beam tube used in the Atomichron, which makes possible the sensing of the cesium frequency response, will be described and the factors affecting the shape of the resonance, its effective Q, and precision will be discussed.

18.3. The Atomichron®—An Atomic Frequency Standard —System Operation and Performance

W. A. MAINBERGER, *National Co., Inc., Malden, Mass.*

The existence of a sealed off atomic beam tube in which cesium exhibits its resonance properties, has led to the development of a frequency producing system which provides an accuracy of 1 part in 10^9 and stability within 5 parts in 10^{10} at all times. The system consists of a beam tube which resonates at 9192.631840 mc, a frequency source which generates an rf field at the resonance frequency, and a control loop for automatic correction of the source. Operation of all circuits and features of the Standard are described. Photographs and recordings of the unit are shown.

18.4. Analysis of The Emissive Phase of a Pulsed Maser

H. H. THEISSING, F. A. DIETER, AND P. J. CAPLAN, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

Combination of the Bloch equations with a power balance for a matched cavity at resonance yields equations which describe processes in which the reaction field is important, such as in the emissive phase of a pulsed maser. These equations contain the same oscillation parameter that appears in the steady-state approximation. Computed plots of rotating moment, total field, spin-emitted power, decay of net inverted spin moment, etc., are shown for different values of the oscillation parameter. It is shown how the behavior of the latter quantity necessitates two different modes of maser operation, the regenerative and the superregenerative. Typical working cycles for both cases are presented.

18.5. A Two-Cavity Unilateral Maser Amplifier

N. SHER, *Philco Corp., Philadelphia, Pa.*

The amplifier described is one of a growing class of devices which uses the phenomenon of induced emission from atomic or molecular particles to obtain coherent oscillation or emission in the microwave region. Following a brief introduction into the principles of inducing coherent radiation from a molecular system, a discussion is presented pertaining to a method for obtaining unilateral amplification by means of a maser employing two isolated resonant cavities with prestimulation. A simplified theory of the amplification process is discussed, outlining the parameters which influence power gain.

The details of an experimental amplifier are described and the results of gain and noise figure measurements are reported.

SESSION 19*

Tues. 2:30–5:00 P.M.

Waldorf-Astoria
Jade Room

BROADCAST TRANSMISSION SYSTEMS AND COMMUNICATIONS SYSTEMS

Chairman: A. L. HAMMERSCHMIDT, *National Broadcasting Co., New York, N. Y.*

19.1. Remote Control of 50-KW Transmitters

R. N. HARMON, *WBC, Inc., New York, N. Y.*

This paper describes planning for remote control of WBZ-AM and KDKA-AM, 50 kw. Examination of performance records of these two stations over a period of years indicates that the transmitter is the most reliable link in the system. Alternate facilities for power, telephone lines, transmitter, and antenna are different for each location; these are described and also how the choice was determined. Remote control experience of 1 kw, WBZA in Springfield, Mass., 100 miles from WBZ-AM, also is discussed.

19.2. Report on Multiplex Experimental Work at WCAU-FM

E. J. MEEHAN, *Station WCAU-FM, Philadelphia, Pa.*

The results of continuing multiplex experimentation by an fm broadcaster in the field of background music are described. Problems relating to distortion, intermodulation between channels, and signal alternation on the multiplex subchannel are reviewed. Practical steps toward the solution of these problems are enumerated. Difficulties encountered with the addition of selective muting to multiplex service and the approach toward a solution are discussed. Direct comparison between main-channel and subchannel service is given. An evaluation is offered of the present status and future promise of multiplex subchannel transmission of background music, based on this experimental work.

19.3. Field Test of Compatible Single Sideband at WABC

R. M. MORRIS, *American Broadcasting Co., New York, N. Y.*

The field test of cssb transmission applied to standard broadcast station WABC is outlined. Problems involved in the interconnection of the cssb exciter with a 50-kw transmitter are treated, together with preliminary tests necessary to assure satisfactory performance. Experiences and conclusions resulting from this test to date are reported.

* Sponsored by the Professional Groups on Broadcast Transmission Systems and Communications Systems. Papers 19.1–19.4 to be published in Part 7 and 19.5–19.7 in Part 8 of the 1958 IRE NATIONAL CONVENTION RECORD.

19.4. Improved CSSB Equipment for the Standard Broadcast Service

L. R. KAHN, *Kahn Research Labs., Freeport, N. Y.*

Design of a new compatible single-sideband adapter for use with conventional broadcast transmitters is described. Measurements of sideband suppression, distortion, signal-to-noise ratio, etc. are given for this improved equipment. Equipment and methods for facilitating operational tests are outlined.

19.5 An Expanded Theory for Signal-to-Noise Performance of FM Systems Carrying Frequency Division Multiplex

D. P. HARRIS, *Collins Radio Co., Dallas, Texas*

The need for a more exact method of calculating snr under weak signal conditions of multichannel fm reception is shown. Demodulated receiver output signal and noise spectra are examined for various levels of carrier-to-noise. Comparison of signal-to-noise thresholds for wide-band fm signals and for frequency division multiplex channels at various frequencies is made. A general method for extending calculated signal-to-noise curves beyond the linear fade terminal point for any channel and any receiver bandwidth is presented. Theoretical effect changing receiver bandwidth on the signal-to-noise characteristic of a given multiplex channel is shown. Design factors limiting actual signal-to-noise performance under various signal conditions are analyzed and compared for different types of fm systems.

19.6. The Generation of Single-Sideband Carrier Telephone Channels by Polyphase Modulation

J. R. MENSCH, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

The vast majority of carrier systems in use today utilize single-sideband suppressed carrier modulation. This type of modulation has generally been produced by balanced modulators and filters to select the desired sideband. In a few cases, single-sideband modulation has been produced by phasing the incoming audio band to cause the unwanted sideband to cancel in a subsequent modulation stage.

A third method of producing single-sideband signals has been described in the literature. This method of modulation has been called "polyphase modulation" by several authors. This paper discusses the theory of polyphase modulation, describes its advantages and limitations for generating the single-sideband signals required for carrier telephone purposes, and presents the characteristics of an experimental channel utilizing polyphase modulation.

It will be shown that the band pass and envelope distortion characteristics of a channel produced by polyphase modulation are as good as or better than those achieved in any other way. This is because the only filtering action in the channel is a low-pass filter which has

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negligible effect on the midband performance of the channel.

In addition, the unwanted sideband lies in the same frequency allocation as the wanted sideband. Thus, even though the suppression may be no better than 25 db, the noise is not disturbing because it is proportional to signal strength and is effectively masked at all times.

19.7. Tele-Map

H. Hoffman, Jr., *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

Tele-Map is a compression code system for transmitting weather maps over teletype facilities with a greatly reduced Time-Bandwidth product as compared to conventional facsimile systems.

SESSION 20*

Tues. **2:30–5:00 P.M.**

Waldorf-Astoria
Sert Room

AUDIO, AMPLIFIER, AND RECEIVER DEVELOPMENTS

Chairman: M. S. Corrington, *Radio Corp. of America, Camden, N. J.*

20.1. Distortion in Audio Phase Inverter and Driver Systems

W. B. Bernard, *Bureau of Ships, Navy Department, Washington, D. C.*

This paper describes an experimental study of some of the most widely used audio phase inverter and driver circuits. The study was directed principally toward the distortion characteristics of these circuits. In making a choice of circuits, the factors of simplicity and the ability to drive the new high power output tubes were considered along with the distortion characteristics. The study shows that, at the usual listening levels, many of the popular circuits give more distortion than the output stages that they drive. In such cases, very simple remedies will reduce this distortion to a fraction of the original amount.

20.2. Latest Advances in Extra Fine Groove Recording

P. C. Goldmark, *CBS Labs., New York, N. Y.*

Immediately after the development of the LP record further investigations have been carried out to put more information on phonograph records. The seven inch $16\frac{2}{3}$ rpm extra fine groove record developed for the automobile phonograph followed next and now an $8\frac{1}{3}$ rpm seven inch record will be discussed which provides four hours of spoken word per record. Some of the underlying theories will be presented and demonstrations will show the results obtained.

20.3. Design of a Transistorized Record-Playback Amplifier for Dictation Machine Application

R. F. Fleming, *The Gray Manufacturing Co., Hartford, Conn.*

This paper outlines the somewhat specialized problems encountered in the design of a transistorized record-playback amplifier for use in a dictation machine. Both Class-A and Class-B amplifiers are compared and discussed in detail and the merits of both are related to the basic requirements relative to power consumption, input and output impedance, gain, stability, and power supply requirements. In order to obtain the desired system performance from record to playback, microphone, reproducer, and recorder head characteristics are carefully analyzed and considerable attention is paid to transformer performance. In order to insure uniformity of product in production, detailed specifications on component performance are established and associated test sets developed. Particular stress is placed on pinpointing transistor parameters.

20.4. Single Tuned Transformers for Transistor Amplifiers

S. H. Colodny, *Philco Corp., Philadelphia, Pa.*

Single tuned circuits have required reexamination since the advent of transistor applications, because the concept of power transfer, rather than voltage transfer, must be used. Match, under the constraint of a specified bandwidth, is defined and the effects of component losses upon the efficiency of the interstage are discussed.

20.5. Design Considerations for Transistorized Automobile Receivers

R. A. Santilli, *Radio Corp. of America, Somerville, N. J.*

The automobile receiver is essentially a high-performance broadcast-band receiver having exceptional agc, sensitivity, selectivity, and noise characteristics. Because these characteristics depend on the control of pertinent transistor parameters, a new line of transitors was developed to meet the stringent performance requirements of this application. This paper discusses the parameters which must be controlled in each of the stages of a typical automobile receiver and describes the characteristics of transistors developed specifically for this service. A high-performance all-transistor receiver using these new transistors is described, and performance data are given.

20.6. Voltage Sensitivity of Local Oscillators

W. Y. Pan, *Radio Corp. of America, Camden, N. J.*

The theory of voltage sensitivity of local oscillators is discussed. Based on this theory, the frequency characteristics under the conditions of oscillator warmup and varying operating voltages can be predicted and controlled.

It has been found that local oscillators which are stabilized for frequency deviations during the warmup period remain stabilized for variations of operating voltages, provided that certain additional requirements are fulfilled. Such additional requirements are discussed and illustrated with a sample oscillator circuit that may be used in television receivers.

Particular considerations pertaining to oscillator circuits and devices are proposed, thereby a nearly perfect oscillator stability may be attained.

* Sponsored by the Professional Groups on Audio and Broadcast and Television Receivers. To be published in Part 7 of the 1958 IRE National Convention Record.

SESSION 21*

Tues. **2:30–5:00 P.M.**

New York Coliseum
Morse Hall

BEAM AND DISPLAY TUBES

Chairman: B. Salzberg, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

21.1. High Transconductance Wide-Band Television Gun

E. Atti, *Westinghouse Electric Corp., Elmira, N. Y.*

A television gun possessing a transconductance about 30 times larger than the transconductance of conventional guns having comparable electron optical performance is described. The improved result is achieved by substantially divorcing the drive performance from the electron-optical performance. The new gun is of the conventional screen grid type, differing in its simplest version by a single additional electrode. It has a multipurpose screen grid and makes more efficient use of cathode area. The low shunt capacitance makes the gun particularly attractive for wide-band operation. Simple design changes permit conversion of most screen grid guns into their high transconductance wide-band counterpart.

21.2. The Annular Geometry Electron Gun: A New Electron Device

J. W. Schwartz, *RCA Labs., Princeton, N. J.*

The annular geometry gun represents a distinct departure in electron gun design and operation. The modulator section contains an annular cathode, annular control grids and accelerating grids, a beam bending probe, and an electron object electrode. Very high modulation sensitivity, inverted modulation characteristics, internal electronic video signal amplification, and automatic "white noise" inversion are among the unique performance features of the annular cathode gun. Beam control is produced by a focus modulation process. The final spot is formed by imaging a geometrical aperture in the object plate. This results in very high resolution capabilities and an optimum focus condition and spot size which are practically independent of beam current.

* Sponsored by the Professional Group on Electron Devices. To be published in Part 3 of the 1958 IRE National Convention Record.

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21.3. Recent Developments in Shaped Beam Display and Recording Techniques

R. M. PETERSON AND R. C. RITCHART, *Stromberg-Carlson Co., San Diego, Calif.*

Recent developments in the electron optics of shaped beam tubes and improvements in structural design are discussed. A new series of cathode ray tubes and their application to three new types of computer output devices are described. The performance capabilities of these indicate a break-through into a new region of speed and flexibility essentially unlimited by mechanical inertial effects. A time shared analog-alphanumeric-symbolic display useful as a man-machine link in real time situations is described and system design equations are derived to enable the designer to select optimum system parameters.

21.4. ELF, A New Electroluminescent Display

E. A. SACK, *Westinghouse Electric Corp., Pittsburgh, Pa.*

The ELF screen is a new, flat electroluminescent display which combines very desirable brightness and halftone characteristics with highly flexible storage capabilities. Ferroelectrics are used with the electroluminescent layer to provide built-in storage and control. The output image is formed in accordance with a charge pattern on the ferroelectric control array. This pattern is established by a signal distribution system, part of which is incorporated in the screen structure.

Highlight brightnesses in excess of 100 foot-lamberts have been obtained with contrast ratios of 200 to 1. Images may be stored for several minutes or, alternatively, the images can be modified many times per second. There are no set restrictions on the size of the screen.

Several low resolution screens have been built to test the ELF principle. Work is in progress toward development of high resolution models. The various elements of a signal distribution system have been tested and found to perform satisfactorily.

21.5. A Tube That Tells Time

W. T. ERIKSEN AND E. J. HANDLY, *Raytheon Manufacturing Co., Newton, Mass.*

A novel subminiature tube type for measuring the total number of hours of operation of any electrical equipment or component is described. The tube is capable of measuring operating times ranging from 200 to 20,000 hours by adjustment of the circuit conditions. Operating characteristics of the device are detailed including the effects of several factors such as temperature, shock, and vibration.

The results are particularly applicable where the tube is used to obtain reliability data so that lifetimes of components and modules can be predicted and replacements made before equipment failure occurs.

SESSION 22*

Tues. **2:30–5:00 P.M.**

New York Coliseum
Marconi Hall

BIOLOGICAL TRANSDUCERS—PANEL DISCUSSION

Chairman: O. H. SCHMITT, *Dept. of Physics, University of Minnesota, Minneapolis, Minn.*

Animals communicate with their environment and process this control information with the aid of several highly specialized and remarkably effective biological transducer systems. Some of these transducer systems utilize cellular geometry and topology, channel diversity, and unusual nonlinear pulse-codings to simply achieve results much sought after in engineering systems and usually achieved only with very complex systems of low functional safety factor.

A group of experts representing biological, biophysical, and engineering approaches will describe some of these transducer systems giving specific examples of codings, transfer functions, and circuit arrangements, and they will attempt to interpret these research facts in engineering context.

* Sponsored by the Professional Group on Medical Electronics. To be published in Part 9 of the 1958 IRE NATIONAL CONVENTION RECORD.

SESSION 23*

Tues. **2:30–5:00 P.M.**

New York Coliseum
Faraday Hall

RELIABILITY THROUGH COMPONENTS

Chairman: L. PODOLSKY, *Sprague Electric Co., North Adams, Mass.*

23.1. Reliability of Missile Guidance Systems—The Statistical Cliche vs Reality

A. R. GRAY, *The Martin Co., Orlando, Fla.*

It will be pointed out that reference to the "Statistical Cliche" in the subtitle casts no reflection on the several excellent treatises that have been written on the statistical analysis of component-part failure, component-part reliability, complexity, and weapon-system reliability. The paper will stress, however, that the time for talking is now past, that the "statistical cliche" must now be reinforced by a hard, cold look at "reality."

Remembering that there has been a pendulum swing during the last three years, from a relatively uncontrolled reliability to a type of reliability effort using "too many Chiefs and too few Indians," the speaker will warn that guided missile managers should now make a critical and objective re-examination of our entire reliability philosophy.

In the body of the paper, an attempt will be made to give an impartial evaluation of numerous "classical" and contemporary reliability theories and practices. Treated at some length will be the "weighting of component-part reliability by categories," and "practical application." It will be emphasized that the future paths to high reliability lie in the direction of "component-part accelerated life testing," and some definite proposals will be made along these lines.

In conclusion, long-range reliability trends, dictated by the economics of the situation, will be discussed. The speaker will prognosticate on future roles of reliability engineering.

* Sponsored by the Professional Group on Reliability and Quality Control. To be published in Part 6 of the 1958 IRE NATIONAL CONVENTION RECORD.

23.2. Component Part Failure Rate Analysis for Prediction of Equipment Reliability

R. L. VANDER HAMM, *Collins Radio Co., Cedar Rapids, Iowa*

Electronic equipment contractors are now required to meet specific reliability specifications. To evaluate these requirements, failure rates of component parts must be known. Published failure rates are highly inconsistent.

This paper presents component part failure rates derived from failures occurring during 1,010,000 observed equipment operating hours, accumulated through 20-hour performance tests on 50,000 production equipment. Accuracy of derived data is verified by correlation with field experience.

Component part application derating to achieve desired reliability and to minimize failure rates is discussed.

23.3. A Progress Report on the Arma Inertial Guidance System Reliability Program

E. F. DERTINGER, *American Bosch Arma Corp., Garden City, N. Y.*

The proposed paper will describe the procedures established and steps taken to render a highly reliable inertial guidance system. The Arma reliability program, as described at the 1957 Convention, represented an all-out attempt to achieve high reliability in the first preproduction equipments, and then to maintain this inherent reliability as the production program continued. A progress report will be presented on the early achievements of this program, giving test results in order that engineers in other development programs might take note of the problems encountered and the shortcomings to be expected of component parts. As a result of the test data collected, the Arma inertial guidance system has been made more reliable. We feel this data might well be shared with others engaged in the development of military equipment.

23.4. An Impulse Test for Evaluating the Vibrational Characteristics of Receiving Tubes over a Wide Frequency Range

S. A. JOLLY AND W. U. SHIPLEY, *General Electric Co., Owensboro, Ky.*

Test equipment has been developed which evaluates the vibrational characteristics of

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vacuum tubes by measuring their electrical output when subjected to a mechanical impulse. In this equipment the impulse is provided by a pendulum-type tapper. Plate-circuit voltage variations in the form of a decaying wavetrain are produced. Both the peak and the integrated values of this wavetrain are measured.

The results obtained using the impulse test are compared with those obtained using a periodic and a random vibration test. Analysis of the data indicates that the impulse test can economically provide essentially the same information as the other two tests.

23.5. Reliability of Power Amplifier Klystrons in Troposcatter Communication Systems

R. F. Lazzarini and H. A. Bailey, *Eitel-McCullough, Inc., San Bruno, Calif.*

Power amplifier klystron reliability, based on performance in a number of troposcatter communication systems, is analyzed. Actual system performance is compared to published life expectancy probability curves.

The major cause of short life or catastrophic failure is discussed.

Methods for improving reliability and short life or catastrophic failure are presented. The basic discussion is expanded to include fundamental amplifier klystron operational requirements.

Improved training programs and communication between tube manufacturers, systems designers, and systems operators is stressed as a means by which klystron reliability can be increased.

SESSION 24*

Tues. 8:00–10:30 P.M.

Waldorf-Astoria Starlight Roof

ELECTRONICS IN SPACE—A PANEL DISCUSSION

Chairman: L. DuBridge, *Pres., California Institute of Technology, Pasadena, Calif.*

A Prelude to Space Travel

Preparation has begun and major new phases of evolution are under way as a consequence. The round table of seven outstanding scientists will discuss informally the major problems to be encountered, including the use of electronics for propulsion, navigation, communications, telemetering, and instrumentation. What new areas must be anticipated for existence en route and in the terminal environment?

24.1 Propulsion and Interplanetary Travel

E. Stuhlinger, *U. S. Army Ballistic Missile Agency, Huntsville, Ala.*, and K. A. Ehricke, *Convair Astronautics Div., San Diego, Calif.*

* Sponsored by the Professional Groups on Aeronautical and Navigational Electronics, Medical Electronics, and Telemetry and Remote Control.

24.2. Navigation and Control

C. S. Draper, *Instrumentation Lab., Massachusetts Institute of Technology, Cambridge, Mass.*

24.3. Man in the Space Environment

D. G. Simons, *Holloman Air Force Base, N. M.*

24.4. Communications and Telemetering

J. B. Wiesner, *Research Lab. for Electronics, Massachusetts Institute of Technology, Cambridge, Mass.*

24.5. Terminal Environment

F. L. Whipple, *Smithsonian Astrophysical Observatory, Cambridge, Mass.*

SESSION 25*

Tues. 8:00–10:30 P.M.

New York Coliseum Faraday Hall

ELECTRONICS SYSTEMS IN INDUSTRY—A PANEL DISCUSSION

Chairman: J. D. Ryder, *Dean, College of Eng., Michigan State University, East Lansing, Mich.*

Panel Members: J. M. Bridges, *Office of the Asst. Secretary of Defense, Washington, D.C.*

C. C. Hurd, *IBM Corp., New York, N. Y.*

T. R. Jones, *Daystrom, Inc., Murray Hill, N. J.*

J. D. Ryder, *Michigan State University, East Lansing, Mich.*

The great impact which electronics had on American industry will be highlighted at this panel symposium.

J. D. Ryder will open the symposium with a paper on "New Trends in Engineering Education." The emphasis in Dean Ryder's talk will be on strengthening the requirements in fundamental sciences, without which neither the demands of industry nor those of our defense establishments can be satisfied.

C. C. Hurd will discuss new ideas which found their entry in industry in connection with fully automatic processes.

* Sponsored by the Professional Group on Industrial Electronics. To be published in Part 6 of the 1958 IRE National Convention Record.

T. R. Jones will speak about organization of complete electronic systems, utilizing the resources of several integrated organizations.

J. M. Bridges will highlight the military aspects associated with electronic systems engineering and their relationship to the electronic engineering professional society.

SESSION 26*

Wed. 10:00 A.M.–12:30 P.M.

Waldorf-Astoria Starlight Roof

AERONAUTICAL AND NAVIGATIONAL ELECTRONICS

Chairman: J. F. Morrison, *Bell Telephone Labs., Inc., Whippany, N. J.*

26.1. Airborne Dual Antenna System for Aerial Navigation

W. M. Spanos and J. M. Ashbrook, *Federal Telecommunication Labs., Nutley, N. J.*

The requirements for obtaining omnidirectional pattern coverage for the various attitudes of aircraft during flight impose serious limitations on the use of ultra-high frequencies for the aerial navigation systems. The large size of many aircraft limits the usefulness of a single antenna in providing operating signals under all encountered flight conditions. Some improvement has been realized through use of multiple antennas and associated switching arrangements. However, these arrangements may be expensive and unnecessary when equivalent or better performance can be obtained through passive antenna arrangements. A dual antenna system using parallel-driven sector antennas satisfies these requirements. Flight tests with an experimental dual antenna system indicated improved performance for DME and Radar Safety Beacon Systems. A flyable-model, dual antenna system for DC-6 and Constellation type aircraft has been developed. A prototype of this antenna system on a DC-3 has been undergoing operational in-flight evaluation using the Tacan navigation system. Preliminary evaluation of the dual antenna system has indicated improved performance over that provided by a conventional belly antenna. Provision of a passive rf hybrid in the system permits the simultaneous operation of Radar Safety Beacon equipment from the same antenna system when so desired.

26.2. Engineering Evaluation of an Automatic Ground Controlled Approach System (AN/MSN-3)

R. M. Brooks and W. F. Hoy, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

* Sponsored by the Professional Group on Aeronautical and Navigational Electronics. To be published in Part 5 of the 1958 IRE National Convention Record.

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This paper presents the results of flight evaluations performed on the final configuration of the Automatic Ground Controlled Approach System AN/MSN-3.

Comparative analysis of data collected on two types of control philosophies designed into this system proved that ground computed position displacement and air derived flight path are preferred to the method where the optimum flight path is computed on the ground and translated to bank magnitudes in the aircraft.

The evaluation program culminated in an approach system with response characteristics satisfactory for controlling varying types of aircraft through IFR minimum, while maintaining position and heading attitudes for transition to manual landing procedures.

26.3. A Quantitative Analysis of Automatic Target Detection—Position Estimation Schemes Observing Scintillating Targets in Noise

C. M. Walter, *Air Force Cambridge Research Center, L. G. Hanscom Field, Bedford, Mass.*

An analytical technique is presented which makes possible the quantitative evaluation of a wide class of automatic target detection and position estimation schemes operating on data from scintillating targets imbedded in receiver noise and in clutter. The effect of target scintillation in introducing intrinsic limitations on the detection efficiency and azimuth accuracy of search radar surveillance systems is discussed in detail. The optimization of such systems parameters as detection efficiency and azimuth accuracy, for scintillating targets in noise, also is treated. Design data and numerous graphs are provided covering a variety of systems configurations.

26.4. Applying the Amplitron and Stabilotron to MTI Radar Systems

T. A. Weil, *Raytheon Manufacturing Co., Wayland, Mass.*

The Amplitron is a basically new microwave amplifier offering many advantages as an rf power amplifier. The Stabilotron is a stabilized oscillator using the same basic structure and offering performance similar but superior to that of magnetrons. The stability requirements on the Amplitron and Stabilotron for use in MTI systems are established by analysis of the methods by which such systems extract MTI information. Tolerable instabilities are calculated, and the modulation sensitivities of the Amplitron and Stabilotron are given and compared with other microwave tube types. Several Amplitron and Stabilotron systems are analyzed and the results of measurements are given.

26.5. Transistorized Airborne Frequency Standard

G. R. Hykes, *Collins Radio Co., Cedar Rapids, Iowa*

Present day single-sideband techniques require high stability oscillators, both on the ground and in the air. Weight, size, and power demands are easily met for ground equipment. For airborne use, many new problems are created. A modular unit of 60 cubic inches and weighing 2 pounds has been developed, to maintain a minimum long-term stability of 1 part per million per month under all environmental conditions of existing aircraft systems. Cost, reliability, and maintenance problems are considered.

SESSION 27*

Wed. 10:00 A.M.–12:30 P.M.

Waldorf-Astoria
Astor Gallery

STATISTICAL APPLICATIONS

Chairman: R. M. Fano, *Dept. Elec. Eng., Massachusetts Institute of Technology, Cambridge, Mass.*

27.1. Frequency-Domain Statistical Model of Linear Variable Networks for Finite Operating Time

G. W. Johnson, *IBM Airborne Computer Labs., Owego, N. Y.*

This paper extends the conventional frequency domain techniques to include an important category of the general nonstationary field wherein a random input, whose statistical properties are stationary for $t \geq 0$, is suddenly applied at $t=0$ to a linear variable network (initially quiescent at $t=0$). A system design based on minimizing the expected value of the squared error after any finite operating time is assumed to be of interest. It is shown that the transient statistical response may be obtained by frequency domain methods if the linear network under consideration is replaced by a modified variable network transfer function which includes a starting switch closing at $t=0$. The frequency-domain transient statistical model reduces to the conventional steady-state model if the linear network under consideration is invariant and the operating time is large compared to the network settling time.

27.2. The Root Square Locus Plot—A Geometrical Method for Synthesizing Optimum Servosystems

S. S. L. Chang, *New York University, New York, N. Y.*

This paper presents a simple geometrical method for synthesizing optimum servosystems with known or stationary random inputs and possibly in the presence of noise and disturbance, with heat loss or saturation tendencies of the fixed components taken into account. Starting from poles and zeros of the fixed components only, without specifying any compensating means, the new method automatically gives the optimum compensating network for each value of gain. It is based on the analytical result (shown in the paper) that the closed loop poles of the optimum system are related to the poles and zeros of the fixed components by the root square locus plot.

* Sponsored by the Professional Groups on Automatic Control and Information Theory. To be published in Part 4 of the 1958 IRE National Convention Record.

27.3. TV Bandwidth Reduction by Digital Coding

W. F. Schreiber and C. F. Knapp, *Technicolor Corp., Burbank, Calif.*

A device has been constructed which uses statistical redundancies in video signals to reduce transmission bandwidth. Using a variant of run-length coding, it creates and stores a code group at each brightness change or run end. The code groups, which indicate the run-end position, are then transmitted at a uniform rate and restored at the receiver. The original signal timing is recreated by reading a code group out of storage and comparing it with the output of a local code group generator until identity is indicated. Brightness information is handled in a parallel storage-transmission channel.

27.4. Subjective Experiments in Visual Communication

R. E. Graham, *Bell Telephone Labs., Inc., Murray Hill, N. J.*

Efficient coding of "natural" information sources, such as speech or pictures, will require much new knowledge of the perceptual properties of the human receiver. Ideally in the case of visual communication we wish to determine a *minimal* set of transmitted pictures which, to a specified degree of observer satisfaction, will approximate any desired picture sequence. Some progress toward this goal is being made by the construction or simulation of proposed coding schemes and subjective evaluation of the resulting pictures. Basic studies of the observer also are needed to understand the processes of visual perception, recognition, and memory. Subjective experiments are currently being performed to explore the observer's ability to recognize and discriminate patterns in both still and motion pictures, and to obtain his reaction to various rates of pictorial information.

27.5. Demonstration of Some Visual Effects of Using Frame Storage in Television Transmission

M. W. Baldwin, Jr., *Bell Telephone Labs., Inc., Murray Hill, N. J.*

An illusion of motion can be maintained at frame rates considerably lower than 30 per second. How much lower depends upon the subject matter and upon the particular way of grouping and repeating stored frames to avoid flicker.

A specially printed motion picture film is used to demonstrate the visual effects of transmission at 12, 8, 6, and 4 frames per second with three different methods of grouping and repeating the stored frames.

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SESSION 28*

Wed. 10:00 A.M.–12:30 P.M.

Waldorf-Astoria
Jade Room

ELECTRONIC COMPONENT PARTS

Chairman: R. J. FRAMME, *Wright Air Development Center, Wright-Patterson Air Force Base, Dayton, Ohio*

28.1. Development of Electronic Components for the Nuclear Radiation Environment

J. W. CLARK, *Hughes Aircraft Co., Culver City, Calif.*

Nuclear radiation environments will become increasingly common in future applications of military electronics. This will be a summary talk, condensing present knowledge concerning radiation effects upon electronics. Electronic components will be ranked in approximate order of radiation vulnerability. The nature of effects of radiation upon the electrical and mechanical properties of components will be indicated insofar as this is known. Both rate effects which are reversible and depend only upon radiation rate, and dose effects which are irreversible and depend upon accumulated radiation dose, will be discussed. The former may produce serious circuit malfunctions without component damage; the latter may produce complete equipment failure.

An orderly program for development of components having predictable performance in the radiation environment will be briefly described.

28.2. Design of Shielded Air-Cored Inductors

R. O. SCHILDKNECHT, *Federal Telecommunication Labs., Nutley, N. J.*

The application of transmission-line theory to shielded single layer solenoidal structures has resulted in a new approach to the design of air-cored miniaturized inductors at frequencies of two to one thousand megacycles. Simple proportionalities are given which predict the Q, volume, and other parameters of optimum design inductors to within better than ten per cent. The Q has been optimized on the basis of total shielded volume in a realistic approach to miniaturization. The data apply to the entire useful frequency range of the particular inductor from quarter wave resonance as a helican resonator down to the point in which the unit behaves as a lumped inductance.

28.3. Ceramic Coating Applications in the Electrical Field

P. HUPPERT, *Gulton Industries, Inc., Metuchen, N. J.*

Ceramic coatings as a replacement of organic materials of various kinds are becoming increasingly important in meeting requirements for higher operating temperature ranges.

A wide variety of metals such as copper aluminum, ferrous alloys, and silver are to be coated.. Both insulating and conductive, as well as semiconductive coatings in combination with heat corrosion and other specialized properties, have to be developed to protect wires, cables, motor housings, etc., to serve as cements between motor and transformer laminates, and to be used as potting compounds. A wide variety of approaches is offered.

28.4. The Components Engineer and the Sales Engineer, Partners in Reliability

P. C. KNOX, *Lutherville, Md.*

It is essential that the components engineer, within a company, and the sales engineer, representing a components manufacturing company, realize their joint responsibilities concerning the performance and reliability of completed equipment.

The components engineer is responsible for the establishment of procurement specifications and the determination of approved manufacturers. He assures that all requirements, electrical, mechanical, and environmnetal, are clearly and completely stated to the mutual agreement of himself, concerned design engineers, and the component manufacturer.

The sales engineer is responsible for the presentation of complete, accurate, and up-to-date component product technical information for use and constant reference by the components engineer and design engineers. He gives prompt and thorough attention to all questions. He regards the components engineer as his supervisor during technical negotiations.

A constant contact and exchange of technical information between the components engineer and the sales engineer is of great assistance to design engincers and greatly contributes to the performance and reliability of completed equipment.

28.5. Miniature Ruggedized Precision Meters

J. F. FAUGHNAN AND R. E. LOISELLE, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

The impact of military requirements for transistor and miniaturized equipment motivated the development of miniature indicating instruments featuring exceptional accuracy, sensitivity, readability, and performance under tactical conditions. As a result of development work with the Hickok Electrical Instrument Company, a meter having many characteristics heretofore unavailable for commercial as well as military applications has been made available. Accuracy of ± 1 per cent of full scale value, 180° pointer deflection, 1 inch barrel diameter, sensitivity of 50 μa, knife edged pointer, and anti-parallax, mirrored dial, are some of the unique features of this meter. Achievement of these characteristics was obtained by a novel magnetic circuit design, and the utilization of optimum jewel and pivot combination. The basic simplicity of the developed meter design as compared to conventional long scale meters offers significantly reduced production costs, thus making readily available high precision miniature meters for use in critical applications.

* Sponsored by the Professional Group on Component Parts. To be published in Part 6 of the 1958 IRE NATIONAL CONVENTION RECORD.

SESSION 29*

Wed. 10:00 A.M.–12:30 P.M.

Waldorf-Astoria
Sert Room

CIRCUIT THEORY I AND ULTRASONICS I: SYMPOSIUM ON MODERN ASPECTS OF DELAY LINES

Chairman: H. A. WHEELER, *Wheeler Labs., Inc., Great Neck, N. Y.*

29.1. Low-Dispersion Wired Delay Lines

M. J. DITORO, *Polytechnic Research & Development Co., Inc., Brooklyn, N. Y.*

A survey is given reviewing the historical development and indicating the present status of wired electrical delay lines. Factors considered include reduction of impulse ripple or dispersion, attenuation, bandwidth-delay product, and physical size. Cascaded lumped, distributed solenoidal and hybrid combined configuration of delay lines are considered. The various methods are reviewed which have been suggested for avoiding the excessive dispersion arising especially from excessive group delay distortion. These methods comprise use of bridged lumped capacitors or distributed capacitive patches, and the circuit and geometrical control of the series mutual inductance of the delay line. The performance obtained with these remedies is compared with that of the ideal delay line having known pole and zero locations in the p plane.

29.2. Electrical Design of the Transducer Networks of a Magnetostrictive Delay Line

L. ROSENBERG AND A. ROTHBART, *Federal Telecommunication Labs., Nutley, N. J.*

Certain basic concepts are presented as a guide in the electrical design of the transducer networks of an ideal magnetostrictive delay line for digital or analog pulse transmission.

For single-coil digital operation, the waveform of input current pulse to the transmitter coil, which offers the best compromise between economy of driving power and acceptable signal-to-noise ratio, is a trapezoid. Design techniques and transient response curves are submitted for transducer networks which can pass a trapezoidal pulse of a desired rise time. Basic relationships are established for the peak amplitudes of the input electrical pulse, the transmitted mechanical pulse, and the output electrical pulse.

In order to obtain increased efficiency for digital operation, it is possible to design multiple-coil transducer networks which have the properties of a directional coupler.

* Sponsored by the Professional Groups on Circuit Theory and Ultrasonics Engineering. To be published in Part 2 of the 1958 IRE NATIONAL CONVENTION RECORD.

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For analog pulse transmission requiring a delay line with a linear transfer response, the pertinent theoretical considerations are derived.

29.3. The Approximation Problem in Lumped Delay Line

A. PAPOULIS, *Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

The design of a lumped delay line can be approached in one of two ways.

1) Approximate ϵ^{-pt_0} by a rational function of a given order (frequency domain approximation).
2) Approximate the delayed impulse $\delta(t-t_0)$ by a sum of n exponentials (time-domain approximation).

Most of the available techniques use the first approach. In this paper, the known methods will be discussed briefly, and the solution of the time-domain problem will be attempted. A partial answer will be given for the RC line in which the approximating function $f(t)$ equals the sum of n real exponentials. It will be shown that the optimum response with equal ripple, minimizing the "rise-time" is given by

$$f(t) = C_n\left[a\left(2\epsilon\frac{-t\ln 2}{t_0}\right)^2 + b\right],$$

where C_n is the Tchebycheff polynomial and a and b suitably chosen constants.

29.4. Coiled Wire Torsional Wave Delay Line

R. N. THURSTON AND L. M. TORNILLO, *Bell Telephone Labs., Inc., Murray Hill, N. J.*

This paper describes a torsional wave ultrasonic delay line in which the delaying medium is a wire 15 feet long coiled up into a spiral having an outside diameter of 4 inches. The bandwidth is limited by the transducers used to generate and receive the torsional waves. In this case, the transducers were made of barium titanate ceramic. The minimum insertion loss for the device, measured between 3000-ohm resistors, is 9 db at 840 kc; the 3-db bandwidth is 98 kc. The measured delay time is 1.65 milliseconds. Of the 9-db insertion loss, approximately 6 db is attributed to attenuation in the wire. The remaining loss is due to the two transducers and the associated tuning inductors.

29.5. Variable Delay Line Using Ultrasonic Surface Waves

J. D. ROSS, S. J. KAPUSCIENSKI, AND K. B. DANIELS, *E. I. du Pont de Nemours & Co., Aiken, S. C.*

A delay line is described that uses ultrasonic surface waves for the delayed signal. Surface waves are directed along a helical path on a cylinder and are detected at the far end, or are reflected to the transmitting transducer. The time delay is determined by the velocity of the surface wave on the cylinder and the path length between the transducers. The time delay can be continuously varied from several microseconds to several milliseconds by sliding the receiving transducer along the helix.

SESSION 30*

Wed. 10:00 A.M.–12:00 NOON

Waldorf-Astoria Grand Ballroom

THE CANADIAN AUTOMATION SYSTEM OF POSTAL OPERATIONS

Chairman: G. S. COBURN, RCAF, *c/o Radio Corp. of America, Camden, N. J.*

30.1. The Canadian Automation System of Postal Operations

M. LEVY, *Canada Post Office Dept., Ottawa, Ont., Can.*

This paper gives an original and detailed survey of the Canadian method of automation of postal operations and describes the first world's installation.

It is shown that, while many alternative methods of automation are possible, it is the most unexpected one which results in the maximum flexibility, reliability, and lower cost. This method, which we call the method of successive operations, applies to all sizes of installations and solves the electronic computer problem in an easy way.

However, a series of difficulties had to be solved before the method became practical. These difficulties have been solved by electronic means (low-noise flying spot scanner, double detector, error checking, code correcting devices, etc.). The result has been a system in which less than one letter out of 10,000 is misrouted.

Finally, a description is given of the first electronic installation (Ottawa Post Office) and designs for large sized installations are detailed.

30.2. Coding and Error Checking in the Canadian System

M. LEVY AND V. CZORNY, *Canada Post Office Dept., Ottawa, Ont., Can.*

The problem of misrouting letters is extremely important in electronic mail sortation. It is shown how, by a detailed code analysis, together with a special error checking device, misrouting has been reduced to the lowest minimum possible. Taking account of all factors, probably less than 1 letter out of 10,000 will be misrouted.

This problem is closely interconnected with that of coding. As coding will become international in a not too far future, an attempt has been made to establish some additional rules which help in the selection of an optimum code and to analyze the advantages and disadvantages of various possible coding keyboards.

The study is completed by analysis of some speed and training tests showing that coding could be learned very quickly.

Finally, description is given of some error checking devices used in the Ottawa installation.

30.3. Organization of the Electronic Computer for the Canadian Electronic Mail Sorting System

A. BARASZCZEWSKI, *Canada Post Office Dept., Ottawa, Ont., Can.*

In the Canadian Electronic Mail Sorting System the addresses on letters are converted into special codes printed on the reverse side of the envelopes. Thereafter, the coded mail is sorted automatically on a conveyor system controlled by an electronic computer.

The aim of this paper is to analyze the organization of an electronic computer for the use of post offices. The functional requirements of this computer are governed by the following main factors: 1) post office requirements; 2) code and coding system used, in particular stressing the difficulties arising in city mail sortation; 3) characteristics of mechanical conveyor; 4) memory size and differences in computer design for use in small, medium, and large post offices.

From these considerations, the organization for an electronic computer for Canadian Post Offices' use is fully described; the circuit details are not given.

30.4. Facing, Code Reading and Electrostatic Printing

H. JENSEN AND K. H. ULLYATT, *Canada Post Office Dept., Ottawa, Ont., Can.*

This paper is divided in two sections. The first deals with the problem of facing letters and the second with an electrostatic printer designed to print code marks on the back of envelopes.

The problems of preselection, and facing operations in the Canadian studies of automation of mail operations are described.

Various types of preselecting methods are possible. They have been analyzed and a method impregnating the stamps with fluorescence has been selected. A detailed analysis is given.

In the facing process letters have been selected for size and their face position in the conveyor. A comparatively simple method of achieving this result is described.

In connection with automatizing mail handling operations in the Canadian Post Office, an essential step is a coding of the address and the printing of this coded information on the back of the envelope.

An electronic printer is described which is used in conjunction with a special coding keyboard. The principle of operation is an electrostatic process whereby a fluorescent powder is attracted to selected areas on the surface of a drum to correspond with particular keys depressed by an operator, followed by transfer of the powder pattern to the envelope and subsequent "fixing" of the pattern.

The machine is essentially an offset rotary printer with an electrically active, though stationary, type matrix set in the drum surface.

* Sponsored by the Professional Groups on Engineering Management and Industrial Electronics. To be published in Part 6 of the 1958 IRE NATIONAL CONVENTION RECORD.

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SESSION 31*

Wed. 10:00 A.M.–12:30 P.M.

New York Coliseum
Morse Hall

RADAR IN MILITARY ELECTRONICS

Chairman: R. SARGENT, *Office of Asst. Secretary of Defense, Washington, D. C.*

31.1. Automatic and Continuous Radar Performance Monitor

W. C. WOODS, *Sperry Gyroscope Co., Div. of Sperry Rand Corp., Great Neck, N. Y.*

A radar performance monitor is described which continuously measures two radar system parameters, noise figure, and transmitter energy-per-pulse. A relative tuning indicator continuously monitors the IF frequency. The vswr in the transmission line can be accurately measured by a simple manual adjustment. (Measurements are continuously made while the radar is operating.) The performance monitor was developed to be fully automatic, compatible with different radar systems, operate reliably, and "fail-safe," and be designed with miniature components. A self-checking circuit and switch permits a checkout of the monitor component circuits to discover and localize a malfunction.

31.2. Analysis and Theoretical Investigation of New Military Electronic Missile and Aircraft-Borne Equipment

D. EHRENPREIS, *David Ehrenpreis, Consulting Engineers, New York, N. Y.*

Today's additional requirements for military electronic and electromechanical equipment, with regard to optimum minimum weight and space envelope and optimum reliability, are goals of the judicious design and analytical procedure discussed in this paper. The dynamic environment acts as forcing functions upon three mathematical regimes set up as a mathematical model defining the dynamic properties and characteristics of the military electronic equipment under investigation.

The new method of analyzing military electronic equipment has resulted in the judicious design procedures and recommendations for optimum performance in the severe environmental conditions of random vibrations, random shock, steady-state vibrations, sudden-impulse shock, and sustained acceleration.

31.3 Packaged High-Power Radar Transceivers

H. N. C. ELLIS-ROBINSON, *Marconi's Wireless Telegraph Co., Ltd., Chelmsford, Essex, Eng.*

This paper discusses certain aspects of the mechanical and electronic problems in the development and design of a series of single unit integral construction 5-mw radar transmitter-receivers, which have been developed to meet a specification for a single packaged unit suitable for world-wide military use in both shipborne and mobile applications. Each design, regardless of frequency band, was required to maintain a high degree of component and assembly interchangeability, and at the same time provide comprehensive inbuilt test instrumentation with unique simplicity of interpretation.

The paper concludes by discussing two of the resulting designs, one in S band and one in L band, based on a common structural form.

31.4. Limitations of the Output Pulse Shape of High-Power Pulse Transformers

R. G. DEBUDA AND J. VILCANS, *Canadian General Electric Co., Ltd., Toronto, Ont., Can.*

Modulators for ground radar and nuclear accelerators require pulse transformers giving sharply defined pulses of high power. This paper shows that these two requirements may conflict.

The geometry of a pulse transformer core is such that (using suitable material constants) a certain function of the pulse shape and power has a minimum value, which is given in the paper.

This can be interpreted as giving a limit on the pulse risetime, below which no pulse transformer with otherwise given characteristics can be designed.

31.5. A Radar Electronic Countermeasures Simulator

L. STERNLICHT, *The Hallicrafters Co., Chicago, Ill.*

Several aspects of the design of an electronic radar countermeasures simulator are described. The ECM and ECCM battle tactics require that countermeasures be used as a single round weapons system capable of the constant and rapid changes required by the tactical situation.

This paper presents a method for the simulation of radar interference waveforms at L, S, and X-band radio frequencies and at 30- and 60-mc intermediate frequencies. The video frequencies are obtained with a multitarget generator composed of two manually controlled single targets, a flight of several targets programmable as a unit, six individually programmed targets, and random targets which occur during an individual scan of the radar set. The targets are programmable in speed, course, pulse, and azimuth width.

The patterns produced by the random targets, programmed targets, and the other interference waveforms on a ppi presentation are analyzed.

* Sponsored by the Professional Group on Military Electronics. To be published in Part 8 of the 1958 IRE NATIONAL CONVENTION RECORD.

SESSION 32*

Wed. 10:00 A.M.–12:30 P.M.

New York Coliseum
Marconi Hall

MICROWAVE MEASUREMENTS

Chairman: W. L. PRITCHARD, *Raytheon Manufacturing Co., Wayland, Mass.*

32.1. Power Limiting Using Ferrites

R. F. SOOHOO, *Cascade Research Corp., Los Gatos, Calif.*

Making use of the nonlinear behavior of ferrites at high microwave signals, a device using a magnetized ferrite could be designed such that its attenuation is a function of the incident power level. If a band-pass filter is introduced in conjunction with such a limiter the combination can be made to reject all signals outside the desired band and limit to a preset low value the power within the band. An automatic-shutter-crystal protector can thus be obtained for pulse radar systems.

The buildup mechanism of the spin-waves (which account for nonlinear effects at high-power levels) and the time required for them to reach steady-state amplitudes are analytically investigated. Experimental verification of the theory and data on the performance of a power limiter will be presented.

32.2. An Ultraprecise Microwave Interferometer

G. R. BLAIR, *McMillan Lab., Inc., Ipswich, Mass.*

Electrical phase may be measured by use of a microwave interferometer. The device may be used to obtain dielectric constant and loss tangent. Equations are presented from which the dielectric constant may be obtained. The interferometer may be used to obtain data on large-size production items for nondestructive testing. There are no temperature limits at which characteristic data on a sample may be obtained.

An ultraprecise interferometer is described. It is used to obtain Δ values directly, accurate to ± 0.0001 inch.

The frequency stabilization unit consisting of a resonant cavity is described.

Curves are shown, portraying the extent of correction and easily obtained accuracy, feasible with this equipment.

32.3. Direct Reading Microwave Phase Meter

H. A. DROPKIN, *Diamond Ordnance Fuze Labs., Washington, D. C.*

A direct reading broad-band phase meter for frequencies between 9 and 10 kmc is described. The microwave signal, whose phase is to be determined, is superheterodyned to a 180-cps intermediate frequency, employing a local oscillator signal obtained by shifting the frequency of the X-band source. The phase of the IF signal, equal to the microwave phase, is readily determined. The phase measurement is independent of the insertion loss of the sample. No null seeking balances are required. The equipment constructed has a ± 4 degree phase error for insertion losses up to 25 db. The errors produced by various imperfect

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components and the ability to make high-precision measurements at selected frequencies are described. Simultaneous microwave amplitude measurement equipment is readily incorporated.

32.4. A Microwave Spin Resonance Spectrometer

R. R. UNTERBERGER, *California Research Corp., La Habra, Calif.*

A microwave instrument operating at X band (9200 mc) for the measurement of the real χ' and imaginary χ'' components of the magnetic susceptibility χ of liquids and solids is described. The sample is placed in a resonant cavity immersed in a uniform and controllable magnetic field. The sample is irradiated by microwaves tuned to cavity resonance. The microwave H field is oriented at right angles to the external magnetic field. An afc system keeps the klystron tuned to the resonant frequency of the cavity while the external magnetic field is swept slowly in time and also modulated by a low-frequency sinusoidal wave. The reflected microwave power is detected, amplified, phase-compared with the modulating sinusoid, and recorded while sweeping the magnetic field. A spin resonance is observed when the magnetic field H satisfies the equation $h\nu = g\beta H$, where ν is the microwave frequency and g, β, and h are constants.

32.5. The Use of Impulse Response in Electromagnetic Scattering Problems

E. M. KENNAUGH AND R. L. COSGRIFF, *Dept. Elec. Eng., Ohio State University, Columbus, Ohio*

It is customary to develop approximate solutions to electromagnetic scattering problems by assuming monochromatic sources and imposing further conditions on the ratio of physical dimensions to wavelength. If the ratio of physical dimension to wavelength is large, the solution is obtained as a perturbation of the geometrical or physical optics solution. If the ratio is small, the solution differs little from that predicted by Rayleigh scattering. It is not usually possible to obtain a single approximate solution which yields the correct result for the limiting cases of zero and infinite frequency.

This paper presents a method of approximating the exact solution to electromagnetic scattering problems for any source frequency by an approximation to the impulsive response of the scatterer. This approximation is chosen so as to yield the correct solution for the limiting cases of zero and infinite frequency. The relation between this approach and the physical optics approximation, as well as the quasi-static approximation, is discussed. The application of this technique is illustrated for several types of scatterers, and the results are compared with those obtained by standard methods.

SESSION 33*

Wed. 10:00 A.M.–12:30 P.M.

New York Coliseum
Faraday Hall

SEMICONDUCTOR DEVICES

Chairman: R. M. RYDER, *Bell Telephone Labs., Inc., Murray Hill, N. J.*

33.1. A New Passive Semiconductor Component

R. M. WARNER, JR., *Bell Telephone Labs., Inc., Murray Hill, N. J.*

A nonlinear component has been developed which can function as a current regulator, current limiter, current pulse shaper, choke, or ac switch. It can be designed for polar or nonpolar behavior. The device is based on the field effect principles developed by Shockley, Dacey, and Ross, and has been named the field effect varistor. All models made to date have embodied diffused junctions. These models have assumed a variety of forms; simple coaxial structures are among the most promising. Field effect device theory has been extended to cover these new geometries.

33.2. Use of the RCA 2N384 Drift Transistor as a Linear Amplifier

D. M. GRISWOLD AND V. J. CADRA, *Radio Corp., of America, Somerville, N. J.*

This paper describes the use of a *p-n-p* germanium drift transistor, the RCA 2N384, in several high-frequency amplifier applications.

The 2N384 has a common-emitter feedback capacitance of 1.3 $\mu\mu$f, a maximum available gain of 23 db at 30 mc per second, and a maximum collector-dissipation rating of 120 mw at 25°C. The frequency at which the common-base current transfer ratio drops to 0.707 times its 1 kc value is 100 mc.

The transistor is described in terms of typical variations of individual parameters as a function of temperature and operating point. Performance in the various amplifier applications is then discussed and data are presented showing the effect of voltage and temperature deviations on gain.

33.3. High-Current Switching Times for a *P-N-P* Drift Transistor. Numerical Analysis on the IBM 704 Digital Computer

A. MITCHELL, *IBM Corp., Poughkeepsie, N. Y.*, AND L. LAPIDUS, *Dept. Chemical Eng., Princeton University, Princeton, N. J.*

A one-dimensional model is proposed for predicting the high current turn-on and turn-off time of a *p-n-p* drift transistor. The system of equations describing this model is highly nonlinear and does not admit of an analytical solution. Numerical calculations on the IBM 704 are discussed with consideration of stability and convergence of the equations. Results are presented for drift transistors currently in use. The effect of the various physical parameters, such as the donor gradient, the intrinsic resistivity, and base width, are discussed.

33.4. A New High-Frequency Diffused Base Transistor

J. SARDELLA AND R. WONSON, *Raytheon Manufacturing Co., Newton, Mass.*

A diffused base transistor has been developed capable of performing as video amplifier and as a straight forward amplifier at frequencies between 5 and 10 mc. Detailed electrical performance characteristics are given indicating the performance potential of this device. Several novel construction details are given which point out the large production potential of this sort of device.

33.5. A New Five-Watt, Class A, Power Amplifier

G. FREEDMAN, J. WILLIAMS, P. FLAHERTY, D. ROOT, D. SPITTLEHOUSE, W. WARING, P. KAUFMANN, AND P. WHORISKEY, *Raytheon Manufacturing Co., Newton, Mass.*

A high-power diffused base transistor with high production potential has been developed. It can deliver 5 watts, Class A, at a temperature of 100°C. It is constructed of silicon utilizing a novel fabrication technique. This is the technique of bulk diffusion, where diffusion is combined with the more conventional fusion. Diffusion proceeds from the recrystallized region. Detailed analysis is made of electrical performance characteristics and their relation to thermal transfer at elevated wattages. Limitations of this design are given, as well as predictions for expansion into design of high-dissipation ranges.

SESSION 34*

Wed. 2:30–5:00 P.M.

Waldorf-Astoria
Starlight Roof

RELIABILITY THROUGH SYSTEMS

* Sponsored by the Professional Group on Electron Devices. To be published in Part 3 of the 1958 IRE NATIONAL CONVENTION RECORD.

* Sponsored by the Professional Group on Reliability and Quality Control. To be published in Part 6 of the 1958 IRE NATIONAL CONVENTION RECORD.

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Chairman: N. H. Taylor, *Lincoln Lab., Massachusetts Institute of Technology, Lexington, Mass.*

34.1. On an Analytical Design Technique

J. B. Heyne, *Hughes Aircraft Co., Culver City, Calif.*

This paper describes a technique for analytical system design. It is offered as a first step in the development of a comprehensive program. Because this technique lends itself to piecemeal application, a primary purpose of this presentation is to foster interest on the part of the system or circuit designer. The type of analysis presented here is generalized. The purpose of this presentation is to indicate the existence of component-system relationships. Insofar as environmental stress affects these relationships, it will also affect the corresponding system transfer functions. It becomes possible to pinpoint fundamental physical limitations on system performance and to explore alternative solutions, prior to hardware fabrication.

34.2. Reliability or Life Performance

A. R. Matthews, *Wright Air Development Center, Wright-Patterson Air Force Base, Ohio*

This report is a brief analysis of the "reliability" subject about which hundreds of specialized reports and bibliographies have been written.

The report includes a preliminary study of reliability concepts, programs, achievements, requirements, capabilities, and potential radical improvements through specific technological advances and contractual mechanisms. The requirements of future operational systems are shown to exceed presently anticipated future capabilities. *The concept of a "life" criteria is proposed and demonstrated as potentially applicable in both the electronics and mechanical areas.* The "life" factor is shown to be the essential quality from a complete system to the smallest part thereof.

Specific management and technological efforts are identified as being essential to progress that will achieve practical operational requirements of the future.

34.3. Reliability Improvement through Redundancy at Various System Levels

B. J. Flehinger, *IBM Watson Lab., Columbia University, New York, N. Y.*

Improvement in computing machine reliability through redundancy is studied as a function of the level at which the redundancy is applied. The reliability achieved by redundancy of complete independent machines is compared to that achieved by redundancy of smaller units.

A machine unit is termed m times redundant when:

1) m independent identical units operate simultaneously with a common input;
2) a failure detector is associated with each unit; and
3) a switch is connected to the outputs of the units, so that the output is taken from some one unit until failure occurs in that unit. Then the switch steps so that the output is taken from the next redundant unit, if that unit is operating correctly. This process continues until the assigned task is completed or all m units fail.

The reliability of m redundant units is expressed in terms of the reliability of one unit and the probabilities of correct operation of the failure detectors and switch.

It is assumed that a complete machine may be broken up into p units, $p=1, 2, 3, \ldots, 24$, of equal reliability. The reliability achieved by redundancy of these units is calculated as a function of p and m, $m=1, 2, 3, 4$, with single-machine reliabilities of 0.2, 0.5, 0.9, and 0.99. These results are calculated for perfect failure detection and switching devices, as well as for moderately unreliable devices. The resultant system unreliability is plotted as a function of p on linear and on logarithmic scales.

34.4. Fundamental Techniques in Doppler Radar Navigation System Reliability Measurements

P. D. Stahl, *General Precision Lab., Inc., Pleasantville, N. Y.*

The problems associated with establishing the performance reliability with ground support equipment for a highly accurate Doppler radar navigation system are investigated. Special attention is directed to the required instrumentation for simulating at the microwave, IF, and Doppler audio levels, the realistic Doppler echo returns as a function of aircraft ground velocity and drift angle, the two prime outputs of a Doppler navigator. Included in the discussion is a brief treatment of the general statistical nature of the Doppler signal and the significant phenomenon which must be simulated with reasonable fidelity in order to obtain the required performance data and a measure of the expected reliability of the prime outputs.

34.5. Reliability Prediction and Test Results on USAF Ground Electronic Equipment

E. Krzysiak and J. Naresky, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

If the military designer were supplied with a reliability prediction technique which would enable fairly accurate prediction of the quantitative reliability to be expected of an equipment in the field, untold funds and manpower might be saved through the elimination of engineering field changes, modification kits, and the need for large numbers of replacement parts.

This paper describes attempts to apply a prediction technique to several representative items of USAF ground electronics equipment. Discussed are: the theory of the prediction technique, the reliability "numbers" obtained for each equipment by use of the technique, the numbers obtained in laboratory tests of the equipment, the numbers obtained from rigidly controlled field failure results, and the correlations or discrepancies among the various figures obtained.

SESSION 35*

Wed. 2:30–5:00 P.M.

Waldorf-Astoria
Astor Gallery

INFORMATION THEORY: CODING AND DETECTION

Chairman: L. G. Fischer, *Federal Telecommunication Labs., Nutley, N. J.*

35.1. On Communication Processes Involving Learning and Random Duration

R. Bellman and R. Kalaba, *The Rand Corp., Santa Monica, Calif.*

Previously we have shown that the fundamental problem of determining the utility of a communication channel in conveying information may be viewed as a problem within the framework of multistage decision processes of stochastic type, and as such may be treated by the theory of dynamic programming. Furthermore, the relations between utility and capacity, in Shannon's sense, have been indicated.

Following a brief résumé of the foregoing, we show how to treat communication problems involving the use of a channel whose properties are not completely known, and those involving processes of random duration. There are special cases of still more general problems in prediction theory.

Our aim is to show how quite general processes can be treated in a uniform fashion by the functional equation technique of dynamic programming.

35.2. The Application of "Comparison of Experiments" to Detection Problems

N. Abramson, *Electronics Research Lab., Stanford University, Stanford, Calif.*

"Comparison of Experiments" is the name given to a recently developed topic in the theory of statistical decisions. This paper presents an explanation of the main concepts of this topic, derives a practical method of employing these concepts, and illustrates their application to radar detection.

The strength of the methods obtained from "Comparison of Experiments" lies in the fact that, in contrast with almost all previous results using statistical decision theory, these methods yield results which do not depend on the particular loss function of the problem or the particular form of the *a priori* distribution assumed.

35.3. Signals with Uniform Ambiguity Functions

R. M. Lerner, *Lincoln Lab., Massachusetts Institute of Technology, Lexington, Mass.*

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The response of a correlation detector or matched filter at time t to a signal differing by Doppler shift w from the expected one is a function of these two variables, $X(w, t)$, known as the ambiguity function. For a signal having a bandwidth, W, and duration, T, the rms value of $|X(w, t)|$ lies approximately $1/\sqrt{TW}$ below its peak value for all other w in W and all other t in T. A signal for which $|X|$ has a single peak, but otherwise lies uniformly at or below $1/\sqrt{TW}$ of the peak, is constructed from a spectrum which is a modified periodic binary shift register sequence. The resulting signal has finite energy and is realizable at band pass.

35.4. Evaluation of Some Error Correction Methods Applicable to Digital Data Transmission

A. B. Brown and S. T. Meyers, *Bell Telephone Labs., Inc., Murray Hill, N. J.*

Several methods of error correction in digital data transmission are evaluated theoretically. The evaluation follows procedures which are outlined and which are considered generally applicable in the orderly determination of the relative merits of these methods. The relative merits are considered on the basis of: 1) the ability of any correction method to achieve a specified final average error rate, given an initial error rate and distribution of lengths of noise bursts; 2) the increase in transmission time due to the addition of redundancy in the message; and 3) complexity of instrumentation. These considerations are summarized in curves of transmission efficiency vs initial error rate for given final average error rates and relative complexity of instrumentation vs coding efficiency.

35.5. Algebraic Decoding for the Binary Erasure Channel

M. A. Epstein, *Dept. of Elec. Eng. and Lincoln Lab., Massachusetts Institute of Technology, Cambridge, Mass.*

This paper presents an optimum decoding procedure for parity check codes that have been transmitted through a binary erasure channel. This procedure decodes the erased digits by means of modulo 2 equations generated by the received message and the original parity check equations. Most previous decoding procedures required a number of computations that grew exponentially with code length. At best, the required number of computations grew as a polynomial of code length. The decoding procedure for convolution parity check codes presented here will decode with an average number of computations per digit that is bounded by a finite number, which is independent of code length, for any rate less than capacity.

SESSION 36*

Wed. **2:30–5:00 P.M.**

Waldorf-Astoria
Jade Room

ELECTRONIC COMPONENT PARTS

Chairman: J. A. Csepely, *Westinghouse Electric Corp., Baltimore, Md.*

36.1. Effect of High-Intensity Radiation on Electronic Parts and Materials

C. P. Lascaro and A. L. Long, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

A selected group of electronic parts and materials were exposed in a static condition to various intensities of radiation from a field nuclear explosion. These exposures were high intensity doses delivered within periods of milliseconds. Electronic and physical measurements were made showing permanent effects on performance characteristics pertinent to electronic usage. Data of this type will be illustrated on electronic parts, and radiation effects observed on materials will be discussed and evaluated for their significance to electronic application.

36.2. Some Guideposts to the Use of Metallized Capacitors

W. C. Lamphier, *Sprague Electric Co., North Adams, Mass.*

This paper discusses two aspects of metallized capacitors. Miniaturization by commercial techniques is related to capacitor performance. The ac behavior of metallized capacitors is also presented along with improved designs with improved performance.

The first question considered is the volume efficiency which can be attained for various structures vs the performance which has been experienced. This comparison covers a high-quality metallized capacitor, a standard quality metallized capacitor, a metallized mylar capacitor, and a miniaturized metallized capacitor. The relative efficiencies of these designs are shown to be a function of capacity. The particular performance and temperature limitations of each are presented. This leads to the conclusion that the end circuit requirements will dictate which capacitor should be used. The greatest volume efficiency is attained at the sacrifice of temperature rating, voltage rating, or hermetic sealing.

The second part of the paper is a discussion of the ac performance of metallized capacitors. The results obtained when standard dc metallized capacitors are subjected to ac stress of 60, 400, and 800 cycles are presented. Two limiting factors to the ac use of metallized capacitors become apparent from the data; these are heat conductivity and voltage. The result is that low-capacity capacitors are superior to high, with respect to ac voltage, and as the voltage is increased added dielectric thickness is of little benefit.

Progress has been made toward increasing the heat conductivity of the capacitor structures and decreasing the sensitivity to voltage. The data collected on capacitors utilizing these approaches lead to the conclusion that dc designs can be improved to give better ac performance of metallized capacitors.

36.3. New Amplifiers for Automatic Control of Active DC Loads

E. Levi, *Microwave Research Institute, Brooklyn, N. Y.*

This paper introduces a class of magnetic amplifiers specifically designed for automatic control of active dc loads, such as the armature of a dc motor or an electrolytic cell.

Their main features are: 1) response within less than a half-cycle, 2) full wave or polyphase output with a single core, 3) zero sensitivity to rectifiers leakage and its variation with temperature and time, 4) limited control in the low counter emf range.

Considerable reduction in weight and cost, elimination of core-matching problems, fast response, and smoother feedback signals are achieved.

36.4. Magnetostriction Transducers for Mechanical Filters

R. L. Sharma and H. O. Lewis, *Collins Radio Co., Burbank, Calif.*

Equivalent circuits are obtained for magnetostriction transducers employing straight nickel wire and straight ferrite rods at intermediate radio frequencies. Losses due to eddy currents, hysteresis, and magnetomechanical damping are considered. Electromechanical coupling factor, transducer bandwidth, and transducer efficiency are defined. Experimental results are described relating important transducer characteristics to temperature and biasing field. An actual ferrite transducer design for a 455-kc electromechanical filter and some over-all filter response curves are shown and discussed.

36.5. Application of Piezoelectric Ceramic Resonators to Modern Band-Pass Amplifiers

A. Lungo and K. W. Henderson, *Clevite Corp., Cleveland, Ohio*

A new unique component for use in IF circuitry is described. Fabricated of a new piezoelectric ceramic of improved stability, the component takes the form of small radially resonant disks.

Two configurations of disks are described and their application as series and shunt couplers, as well as substitutes for emitter bypass capacitors in transistor IF stages, is discussed.

A special configuration, called ring and dot, is described. Also described is the application of these components in a reflex receiver.

These components, now becoming available, enable circuit designers to depart from conventional design considerations of shielding, size, and accessibility.

SESSION 37*

Wed. **2:30–5:00 P.M.**

Waldorf-Astoria
Sert Room

COMPUTERS AND CONTROL

* Sponsored by the Professional Group on Component Parts. To be published in Part 6 of the 1958 IRE National Convention Record.

* Sponsored by the Professional Group on Electronic Computers. To be published in Part 4 of the 1958 IRE National Convention Record.

World Radio History

Chairman: F. M. VERZUH, *Computation Center, Massachusetts Institute of Technology, Cambridge, Mass.*

37.1. A Preventive Maintenance Program for Large General Purpose Electronic Analog Computers

R. P. SYKES, *The Ramo-Wooldridge Corp., Los Angeles, Calif.*

The need for a systematic preventive maintenance program for large general purpose electronic analog computers, particularly in the combined analog-digital simulation situation, is shown. An analysis of the problem is given, and details of the preparation for and institution of the program are given. Details of the program, equipment required for its execution, and conclusions drawn as a result of several months' experience are described.

37.2. The TRICE—A High-Speed Incremental Computer

J. M. MITCHELL AND S. RUHMAN, *Packard-Bell Computer Corp., Los Angeles, Calif.*

A new parallel digital differential analyzer is described capable of 100,000 iterations per second at a maximum precision of 28 binary digits and sign. The TRICE (Transistorized Real-Time Incremental Computer—Expansible) consists of a number of self-contained digital integrators (the number depending on the complexity of the problem) interconnected by a plugboard that serves to program the computer. Thus, the TRICE combines the expansibility and ease of programming of the analog computer with the precision of the digital computer, at a speed several orders of magnitude greater than that of existing digital differential analyzers.

Both the logic and circuitry of the basic integrator are treated. Three delay-line registers and three adders, operating at a clock rate of 3 mc per second, perform trapezoidal integration. The component count is approximately 100 transistors and 450 diodes.

37.3. Digital Moon Radar Antenna Programmer with Analog Interpolator Servo

O. A. GUZMANN, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

Past attempts have been made to employ an analog computer for computing azimuth and elevation positioning data for a moon radar antenna. Due to the complicated orbit equation of the moon, the analog equipment became very complex and its accuracy was insufficient.

This paper describes a system for producing continuous flow of moon azimuth and elevation data, using digital techniques for computation, and storage and analog techniques for data transmission and interpolation. The discrete digital intervals are interpolated by means of a combined position-velocity servo. The stored data are time, azimuth, and elevation angles and their rates of change. The accuracy of the digital system is 25 times better than that of the analog computer, and costs less than 30 per cent of the latter.

37.4. A Balanced Precision Reference Regulator for Computer Application

D. A. NODEN, *The Martin Co., Baltimore, Md.*

This paper describes a dual output reference regulator that is designed to hold its outputs at 300 v dc ±0.01 per cent for input variations of −8+5 per cent and load variations of 0–250 ma. General design criteria of precision regulators are discussed first, with special consideration being given to methods of obtaining long-term stability. Choice of circuitry is discussed next, and a description is given of the development of a regulator to meet the following requirements.

1) Dual outputs, +300 v dc and −300 v dc.
2) Voltage regulation, 0.005 per cent.
3) Combined line and load regulation, 0.01 per cent.
4) Combined line and load regulation not to exceed 0.02 per cent for 100 hours.
5) Outputs to track each other in absolute magnitude to within 25 mv.
6) Output ripple, not to exceed 1 mv rms.
7) Output current, 0–250 ma each output.
8) Response time (time to return to within 1/10 Ep of quiescent value, where Ep is peak fluctuation, when load is switched from ½ load to ¾ load), 1 millisecond.
9) Input, + and −395 v dc −8+5 per cent.
10) Input ripple, 2 v rms maximum.

The discussion of the development includes consideration given to the choice of components with regard to reliability and interchangeability, such that the final design might be reproduced on a production line basis without special selection of any component. Finally, the instrumentation and testing of the regulator shall be described, including tests run to determine the maximum output current rating obtainable using paralleled series regulator sections.

37.5. A Solid-State Analog-To-Digital Conversion Device

M. PALEVSKY, *Packard-Bell Computer Corp., Los Angeles, Calif.*

A high-speed (4 μsec per bit), completely solid-state analog-to-digital conversion device that is accurate to 0.01 per cent is described. It employs a new switching scheme that does not require a high voltage reference, but rather takes advantage of certain saturation characteristics of transistors that have not been explored heretofore. As a result, the digital value that is formed can be the quotient of two input voltages or, using another configuration, the digital value of the square root of the quotient. A reference voltage supply with commensurate accuracy and response is described which employs a silicon transistor chopper and a zener diode reference. Various applications are discussed.

37.6. J-Axis Translation of Transfer Functions

J. L. RYERSON, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

Present day analog computers use many individual operational amplifiers which are chopper stabilized to provide low drift rate. Problems are solved by the use of a large number of such amplifiers, the process of modulation and demodulation taking place in each amplifier.

This paper gives a method of performing the desired operations by the use of a carrier system in which the desired operations take place in the sidebands.

SESSION 38*

Wed. **2:30–5:00 P.M.**

New York Coliseum Morse Hall

INSTRUMENTATION SYSTEMS

Chairman: F. HAMBURGER, JR., *Elec. Eng. Dept., The Johns Hopkins University, Baltimore, Md.*

38.1. An Earth Satellite Instrumentation for Cloud Measurement

R. HANEL AND R. A. STAMPFL, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

During the IGY, a number of scientific experiments are planned for use in an earth satellite. Instrumentation has been developed for these experiments. Restrictions such as weight, size, bandwidth, etc. were established by Project Vanguard of the Naval Research Laboratory, and instrumentation had to be developed within these limitations. This paper deals with an instrumentation for measurement of the cloud distribution over the earth. The apparatus consists of an optical system, a tape recorder, transmitter amplifiers, and switching circuitry. The design parameters for each of these components are derived. Illustrations show the compact construction of the completed package, the circuitry used, and the efficiencies obtained.

38.2. A Precise Optical and Radar Tracking Range

E. V. KULLMAN, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

Rome Air Development Center has instrumented an integrated optical and electromagnetic tracking range used for the testing and evaluation of electronic missile guidance systems. It consists of four precisely positioned Askania cine-theodolites, deployed in a diamond configuration for acquiring flight test data.

This paper outlines the techniques for accomplishing radar acquisition information for the optical trackers, the problems of synchronizing the operation of the remotely installed Askanias, the radio communications netting for technical control of the range, the acquisition of meteorological information to determine index of refraction, and the effects causing optical and electromagnetic tracking errors.

* Sponsored by the Professional Group on Instrumentation. To be published in Part 5 of the 1958 IRE NATIONAL CONVENTION RECORD.

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38.3. A High-Speed Radar Signal Measurement and Recording System

A. NIRENBERG AND R. BURFIEND, *Airborne Instruments Lab., Inc., Mineola, N. Y.*, M. BALLER, *Cambridge Research Development Center, Cambridge, Mass.* AND A. WRIGHT, *Laboratory for Electronics, Inc., Boston, Mass.*

An accurate, high-rate, digital radar data recording and handling system is described. The system was designed and fabricated for the purpose of gathering data on the physical and statistical characteristics of radar ground clutter, with the intention of providing a basis for new and improved techniques and circuitry for mti radar systems. Each of eight discrete parameters, describing a single radar target, can be quantized into an 11-bit binary number for each radar repetition period, and recorded on magnetic tape. These data are converted to punched paper tape form and made available to a digital computer for the calculation of such clutter characteristics as signal time functions, correlation functions, and probability distributions.

A concurrent circuitry study yielded information on the phase characteristics of IF amplifiers and limiters and on the accuracy of various phase detector circuits.

The clutter measurement system has been installed at an Air Force radar site and an active data gathering program is anticipated.

38.4. A High-Performance Multichannel Instrumentation System

W. G. WOLBER, *Bendix Aviation Corp., Detroit, Mich.*

Fuel controllers for modern jet engines are precision devices. A high-performance instrumentation system is required to test them. This paper describes such a system now in operation. The Dynamic Data Plotter measures twelve of forty-three possible input variables (gauge, absolute, and differential pressure, flow, rpm, temperature, and linear travel). Test data are recorded on magnetic tape. Final readout is on a large scale *X-Y* plotting board. Frequency response is 10 cps for reduced speed tape playback. Over-all accuracy (transducer input through plot output) is 0.1 to 0.6 per cent, depending upon the variable. Drift is less than 0.3 per cent per month. Operational reliability has been good.

38.5. Instrumentation Dynamically Analyzed for Optimum Reliability, Weight, and Geometric Space Envelope Subjected to Severe Vibrations and Shock

D. EHRENPREIS, *David Ehrenpreis, Consulting Engineers, New York, N. Y.*

Today, new severe military requirements for instrumentation with regard to optimum reliability, weight, and geometric space envelope are successfully fulfilled through rigorous dynamic analysis, engineering study, and theoretical investigation as discussed in this paper. The dynamic environment of random vibrations, sudden-impulse shock, and sustained acceleration acts as forcing functions upon three mathematical regimes defining the dynamic properties and characteristics of the instrumentation. The mathematical model created simulates the performance and response of the instrumentation to the environment under excitation in three mutually perpendicular axes. Recursion equations, dynamic matrices, and recursion tables are set up. Error functions and natural frequencies are determined utilizing the aid of a high-speed digital computer for numerical computations. The requisite deflections, rotations, stresses, and forces are determined. Open parameters are utilized in the analysis, to determine the dynamic effect of each design modification proposed for the instrumentation. The new method of analysis results in judicious design procedures and recommendations for optimum performance in the severe environment, especially for missile-borne and high-speed military aircraft-borne instrumentation.

SESSION 39*

Wed. 2:30–5:00 P.M.

New York Coliseum
Marconi Hall

MICROWAVE COMPONENTS

Chairman: S. B. COHN, *Stanford Research Institute, Menlo Park, Calif.*

39.1. A New Microwave Rotary Joint

W. E. FROMM, E. G. FUBINI, AND H. S. KEEN, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

A novel annular microwave rotary joint will be described. This rotary joint is characterized by very low loss, swr, and "wow," and has wide-band performance. In addition, it has a flat, annular form and a number of them can be stacked with all input (or output) lines passing through the hole in the center. The basic principle is the use of a large circular cylindrical coupling cavity operating in the TEM mode. By means of strip transmission line binary feeds, this cavity is excited uniformly around the circumference preventing excitation of the most troublesome higher modes. Additional mode suppressors are incorporated to produce very high performance.

39.2. High-Power, Broad-Band, Microwave Gas Discharge Switch Tube

S. J. TETENBAUM AND R. M. HILL, *Sylvania Electric Products, Inc., Mountain View, Calif.*

* Sponsored by the Professional Group on Microwave Theory and Techniques. To be published in Part 1 of the 1958 IRE NATIONAL CONVENTION RECORD.

An electronically controlled, broad-band, gas discharge switch tube has been developed which is capable of rapidly switching high-power pulsed microwaves at *S* band. The switch tube consists of a section of waveguide sealed off at both ends by vacuum windows and filled with a gas at low pressure. It is located in a magnetic field oriented perpendicular to the electric field in the waveguide. The operation of the switch tube is based on the phenomenon of electron cyclotron resonance. The tube is controlled by the strength of the applied magnetic field. Bandwidths of more than 30 per cent can be achieved. The electrical characteristics of the tube and their variation with type of gas filling, pressure, input power, magnetic field, and geometry were studied, both experimentally and theoretically. The experimental results are in qualitative agreement with theory.

39.3. High-Power Microwave Filters

J. H. VOGELMAN, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

Filter sections have been designed suitable for combination in high-power microwave filters capable of handling 700 kw at 10 pounds pressure in 0.900 inch by 0.400 inch-ID waveguide. Design procedure for a multielement filter will be described, together with the measured results of a typical design.

39.4. A Band Separation Filter for the 225–400-MC Band

A. I. GRAYZEL, *Lincoln Lab., Massachusetts Institute of Technology, Cambridge, Mass.*

This paper describes the design and construction of a band separation filter for the 225–400-mc band. The filter utilizes transmission line components and can therefore handle relatively high power. The vswr over the band is less than 1.6:1 and the insertion loss is less than 1 db. There is a 23-mc crossover between 287 and 310 mc.

The individual high and low-pass filters consist of series lines and shunt short-circuited stubs. The pass and stop bands of the two filters are staggered to give the appropriate stop and pass characteristics over the band of frequencies of interest.

The individual filters were designed such that the characteristic impedance of each in its stop band approximates the impedance of the shunt element of the other. Then when the shunt elements at the junction were removed and the two interconnected a match was obtained over the band.

After fabrication, the filter was tuned experimentally to compensate for discontinuities not taken into account in the theory.

39.5. Direct-Coupled, Band-Pass Filters with $\lambda/4$ Resonators

G. L. MATTHAEI, *The Ramo-Wooldridge Corp., Los Angeles, Calif.*

Filters can be designed with resonators approximately $\lambda/4$ long using, alternately, series and shunt coupling discontinuities. Advantages over filters with $\lambda/2$ resonators include: shorter length, second pass band is centered at $3f_0$ instead of $2f_0$ (f_0 is design band center), midstop-band attenuation is higher, precision de-

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sign for a prescribed insertion loss characteristic is tractable to greater bandwidths, and can be made in "bar transmission line" form without dielectric supports. The design procedure used is on the insertion loss basis and incorporates an improved low-pass to band-pass transformation due to Cohn. Use of $\lambda/4$ direct-coupled resonators for mounting crystals on strip transmission line band-pass filters is also discussed.

SESSION 40*

Wed. 2:30–5:00 P.M.

New York Coliseum
Faraday Hall

PROPAGATION AND ANTENNAS I—GENERAL

Chairman: H. G. Booker, *Elec. Eng. Dept., Cornell University, Ithaca, N. Y.*

40.1. Extreme Useful Range of VHF Transmission by Scattering from the Lower Ionosphere

R. C. Kirby, *National Bureau of Standards, Boulder, Colo.*

Results are given from an experimental study of vhf propagation in the extreme distance range for scattering from the lower ionosphere. Signal intensity at 36.0 mc was measured continuously for a year over the 1411 mile path from St. Johns, Newfoundland, to Terceira Island, The Azores, using high transmitting and receiving sites. The median transmission loss is approximately 13 db greater than for the same system operated over the 743 mile path, Cedar Rapids, Iowa, to Sterling, Va.; this ratio is interpreted in terms of the geometrical restriction of the effective scattering volume. Less pronounced diurnal and seasonal variation is related to occluding by earth curvature of scattering from heights below about 80 km. Continuous "height gain" observations give evidence of variation of scattering height diurnally and seasonally. Results are given on the nature of the signal fading, space-correlation, realizable gain from arrays having extensive vertical aperture, and polarization effects.

40.2. Meteor Trail Propagation

J. T. deBettencourt and A. Ward, *Pickard & Burns, Inc., Needham, Mass.*, and B. Goldberg, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

This program was initiated in early 1957 for the purpose of investigating the feasibility, advantages, and limitations of meteor scatter communications at ranges of less than 500 miles. Almost all prior experimental and theoretical investigation had been centered on either radar (backscatter) or maximum (1200 mile) circuits. Two mobile stations were instrumented. The transmitter consisted of a 1-kw linear amplifier excited from a stable (1 part 10^9) frequency Shift Keyer generating simulated standard speed teletype signals. The receiving station uses space diversity antennas and special receiving techniques based on "predicted wave" correlation schemes. An interfering signal was also available from a third station located on the South Jersey shore.

Tests were conducted at various ranges using frequencies in the 30–60-mc band. The evaluation of these data is presented showing qualitative and quantitative conclusions with respect to capacity and capability of this type of radio communication for the distances previously noted.

40.3. The Duty Cycle Associated with Forward-Scattered Echoes from Meteor Trails

H. J. Wirth and T. J. Keary, *U. S. Navy Electronics Lab., San Diego, Calif.*

Signals propagated by meteor trails from a transmitter at Stanford, Calif. have been monitored in San Diego. The duty cycle, per cent of time that received signal exceeds a given threshold level, is analyzed in terms of its variation during the day at a fixed threshold and its variation with level at a fixed time. Overdense trails account for most of the duty cycle and its diurnal variation follows the diurnal variation of the rates of occurrence of overdense trails. The duty cycle varies with threshold level as $1/A^m$. This variation is similar to the distribution of peak amplitudes of the signal bursts.

40.4. A New Low-Frequency Antenna

E. W. Seeley and J. D. Burns, *U. S. Naval Ordnance Lab., Corona, Calif.*

The basic problem in very low-frequency antenna design is to obtain higher radiation resistance and bandwidth, and less ground and corona losses from a very short antenna. A new type short resonant monopole has been designed having several advantages over conventional loaded monopoles. As a result of a theoretical study and measurements made on the loaded folded monopole, it appears possible to build efficient antennas as low as 0.03λ high with resonant radiation resistance of 20 ohms and a half-power bandwidth of 10 per cent. Relative power measurements show the radiated field intensity is approximately equal to a $\lambda/4$ monopole.

40.5. Logarithmically Periodic Antenna Designs

R. H. DuHamel and F. R. Ore, *Collins Radio Co., Cedar Rapids, Iowa*

Research on new types of broad-band logarithmically periodic antenna structures is described. The antennas have patterns and impedance characteristics which are essentially independent of frequency over bandwidths of ten to one. A turnstile structure which has a horizontally polarized omnidirectional pattern and a planar structure which has a vertically polarized bidirectional beam are described. Wire approximations to the metal sheet structures also are discussed. The results of current distribution measurements on the structures are presented and analyzed.

40.6. Phase Center of Helical Beam Antennas

S. Sander, *RCA Defense Electronic Products Div., Camden, N. J.*, and D. K. Cheng, *Syracuse University, Syracuse, N. Y.*

When a helical beam antenna is used as a primary source of circularly polarized waves to illuminate a reflector, it is necessary to know the location of its phase center. The first part of this paper formulates an analytical method of determining the phase-center location of helical beam antennas based upon the expressions of the far-zone diffraction fields. Computations have been carried out for typical cases for both E_θ and E_θ components in the principal planes. The second part presents the experimental results. Dependence of the location of the phase center on the number of turns, the pitch angle, and the size of the ground plane is discussed.

SESSION 41*

Thurs. 10:00 A.M.–12:30 P.M.

Waldorf-Astoria
Starlight Roof

MAGNETICS AND COMPUTERS

Chairman: R. E. Meagher, *Digital Computer Lab., University of Illinois, Urbana, Ill.*

41.1. A High-Speed N-Pole, N-Position Magnetic Core Matrix Switch

A. L. Lane and A. Turczyn, *Technitrol Eng. Co., Philadelphia, Pa.*

This paper describes how miniature square loop ferrite cores can be used in a high-speed n-pole, n-position switch. A magnetic core matrix utilizing a coincident current bias system is utilized as the nucleus of this switch. Three matrix switches, namely, a three by three, ten by ten, and fifty by fifty, have been constructed to demonstrate the feasibility of operation at frequencies up to 1 mc. A novel scheme for core noise cancellation has been developed. The associated electronic circuits are also described. Though this paper presents a solution to a specific problem, the design techniques discussed can be applied to other similar problems such as those found in the computer and telemetering fields.

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* Sponsored by the Professional Group on Antennas and Propagation. To be published in Part 1 of the 1958 IRE National Convention Record.

* Sponsored by the Professional Group on Electronic Computers. To be published in Part 4 of the 1958 IRE National Convention Record.

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41.2. Apertured Plate Memory: Operation and Analysis

W. J. HANEMAN AND J. LEHMANN, *RCA Labs., Princeton, N. J.*, AND C. S. WARREN, *Radio Corp. of America, Camden, N. J.*

At the 1956 Eastern Joint Computer Conference, J. A. Rajchman delivered a paper on "The Ferrite Apertured Plate for Random Access Memory." The three methods of operating the plates proposed in that paper which have been investigated at RCA are: 1) the dc biased-switch driven memory reported by V. L. Newhouse and M. M. Kaufman (Conference on Magnetism, Washington, D. C., November, 1957); 2) set-a-line switch system; and 3) current-coincidence.

The present paper includes: 1) a detailed report on a set-a-line switch driven memory based on the performance of a prototype consisting of 64 pairs of 256-hole memory plates and having a storage capacity of 16,348 bits, including details of the transistorized sensing and rewrite circuits; and 2) a comprehensive analysis of the three methods and an appraisal of their respective merits.

The aperture memory plate appears to offer many advantages in compactness, simplicity of assembly, simplicity of driving circuits, and economy of driving power for a given access time. The dc biased switch-driven and current-coincidence systems appear to be the most promising.

41.3. Molecular Storage and Read-Out with Microwaves

C. H. BECKER, R. L. PIERCE, AND J. R. MARTIN, *Trionics Corp., Madison, Wis.*

A new principle of molecular storage and read-out with conducted microwaves is described. The domains of ferrites or garnets are permanently magnetized from the torque applied to the magnetic moments by properly oriented magnetic fields of microwave frequencies. The microwave interaction occurs along a certain propagation length in a molecular amplifier arrangement. The resonance sharpness is proportional to the ratio of the positively polarized phase velocity to the propagation length. A metallic or electron beam microwave conductor produces permanent dots in the ferrite or garnet material. These permanent magnetic dots (or bits of stored information) are read out nondestructively with the same type of molecular amplifier arrangement as for the radiomagnetic storage.

41.4. Calculation of Flux Patterns in Ferrite Multipath Core Structures

S. A. ABBAS AND D. L. CRITCHLOW, *IBM Research Center, Poughkeepsie, N. Y.*

Methods are described for the prediction of the final flux patterns resulting from driving multipath ferrite cores by a step mmf. It is shown that there exist three regions of constant incremental permeability within a core, based on an idealized *B-H* loop for the material. Laplace's equation is solved taking into account the potentials due to the current, surfaces of distribution of magnetic polarity at the physical boundaries of the core, and polarities induced at the boundaries separating the different regions of constant incremental permeabilities. Various drive winding configurations are considered. The results of the mathematical analysis are approximate but can be further refined.

41.5. Logic by Ordered Flux Changes in Multipath Ferrite Cores

N. F. LOCKHART, *IBM Research Center, Poughkeepsie, N. Y.*

A new technique has been developed for performing logic using multipath cores. The geometry of the cores is such as to cause quantification of the input flux changes and predetermine the order of switching of the output flux paths. By utilizing the many patterns of magnetization which can be realized in such structures, more than one functional relationship among the input variables can be realized. There is a saving in space and in the number of components required for performing a given amount of logic. Specific examples of devices and their operating characteristics will be shown.

41.6. Flux Responsive Magnetic Heads for Low-Speed Read-Out of Data

L. W. FERBER, *Clevite Corp., Cleveland, Ohio*

The basic principle of operation of flux responsive magnetic heads is that the output signal voltage generated is substantially proportional to the magnitude of the magnetic flux on the recorded medium, and is therefore independent of the speed of the medium. This differs from the conventional type of magnetic heads whose output signal voltage is proportional to the time derivative of flux on the recorded medium. Flux responsive magnetic heads also reproduce an accurate facsimile of the recorded flux since there is no phase shift between the recorded flux and the output signal.

This paper presents sine wave and pulse information in the form of basic performance data to aid the systems designer in producing the most effective system to meet his requirements. For example, the data are applicable in evaluating the performance of flux responsive heads as used in measurement and control systems for sensing elements or in computer systems for low-speed read-out transducers to actuate output devices.

SESSION 42*

Thurs. 10:00 A.M.–12:30 P.M.

Waldorf-Astoria Astor Gallery

* Sponsored by the Professional Group on Circuit Theory. To be published in Part 2 of the 1958 IRE NATIONAL CONVENTION RECORD.

CIRCUIT THEORY II—UNUSUAL ASPECTS OF FILTER DESIGN

Chairman: B. J. DASHER, *Director, School of Elec. Eng., Georgia Institute of Technology, Atlanta, Ga.*

42.1. Multichannel-Filter Synthesis in Terms of Dipole Potential Analog

H. A. WHEELER, *Wheeler Labs., Great Neck, N. Y.*

A filter whose frequency band is divided among parallel channels is beyond the capabilities of the usual unipole form of the potential analog. The synthesis of such a filter is based on additive transfer impedances, so it requires the less familiar dipole form of the potential analog. In this form, each natural frequency of resonance is represented by a dipole whose moment is a vector equal to the complex coefficient of the corresponding term in a partial-fraction expansion; then the complex potential is the analog of the complex impedance. Multichannel dividing and recombining filters for contiguous bands are exemplified by simple cases showing the application of the dipole analog.

42.2. Minimum Insertion Loss Filters

E. G. FUBINI, *Airborne Instruments Lab., Mineola, N. Y.*, AND E. A. GUILLEMIN, *Massachusetts Institute of Technology, Cambridge, Mass.*

A generalized criterion is given for the design of minimum loss Butterworth and Tchebycheff filters of arbitrary bandwidth and component quality. This criterion is chosen to minimize the insertion loss in the center of the band and is particularly useful for microwave filters, where insertion loss must be minimized. Two general curves will be given for this design, together with information as to the method followed in obtaining them. It will be shown that, if insertion losses are to be minimized, the use of Tchebycheff filters with ripples greater than very small fractions of one decibel must be discouraged.

42.3. A New Approach to the Design of High-Frequency Crystal Filters

R. A. SYKES, *Bell Telephone Labs., Inc., Whippany, N. J.*

The problems peculiar to the performance and design of high-frequency crystal filters are presented and discussed. The derivation of design equations for lattice and ladder type sections are shown which allows a quick appraisal of the feasibility for a given set of filter requirements. Finally, specific designs for single-sideband transmission and reception are presented for frequencies in the vicinity of 9 mc.

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42.4. Synthesis of Active RC Single-Tuned Band-Pass Filters

J. J. Bongiorno, *Microwave Research Institute, Brooklyn, N. Y.*

This paper deals with the synthesis of active RC single-tuned band-pass filters wherein classical negative feedback and transistors are employed. Four different loop transmissions are examined and compared. One of the four is shown to yield better performance, from a sensitivity standpoint, than the others.

The limitations imposed on the filter synthesis, in the best case, by transistor parasitics is investigated. An approximate relationship between the Q obtainable with parasitics present and the Q obtainable with the transistor considered an ideal current amplifier is developed. The sensitivities obtainable with parasitics present are shown to be essentially the same as those obtainable under ideal conditions.

42.5. A New Class of Filters

A. Papoulis, *Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

A class of filters is developed having an amplitude characteristic

$$A(\omega) = \frac{A_0}{\sqrt{1 + L(\omega^2)}}$$

without ripple in the pass band and with a high rate of attenuation in the stop band; it thus combines the desirable features of the Butterworth and Tchebycheff response. This new class is determined from the polynomials $L(\omega^2)$ given by

$$L(\omega^2) = \int_{-1}^{2\omega^2-1} v(x)dx,$$

where

$$v(x) = a_0 + a_1P_1(x) + \cdots + a_kP_k(x)$$

$$a_0 = \frac{a_1}{3} = \cdots = \frac{a_k}{2k+1} = \frac{1}{\sqrt{2}(k+1)}$$

and the P_k's are the Legendre polynomials of the first kind. Among all filters of a given order it has the maximum cutoff rate under the condition of a monotonically decreasing response.

SESSION 43*

Thurs. 10:00 A.M.–12:30 P.M.

Waldorf-Astoria
Jade Room

ULTRASONICS II—DELAY LINE MEASUREMENTS

Chairman: C. M. Harris, *Electronics Research Labs., Columbia University, New York, N. Y.*

* Sponsored by the Professional Group on Ultrasonics Engineering. To be published in Part 2 of the 1958 IRE National Convention Record.

43.1. Measurements of Delay in Ultrasonic Systems

D. L. Arenberg, *Arenberg Ultrasonic Lab., Inc., Jamaica Plain, Mass.*

Measurement of delay time in ultrasonic systems with any precision can involve many difficulties and errors, dependent on the range of delay, sample size frequency range covered, and whether continuous or transient waves are considered. The different problems will be discussed and means for obtaining the greatest accuracy illustrated, using readily available apparatus. Comparison methods of obtaining differences of 1×10^{-6} parts will be shown.

43.2. Precise Measurement of Time Delay

J. E. May, *Bell Telephone Labs., Inc., Whippany, N. J.*

A time delay measuring system is described capable of at least ±10 ppm precision for delays from two to several thousand microseconds. Developed for ultrasonic delay lines, the method is generally applicable to the determination of delay between repetitive pulses provided their repetition rate can be synchronized with the measuring circuit. Both undelayed and delayed pulses are displayed on the vertical axis of an oscilloscope, against a horizontal time base provided by a stable oscillator whose frequency is varied to superimpose the two pulses. Thus, time delay is determined in terms of frequency which can be measured with high precision.

43.3. The Measurement of Delay-Line Transducer Resistance

J. J. G. McCue, *Lincoln Lab., Massachusetts Institute of Technology, Lexington, Mass.*, and J. A. Leavitt, *Harvard University, Cambridge, Mass.*

Some delay lines of recent design have quartz transducers with equivalent shunt resistance less than 1000 ohms. The circuit designer needs to know the resistance when it is as low as these figures. The present discussion will deal with three methods of measurement: the variable external shunt, the Q meter, and the admittance bridge. When adequate precautions are taken, all these methods agree within about ten per cent; over the range primarily of interest, the Q meter gives the most rapid results.

43.4. Ultrasonic-Delay-Line Terminating Circuits and Pass-Band Measurements

M. Axelbank, *Lincoln Lab., Massachusetts Institute of Technology, Lexington, Mass.*

Recent experience with high-frequency, large-capacitance delay lines indicates that internal series lead inductance may prevent proper operation of conventional delay-line terminating circuits. A single-tuned terminating circuit is described that is not disturbed by series lead inductance. Its gain-bandwidth product is the same as that of the conventional single-tuned circuit. It is well suited to multichannel and single-channel large-bandwidth operation.

The circuit has been adapted to the problem of measuring the band-pass curve of a delay line. Satisfactory measurements have been made on delay lines up to and beyond the frequency of input series resonance, where the series lead inductance resonates the transducer capacitance and the input impedance drops to a very low value.

43.5. Measurement of Temperature and Frequency Dependence of Insertion Loss in Delay Lines

A. H. Meitzler, *Bell Telephone Labs., Inc., Whippany, N. J.*

Ultrasonic delay lines using fused silica are commonly made having delay times as long as 3000 μsec and operating at carrier frequencies from 10 to 60 mc. Such lines may have an appreciable temperature dependence of insertion loss, depending upon the length of the lines and the carrier frequency at which they are operated.

This paper discusses a useful method of measuring the temperature dependence of insertion loss for delay lines, and a method of relating the measured insertion loss of a given line to a quantity which depends primarily upon the length of path and the delay medium, rather than upon the particular combination of transducers and termination conditions. Sample data for the measured variations in insertion loss with temperature and frequency in the case of fused silica delay lines will be presented. The observed variations are briefly interpreted in terms of a relaxation effect in fused silica.

43.6. The Measurement of the Total Spurious Responses of an Ultrasonic Delay Line

M. S. Zimmerman, *General Atronics Corp., Bala Cynwyd, Pa.*

In certain applications of ultrasonic delay lines, a more accurate prediction of the distortion which will be produced as a result of the spurious transmission paths in the delay line can be made on the basis of the combined effect of all such responses rather than the maximum amplitude of any one response.

A system which can directly and conveniently measure total spurious responses has been developed, and is capable of measuring total spurious responses as low as 55 db below the level of the main delayed signal. A substitution method is utilized, allowing accurate measurements to be made without relying on the linearity of system amplifiers.

SESSION 44*

Thurs. 10:00 A.M.–12:30 P.M.

Waldorf-Astoria
Sert Room

* Sponsored by the Professional Group on Industrial Electronics. To be published in Part 6 of the 1958 IRE National Convention Record.

INDUSTRIAL ELECTRONICS

Chairman: E. W. Leaver, *Electronic Associates, Ltd., Willowdale, Ont., Can.*

44.1. Distributor Test Stand

J. A. Lovell, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

This paper describes the underlying considerations that went into the development of a test stand that is used for the final testing of mass-produced automobile distributors. This test stand automatically displays the following characteristics of a distributor on a large-screen oscilloscope: 1) advance angle vs speed; 2) advance angle vs vacuum; and 3) dwell angle. All of the distributor characteristics can be thoroughly checked in less than 36 seconds with an accuracy of ±2 per cent. The angular measurements are made by taking the time integral of the voltage produced by a precision dc tachometer.

44.2. A Digital Setting System for an X-Ray Thickness Gauge

V. A. Blumhagen, *General Electric Co., Milwaukee, Wis.*

The digital setting system is a system used to set up an X-ray thickness gauge to the nominal thickness of material to be gauged or measured. The thickness gauge is set up by positioning a precision reference wedge in the reference X-ray beam by means of a balanced bridge servosystem. One side of the bridge is a potentiometer coupled to the wedge and motor and the other side of the bridge is made up of four decade resistors. The setting is displayed in thousandths of an inch with an in-line electronic display unit both at the control station and at several remote locations.

44.3. Application of Magnetic Core Logic to Industrial Controls

H. Tellefsen and S. Alessio, *Panellit, Inc., Skokie, Ill.*

The series magnetic complementary amplifier proved in digital computers has been adapted for industrial systems. In such systems, reliability is a more stringent requirement than speed of operation, which is limited by transducer response. These requirements have been met in the development of a simpler, more reliable pulse type series magnetic amplifier.

Design and experimental verification of the magnetic logical element is discussed. The paper describes the circuitry evolved from the basic element. In particular, a shift register is illustrated having the features of series or parallel input and output with asynchronous forward-backward operation.

44.4. A Coordinated System of Automatic Control

R. R. Batcher, *Douglaston, N. Y.*

In applying automatic control using electrical and electronic methods to many processes and machines for some potential applications, the limited sales possibilities do not justify a large expense for engineering the specific design and for tooling the special parts needed. Thus, many useful applications do not even get started. This paper deals with a coordinated system of control elements that can be assembled mainly from stocked universal parts, that will handle both simple and intricate problems in this field with a minimum of designers' attention and development expense. Along with descriptions, models of this new approach will be shown and their operation demonstrated.

The system includes the control of circuit switching by punched cards and punched tape, relay arrays and switching networks, sampling and programming systems, high-speed brushless time-sharing commutators, coordinated electrical, electronic, and pneumatic possibilities, and, finally, a system of power switchboard construction using modules that eliminate all assembly and wiring drawings and cables. The design of these basic components of control permits extremely simple constructions of complete systems.

SESSION 45*

Thurs. 10:00 A.M.–12:00 Noon

Waldorf-Astoria Grand Ballroom

ASPECTS OF RF INTERFERENCE IN MILITARY ELECTRONIC AND COMMUNICATIONS SYSTEMS

Chairman: C. L. Engleman, *Engleman & Co., Washington, D. C.*

45.1. Treatment and Methods for the Reduction of Pulse and Random Interference

P. M. Creutz, *Packard-Bell Electronics Corp., Los Angeles, Calif.*

This paper presents a study and evaluation of pulse and random noise types of interference experienced by the various communications systems considered (pdm, pcm, ppm, fm, cw, and AM). A conclusion is made regarding the limitations of improvement in each of the systems, and possible methods of noise reduction through new circuitry approaches are outlined.

45.2. Reduction of Bandwidth Requirements for Radio Relay Systems

D. L. Jacoby and R. H. Levine, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*, and A. Mack and A. Meyerhoff, *Radio Corp. of America, Camden, N. J.*

Reduction of bandwidth required for TDM transmission in military fm radio relay systems is extremely important from the standpoint of spectrum utilization, interference, and vulnerability. Minimum video bandwidth may be calculated on the basis of interchannel crosstalk requirements and sync pulse resolution in the case of ppm systems, and on the basis of maximum tolerable intersymbol interference in the case of pcm systems. Pulse shaping networks are provided in the transmitter video, the IF amplifier, and the receiver video circuits to limit the spectrum transmitted and simultaneously reduce the bandwidth of the receiver. Narrow pulses are provided at the input to the transmitter in order that the pulses at the output of the receiver will represent the impulse response of the transmitter-receiver combination. By employing video reshaping at each repeater point, minimum receiver bandwidth is achieved. A comparison is made between the minimum bandwidth requirements for ppm, binary pcm, and quaternary pcm signals. Use of bipolar pcm results in additional reduction in bandwidth requirements, as well as an improvement in noise threshold when compared with conventional pcm. A reduction of 12 to 1 in receiver bandwidth has been achieved relative to existing systems.

* Sponsored by the Professional Groups on Communications Systems, Military Electronics, and Radio Frequency Interference. To be published in Part 8 of the 1958 IRE National Convention Record.

45.3. Analysis of the Spectral Shape of Modulation Splatter

R. Price, *Lincoln Lab., Massachusetts Institute of Technology, Lexington, Mass.*

Transmitter overload gives rise to "modulation splatter," an undesirable spreading of the transmitted spectrum beyond the normal band. Using Gaussian noise to represent the modulating signal, the decrease in the splatter spectrum is investigated as a function of ω_d, the frequency relative to that of the normal band, both for SSB and AM transmitters. Solution for the asymptotic spectral behavior involves an apparently novel application of the Central Limit theorem. Five smooth-saturation-type overload characteristics are studied in detail, and the spectrum is found to fall off initially as ω_d^{-3} for SSB and ω_d^{-2} for AM. At still greater ω_d, a faster decrease is found, which relates to the degree of smoothness of the saturation characteristic.

45.4. Near-Zone Power Transmission Formulas

M. K. Hu, *Elec. Eng. Dept., Syracuse University, Syracuse, N. Y.*

A general power transmission formula for a matched lossless two-antenna system is derived rigorously from Maxwell's equations through the use of reciprocity theorems. The formula is expressed in terms of field quantities only. It is applicable for a system where the two antennas can be of any physical size, at any relative position, and of any polarization. For linearly-polarized uniform-phase plane aperture antennas, facing each other directly and with the reflections between the antennas neglected, a considerably simplified formula is given. In far zone, the above formulas can be reduced to the well-known far-zone transmission formula expressed in terms of "gains" or "absorption cross sections" of both antennas.

It is shown that the concepts of gain and absorption cross section are applicable in the far zone only, and it is not possible to generalize in either the near zone or the Fresnel zone.

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It is also pointed out that the so-called Fresnel-zone gain appearing in the literature is of very limited application; it is limited to the case where one of the antennas is a point source or of very small size.

SESSION 46*

Thurs. 10:00 A.M.–12:30 P.M.

New York Coliseum
Morse Hall

DATA REDUCTION AND RECORDING

Chairman: R. J. BIBBERO, *Bulova R & D Labs., Inc., Woodside, N. Y.*

46.1. Instrumentation for Recording and Analysis of Audio and Subaudio Noise

D. D. HOWARD, *Naval Research Lab., Washington, D. C.*

Audio and subaudio noise are of concern in many fields of science, and considerable effort has been placed on analysis of noise data. Conventional techniques for graphical analysis of noise data are laborious and time consuming resulting in the instrumentation of automatic noise analysis equipment. The techniques for recording subaudio noise and for use of frequency multiplication of the noise for ease of handling are described, along with instrumentation for providing both automatic spectrum analysis and amplitude probability distribution analysis of the noise data. This analysis technique which requires approximately 15 minutes eliminates the many man hours required by laborious graphical methods.

The spectrum analysis and probability distribution provide a convenient presentation of noise data from which total power, rms voltage, and other values may be readily computed.

46.2. A Xerographic Cathode-Ray Tube Recorder

H. H. HUNTER, O. A. ULLRICH, AND L. E. WALKUP, *Battelle Memorial Institute, Columbus, Ohio*

This paper describes research on a high-speed, direct-writing, cathode-ray tube recorder.

Three methods of direct recording are reported. Two of these methods are described briefly, and the most promising system is described in detail. The most promising method utilizes a special cathode-ray tube with a row of small conducting pins imbedded in the face of the tube. In operation, these pins transfer electrical charge from the electron beam of the tube to an insulating sheet adjacent to the outer ends of the pins, forming on this insulating sheet an electrostatic image which may be processed into a permanent, visible image by xerography.

Writing speeds of 100 meters per second and image resolution as great as 15 lines per millimeter have been achieved, and neither appears to be the ultimate attainable.

An analysis of the system is made to show its limitations and potentialities. Preliminary design data are discussed and suggestions are made for improvements and applications.

46.3. Theory of Magnetography

S. J. BEGUN, *Clevite Corp., Cleveland, Ohio*

It is known in magnetic tape recording that poles of induced magnetic fields attract fine particles of ferromagnetic powder. This method of obtaining a visible representation of recorded signals is called magnetography.

Our studies were directed to controlling the formation of latent images which can serve as a suitable source of stored energy for producing visible pictures or traces.

Three different types of magnetic heads were explored for creating latent images, one of essentially conventional design, another one employing principles similar to those used in the boundary displacement technique, and a third using single line conductors. The operation of these three heads will be evaluated in relation to three functions: 1) producing facsimiles of acceptable tone scale, 2) producing a graphic trace, and 3) producing letter and number characters.

46.4. Applications of Magnetography to Graphic Recording

J. B. GEHMAN, *Clevite Corp., Cleveland, Ohio*

A high-speed oscillographic recorder is described which uses magnetic tape for a graphic recording medium. Such a graphic recorder features a frequency response in excess of 10 kc and has no moving parts in the writing head. Besides producing a rectilinear trace, it prints its own timing and calibration lines simultaneously.

The latent image is induced in the magnetic tape. The magnetic tape is driven in excess of 200 inches per second and is processed continuously within the same instrument at a reduced speed of 10 inches per second.

Recordings may be made under extreme ranges of atmospheric pressure, humidity, temperature, and vibration.

Magnetic tape is an economical recording medium since it may be reused many times. It also provides a means for duplicating graphic recordings with photographic techniques.

46.5. A Shaft Position Digitizer System of High Precision

L. G. DEBEY, *Ballistic Measurements Lab., Aberdeen Proving Ground, Md.*, AND R. C. WEBB, *Colorado Research Corp., Denver, Colo.*

A new high-precision shaft angle digitizer system has been developed over a period of about three years under sponsorship of the Ballistic Research Laboratories of the Aberdeen Proving Ground, Md. It is comprised of a mechanical angular sensing head or transducer and a set of electronic digitizing apparatus providing both visual read-out as well as electrical signals. Design of the transducer is based upon the electrostatic tone wheel principle and involves the generation of two identical sinusoidal signals by motion of a continuously rotating internal member. Phase displacement of the two signals is representative of angular settings. Resolution is one part in 360,000 of the complete circle.

46.6. A High-Precision Digital Shaft Position Indicator

D. H. RAUDENBUSH, *Telecomputing Corp., North Hollywood, Calif.*

A unique system for the digital indication of a shaft position to a precision of six seconds of arc is discussed. A transducer, comprising a pair of dimensionally stable disks, one of which is movable with respect to the other, is attached to the shaft whose position is desired. Digital output of the system is presented by a register comprising a set of five novel cathode-ray tubes each having both a visual display and ten electrical outputs, one of which is energized at a time. Application of the system to an Askania Cinetheodolite range is described.

SESSION 47*

Thurs. 10:00 A.M.–12:30 P.M.

New York Coliseum
Marconi Hall

ANTENNAS II—GENERAL

Chairman: A. H. WAYNICK, *Elec. Eng. Dept., Pennsylvania State College, State College, Pa.*

47.1. Early Warning Radar Antennas

J. M. FLAHERTY AND E. KADAK, *Westinghouse Electric Corp., Pittsburgh, Pa.*

The current military requirements for the detection of missiles at ranges of 2000 to 3000 miles has led to the need for high-gain antennas at wavelengths of approximately one meter. Because of the unwieldy size of such antennas the authors have been investigating antenna types in which the beam directional control can be accomplished by switching the feed point rather than by moving the entire structure. Three main large-scale types have been considered: a flat "pancake" structure in which surface wave radiation is utilized, a luneberg lens structure utilizing artificial and loaded dielectrics, and a spherical structure employing a unique polarization relationship to provide a folded paraboloidal reflector effect.

47.2. Phase and Amplitude Measurements in the Near Field of Microwave Lenses

* Sponsored by the Professional Group on Instrumentation. To be published in Part 5 of the 1958 IRE NATIONAL CONVENTION RECORD.

* Sponsored by the Professional Group on Antennas and Propagation. To be published in Part 1 of the 1958 IRE NATIONAL CONVENTION RECORD.

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C. W. MORROW, P. E. TAYLOR, AND H. T. WARD, *Melpar, Inc., Falls Church, Va.*

An investigation has been made of the near-field phase and amplitude characteristics of several microwave antennas. From these measurements the phase center and focal point have been found. Calculations of the far-field pattern are based on the near-field measurements.

Particular data are shown for lenses having a constant index of refraction and for those constructed of a variable index medium. The utility of the measurements for evaluating and judging microwave lenses is pointed out.

47.3. Annular Slot Direction Finding Antenna

H. H. HOUGARDY AND N. YARU, *Hughes Aircraft Co., Culver City, Calif.*

The radiation characteristics of a flush-mounted cardioid-pattern direction-finding antenna are discussed. Using the general equations for the far field of an annular slot in an infinite and perfectly conducting ground plane, expressions are derived for the radiation field of an antenna consisting of a single annular slot fed with a cylindrical waveguide excited with two orthogonal TE_{11} modes and the TEM mode. The structure of the antenna and the method of exciting the desired modes is described. The effect of the cross-polarization component on the three-dimensional pattern is considered, and theoretical and experimental patterns are included for comparison.

47.4. A Novel Antenna for Mobile Radio Relay Operation in the UHF Range

F. J. TRIOLO, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

A novel type antenna called "double-rhombic," combining two rhombics whose planes form an angle with each other, has been designed and constructed for the frequency range 400 to 600 mc. The gain over a reference dipole was found to be in the order of 6 to 12 db. The swr including feed cable, was found to be less than 1.4. Some advantages are its low wind drag and light weight (25 pounds) in spite of its fairly large dimensions. Another advantage is that it can be easily assembled and disassembled thus making it adaptable for mobile radio relay application.

47.5. Lightweight, High-Gain Antenna

R. G. MALECH, *Airborne Instruments Lab., Inc., Mineola, N. Y.*

A novel high-gain, end-fire antenna has been developed for operation in the vhf range. The antenna is very long and has been designed for mounting on four telephone poles. The antenna consists of two ladder arrays mounted at right angles to each other on a single central boom. One ladder array covers a 20 per cent band, and the other covers the next higher 20 per cent band. Each array is about eight wavelengths long and has a gain of 18 to 20 db over its band.

The antenna for each band is energized by a dipole and reflector structure. A surface wave is launched by the feeding dipole along the modulated ladder structure. Both the spacings and the lengths of the elements are modulated to control the phase velocity of the surface wave along the array, making possible a long, high-gain antenna structure.

This paper discusses the model measurements performed at S band, and the mechanical aspects of the full-scale antenna.

47.6. Voltage Breakdown Characteristics of Microwave Antennas

J. B. CHOWN, T. MORITA, AND W. E. SCHARFMAN, *Stanford Research Institute, Menlo Park, Calif.*

At low pressures, antennas are susceptible to voltage breakdown. In the case of missiles, for example, there are indications that very low power is sufficient to initiate and maintain breakdown.

Since it is essential that the system performance is not interrupted for high-altitude operation, an experimental investigation has been made to study the properties of antennas under breakdown conditions. When voltage breakdown occurs, the effect is fourfold: the input impedance is altered, the total radiated power is decreased, the radiation pattern is changed, and the pulse shape is distorted. The effect on these properties of input power level, pulse width, pulse repetition rate, and pressure is discussed.

SESSION 48*

Thurs. 10:00 A.M.–12:30 P.M.

New York Coliseum
Faraday Hall

MICROWAVE TUBES

Chairman: L. S. NERGAARD, *RCA Labs., Princeton, N. J.*

48.1. Noise Characteristics of A Backward-Wave Oscillator

J. B. CICCHETTI AND J. MUNUSHIAN, *Hughes Aircraft Co., Culver City, Calif.*

Problems associated with the representation of signal and noise in oscillators are discussed. Various methods for the measurement of oscillator stability and for the time-domain characterization of oscillator signals are described; the usefulness of each method is interpreted in terms of particular oscillator applications. The method of models as described by Middleton and Gottschalk is used to measure the noise characteristics of a typical backward-wave oscillator. The interpretation of these results indicates that the output signal of a backward-wave oscillator may be represented by a cw carrier simultaneously amplitude and frequency modulated by partially correlated noise. The various measured fluctuating characteristics of both the signal and the beam are presented as functions of the operating characteristics of the tube.

* Sponsored by the Professional Group on Electron Devices. To be published in Part 3 of the 1958 IRE NATIONAL CONVENTION RECORD.

48.2. The Pulsed M-Type Backward Wave Oscillator and Its Modes of Operation

G. KLEIN AND A. L. WINTERS, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

The pulsed M-type backward wave oscillator is a high-power, high-efficiency, electronically tunable microwave generator which offers a possible means for the development of highly sophisticated electronic systems. This paper is a report on the results obtained to date with the CMI 150, 151, and 152 *S*-band pulsed tubes developed by the Compagnie Generale de Telegraphie Sans Fils, Paris, France, under U. S. Army Signal Engineering Laboratories sponsorship.

Compromises required in the design of a tube for operation at 150 kw in *S* band lead to relatively high electrode potentials and large current densities. Analysis of the performance characteristics of a typical tube points to a number of problems in the use of this device. Various methods of operation are possible, each presenting a unique set of problems. Data obtained from operation of these tubes by some of these methods are presented and analyzed

48.3. The Estiatron—An Electrostatically Focused Medium-Power Traveling-Wave Amplifier

D. J. BLATTNER AND F. E. VACCARO, *Radio Corp. of America, Princeton, N. J.*

This paper describes a developmental medium-power S-band traveling-wave tube which employs bifilar helices to focus a 50-ma beam electrostatically. With this tube, a beam-transmission efficiency of greater than 97 per cent under saturated rf conditions, a power output of 10 watts, and a gain of 20 db have been attained over the frequency range from 2000 to 3500 mc per second. The tube, which is self-focusing, weighs only 8 ounces, including rf couplers and attenuator. This weight represents a seven-fold reduction from that of an equivalent tube employing periodic permanent-magnet focusing. Focusing performance and rf characteristics of the estiatron are discussed in detail.

48.4. The Generation of Shaped Pulses Using Microwave Klystrons

D. H. PREIST, *Eitel-McCullough, Inc., San Bruno, Calif.*

This paper deals with a problem of current interest because of the need for pulsed signals with minimum possible sideband energy in air navigation systems such as Tacan.

The most desirable pulse shape is arrived at from a brief analysis of the fundamental problem. Such pulses can be generated in a reliable and practical way at radio frequencies of around 1000 mc, for example, by the use of a modulating anode klystron final amplifier and a special modulating circuit for applying the appropriate waveforms to the modulating anode of the klystron.

48.5. Wide-Band UHF 10-KW Klystron Amplifier

H. GOLDMAN, L. F. GRAY, AND L. POLLACK, *Federal Telecommunication Labs., Nutley, N. J.*

A power amplifier operating in the 690–890-mc band is described. The amplifier is capable of 20-mc, 3-db bandwidth with a gain of 30 db.

The reasons for the choice of a six cavity klystron are outlined. The external tuned circuits used across the interaction gaps are described and the relationship of the tuned circuit configuration to klystron theory, particularly with respect to broad-band operation of the klystron, is described.

The amplifier circuitry and mechanical arrangement are outlined in detail.

Typical test results with respect to frequency, such as amplitude response, phase response, and gain are discussed.

SESSION 49*

Thurs. 2:30–5:00 P.M.

Waldorf-Astoria Starlight Roof

GENERAL SYSTEMS

Chairman: K. E. IVERSON, *Computation Lab., Harvard University, Cambridge, Mass.*

49.1. Combat Computers

W. F. LUEBBERT, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

Multiple use electronic computers now under development for use by Army units in the field will make possible significant improvements in the over-all effectiveness of combat operations. This paper describes equipment design and development requirements, the characteristics of some equipments under development, others planned for development, and the equipment family concept for their integration. Classification of areas of application, some of the analyses undertaken, and the unusual characteristics of the information retrieval equipment under development are discussed.

49.2. The USAF Automatic Language Translator, Mark I

G. A. SHINER, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

* Sponsored by the Professional Group on Electronic Computers. To be published in Part 4 of the 1958 IRE NATIONAL CONVENTION RECORD.

The present Mark I is built to translate Russian into English. The bilingual dictionary contains about 40,000 words in their inflected form. Physically, the dictionary is a 10-inch glass disk carrying 30 million bits in 700 concentric tracks in 0.36 inch wide annulus. The disk's rotation speed is 1200 rpm. The output current from the photomultiplier follows the succession of black and white squares. This signal is fed into computer-type circuits, and compared with words in the input register. Upon obtaining a match, the subsequent information is delivered to output register. Here, information is temporarily stored until delivery to high-speed printer.

49.3. Nonbinary Switching Theory

O. LOWENSCHUSS, *Sperry Gyroscope Co., Div. of Sperry Rand Corp., Great Neck, N. Y.*

Present-day electronic digital computers consist basically of *binary* elements. Switching devices with *more than two* distinct states make possible the design of a nonbinary digital computer, which may exhibit certain advantages over its binary predecessor. Various nonbinary algebras are available. Each of these has certain advantages. Only "functionally complete" algebras are useful in circuit design. Special algebras can be formed for nonbinary devices, and can be used to express any function of many variables, given in terms of a nonbinary truth table. The design of combinational and sequential circuits is discussed, using the Rutz transistor as an example. Nonbinary algebras also can be used to analyze the transition between states in binary and nonbinary devices.

49.4. Automatic Type Size Normalization in High-Speed Character Sensing Equipment

A. I. TERSOFF, *Intelligent Machines Research Corp., Alexandria, Va.*

A high-speed character sensing system is described which makes use of the length and within-character location of character strokes for purposes of analysis and identification. To permit the reading of documents prepared in a multiplicity of type sizes, a special circuit is employed to provide a measuring standard which automatically proportions itself to the type size being read. Measurements made therefore are relative ones, and are essentially independent of type size. A single associated logical program can then be applied to each document to correctly analyze the characters scanned despite size variations.

49.5. Minimum Time Programming on a Drum Computer

B. SHIFFMAN, *The Ramo-Wooldridge Corp., Los Angeles, Calif.*

In a general purpose, stored program, magnetic drum, high-speed, digital control computer, it is possible, by a suitably chosen bookkeeping system, to achieve minimum time programming. This implies continuous computing with no time wasted in locating operands. The bookkeeping system keeps a tally of the word times, or sectors on the drum, available for any particular command and its operands. Thus, the accounting which progresses as the program unfolds determines the storage locations of the commands and the constants and assures compatibility of the temporary storage. Programs have been written with a computing speed 95 per cent as fast as that of a random access memory machine with the same command instruction and execution time.

SESSION 50*

Thurs. 2:30–5:00 P.M.

Waldorf-Astoria Astor Gallery

CIRCUIT THEORY III—APPLICATION OF TOPOLOGICAL AND GROUP CONCEPTS

Chairman: M. E. VAN VALKENBURG, *University of Illinois, Urbana, Ill.*

50.1. Signal Flow Graph and Network Topology

O. WING, *Dept. of Elec. Eng., Columbia University, New York, N. Y.*

The foundation of signal flow graph is established via network topology in terms of submatrices of the matrices generally encountered in topology. It is shown that the flow graph is completely specified by four such submatrices. As a consequence, the formulation of a signal flow graph is made extremely simple and always can be done by inspection. Moreover, it is now possible in some cases to determine the functions of a network directly from the network graph without constructing its signal flow graph and without using the method of tree products.

50.2. New Transpositions in Power Transformer Windings

R. G. DEBUDA, *Canadian General Electric Co., Ltd., Toronto, Ont., Can.*

Transpositions, as used in power transformer windings to ensure equal current division in parallel strands, have not been treated very much in the literature.

It is possible to classify the transposition systems by noting that they correspond to certain simple finite groups. When this was done, it turned out that two systems of transpositions had been overlooked.

This paper calculates, for different transposition systems, the extra losses due to unequal current sharing and compares the different systems on the basis of losses and mechanical construction.

It is shown that the two new systems of transpositions give better current sharing and therefore lower extra losses than a comparable system generally used, while there is little difference in the mechanical construction.

* Sponsored by the Professional Group on Circuit Theory. To be published in Part 2 of the 1958 IRE NATIONAL CONVENTION RECORD.

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50.3. Two-Terminal Pair Symmetry Relations

R. C. Kiessling, *Lenkurt Electric Co., San Carlos, Calif.*

This paper illustrates an application of the signal flow graph technique. Flow graphs are used to demonstrate relationships between the two-dimensional spaces describing a two-terminal pair to the four-dimensional space in which the network can appear. A general procedure for transforming any one matrix into any other matrix representation is described and applied to computer programming.

The structural frame established by the various directed line segments representing the six matrix forms used to describe the two-terminal pair is shown to form a rigid structure in three-dimensional space. The various operations, which rotate the cube about a major diagonal, or rotate the faces about an axis through the center of two opposite sides, form a permutation group of the four variables. Only one basic operation is required to relate all six sides of the cube to each other. All other transformations performed on the cube can be expressed in terms of this basic operation. The main body of the paper describes this basic operation. Two appendixes are attached. Appendix I is an example problem illustrating the use of the permutation operation. The problem is a simple block reduction of a circuit to a single block. Appendix II describes the properties of a "group" and demonstrates that the operation described forms a group.

50.4. Analysis of Nonreciprocal Networks by Digital Computer

W. Mayeda and M. E. Van Valkenburg, *Dept. of Elec. Eng., University of Illinois, Urbana, Ill.*

Percival and, more recently, Mason have given topological formulas for the computation of network functions for nonreciprocal networks. These formulas differ from those for the passive case in two respects: two graphs are involved, and the sign for the terms must be found. In the paper, the topological formulas are presented in a form convenient for digital computer programming. Features of the program are outlined. The advantages of speed and accuracy are illustrated by several examples which include stability and sensitivity considerations.

50.5. On Non-Series-Parallel Realization of Driving-Point Function

W. H. Kim, *Dept. of Elec. Eng., Columbia University, New York, N. Y.*

This paper presents the results of an investigation on the non-series-parallel network realization of minimum functions. The network chosen is the unbalanced bridge, which has the least number of elements among all non-series-parallel structures.

The following results were obtained:

1) At least three reactive elements are required to realize a minimum function without ideal transformer.
2) The subclass of the class of positive-real functions which can be realized in the chosen structure is found.
3) The network which realizes the minimum function is found to contain the minimum number of elements. Examples are given which show that this structure in fact requires less elements than Brune or the modified Bott-Duffin structures.

SESSION 51*

Thurs. 2:30–5:00 P.M.

Waldorf-Astoria Jade Room

ULTRASONICS III—MEASUREMENT OF RADIATED ACOUSTIC POWER

Chairman: O. E. Mattiat, *Aerophysics Development Corp., Santa Barbara, Calif.*

51.1. Power Handling Capability of Ferroelectric Ceramics

G. W. Renner, R. A. Plante, and T. F. Hueter, *Raytheon Manufacturing Co., Wayland, Mass.*

Titanates of various compositions are commonly used in sonar and commercial ultrasonic transducers. At the high drive levels required for power applications, certain limitations of ferroelectric materials have to be considered. They are related to the hysteresis properties inherent in such materials and to their thermal characteristics. An effort is made to determine the number of watts per unit volume that can be handled safely by ceramic transducers. This number depends on the frequency, the Q, and the coupling coefficient of the material, as well as the heat transfer characteristics which, in turn, are related to the shape and configuration of the transducer elements. Some results of an experimental investigation of these factors will be discussed. The instrumentation required for meaningful measurements of the linearity and power handling capacity of ceramic transducers will be described.

51.2. Measurement of Acoustic Power Radiated from Underwater Sound Transducers

R. J. Bobber, *Office of Naval Research, Orlando, Fla.*

Some of the measurement methods used in the evaluation of underwater sound transducers are potentially applicable to the evaluation of transducers used in ultrasonics engineering. Two methods for the measurement of radiated acoustic power of underwater sound transducers are described. The advantages, disadvantages, and limitations of each method are discussed. One method requires measurement of the sound pressure at one point in the medium and the directivity index of the transducer while it is radiating into a free field or a large medium. The second method requires measurement of the electrical impedance of the transducer under loaded and unloaded conditions. Typical data are presented.

* Sponsored by the Professional Group on Ultrasonics Engineering. To be published in Part 2 of the 1958 IRE National Convention Record.

51.3. An Instrument for Determining Intensity of Ultrasound

J. F. Herrick, B. H. Anderson, and M. Neher, *Mayo Clinic and Mayo Foundation, Rochester, Minn.*

A unique type of transducer assembly converts the motion of a small rigid sphere (the sonic radiation pressure detector) into a voltage difference that determines the current through a galvanometer. Only minimal displacements of the sphere from its null position are required for accurate determinations of the intensity of ultrasound, the maximal displacement being of the order of magnitude of 0.2 mm.

The rigid sphere is attached to a wire that is mounted in a restraining metal diaphragm. A vane attached to the other end of the wire intercepts a light beam passing from a suitable source to a sensitive "split" photovoltaic cell. Motion of the sphere results in motion of the vane which causes an unbalance in the output of the photocells and, hence, a deflection of the galvanometer.

51.4. Measurements of Acoustic Power in Industrial Ultrasonic Equipment

W. Welkowitz, *Gulton Industries, Inc., Metuchen, N. J.*

One of the problems that is of increasing importance in the design of industrial ultrasonic equipment is the measurement of acoustic power generated by the equipment. This paper deals with the measurement techniques for various industrial devices. Calibration techniques for ceramic probe microphones are presented. The methods of measuring power in noncavitating liquids and in horn type transducers (for drills and welders) are discussed. Finally, an instrument to measure cavitation in liquids is described.

51.5. Panel Discussion—Problems in Power Measurement

Panel Members: G. E. Henry, *General Electric Eng. Lab., General Electric Co., Schenectady, N. Y.*

S. E. Jacke, *Detrex Corp., Detroit, Mich.*

F. Massa, *Massa Labs., Inc., Hingham, Mass.*

M. Strasberg, *David Taylor Model Basin, Washington, D. C.*

As applications for ultrasonic energy are found in more diverse fields, many problems arise in the measurement of the power radiated by the transducers. Knowledge of the acoustic energy present in the sound field is

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important to both the user and the manufacturer of ultrasonic equipment. As an example, one current problem is the determination of the intensity of ultrasound in a confined vessel, such as an ultrasonic cleaning tank. Such problems and their possible solutions will be discussed.

SESSION 52*

Thurs. 2:30–5:00 P.M.

Waldorf-Astoria
Sert Room

LONG DISTANCE COMMUNICATIONS

Chairman: A. G. CLAVIER, *Federal Telecommunication Labs., Nutley, N. J.*

52.1. Single-Channel Radioteletype Communication

H. B. VOELCKER, JR., *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

The evolution of long-range, single-channel radioteletype systems is traced briefly from the viewpoint of modulation and detection efficiency. The desirability of linear, synchronous detection is illustrated, and two methods for achieving this objective are considered. The first is the phase-modulated Kineplex technique of Collins Radio Company, and the second is a Signal Corps project which utilizes phase-coherent fsk modulation. The latter system represents an elementary attempt to measure certain ionospheric characteristics and to utilize them in an optimum, linear detection process in accordance with the analytical results of Turin and Price.

52.2. A World-Wide High-Frequency SSB Radio Network

E. BRAY, *Collins Radio Co., Cedar Rapids, Iowa*

This paper describes a high-frequency communication network designed to provide communications between stations located at various points throughout the world; the use of SSB transmission, high-gain antennas, proper selection of operating frequencies for optimum propagation, and radio relaying to provide alternate routing of messages results in successful communication at nearly all times. Performance predictions based on propagation studies are discussed, and results of actual network operation are presented and compared with predicted performance.

52.3. Comparison of Multichannel Radioteletype Systems over a 5000-Mile Ionospheric Path

A. T. BRENNAN, *Stromberg-Carlson Co., Rochester, N. Y.*, B. GOLDBERG AND A. ECKSTEIN, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

A three-month operational comparison is made between the Collins Kineplex radioteletype system and the standard military fsk radioteletype terminal, the AN/FGC-29. This comparison is made under many propagation conditions, including selective frequency fading, flat fading, and impulse noise.

A radio link in the 10–20-mc range is used, with the transmitting equipment in the Territory of Hawaii and the receiving terminals at Fort Monmouth, N. J.

The comparison is made on the basis of the relative number of errors obtained when equal transmitter power per channel is used by each system. Simultaneous transmissions are made on twelve channels of each system, copy being taken on twenty-four teleprinters.

Instrumentation includes means of measuring signal-to-noise ratios, degree of multipath, and selective fading characteristics. These are related to the muf which existed during each test period.

Results are presented on curves which relate the signal-to-noise ratio measured at the receiver site to the normalized error rate calculated from the teletype copy. Individual curves describe system performance under several combinations of propagation conditions.

52.4. Basic Analysis on Controlled Carrier Operation of Tropospheric Scatter Communications Systems

L. P. YEH, *Westinghouse Elec. Corp., Baltimore, Md.*

This paper analyzes the possibility of applying controlled carrier operation to tropospheric scatter communication systems which means that the transmitter power will change as closely as possible in accordance with the fluctuation of the received signal level with time.

Basic requirements, mainly in the field of propagation, are discussed in detail. Certain advanced thinking in tropospheric scatter propagation is also presented, such as:

1) One *minute* as the sampling period of Rayleigh Distribution Fast Fading,
2) Long-term *minutely* median distribution as the long-term slow fading distribution,
3) Probability combination of fast and slow fading distributions as the combining fading distribution to determine system reliability, and
4) Effect of AGC detector time constant on the fast fading distribution.

52.5. Transportable Tropospheric Scatter Communications Systems

A. J. SVIEN, *Collins Radio Co., Tucson, Ariz.*, AND J. C. DOMINGUE, *Fort Huachuca, Ariz.*

This paper describes transportable uhf scatter communications systems and details performance records of these systems over a large number of circuits varying in length from 50 to 150 miles with representative samples of intermediate terrain and horizon angles.

Two versions of the systems are described: one van-mounted, the other mounted in shelters suitable for either truck or air transportation. The latter units feature air inflatable 15-foot diameter paraboloidal reflectors constructed in such a manner that no radome protectors are required to assure unimpaired performance under high wind loading. The unique construction of this antenna permits packaging of a 1-kw fm transmitter, dual-diversity receivers, and associated antennas in two shelters 76×76×96 inches.

52.6. Evaluation of IF and Baseband Diversity Combining Receivers

R. T. ADAMS AND B. M. MINDES, *Federal Telecommunication Labs., Nutley, N. J.*

A comparative study of the relative performance of an fm dual diversity receiver with either ratio-squarer post-detection or linear-addition predetection combining is presented. An examination of the distortion produced by multipath propagation in an fm receiver using the two combiners occupies the major portion of the study. The problem was treated experimentally and theoretically with good agreement found between the two approaches. Other criteria considered are threshold performance, signal-to-noise ratio, and physical characteristics (size, weight, complexity, etc.). The methods of operation of the two combiners also are outlined. The results appeared to favor the predetection combiner.

52.7. Transmission of Digital Data over Multihop Tropospheric Scatter Links

C. N. LAWRENCE AND R. L. MARKS, *Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y.*

This paper summarizes experience in transmission of digital data over certain tropospheric scatter links operated by the Air Force. Error count vs bandwidth and noise level is considered, together with some methods of error rejection.

SESSION 53*

Thurs. 2:30–5:00 P.M.

New York Coliseum
Morse Hall

HIGH-ACCURACY INSTRUMENTS, MEASUREMENT, AND CALIBRATION

Chairman: F. G. MARBLE, *Boonton Radio Corp., Boonton, N. J.*

* Sponsored by the Professional Group on Communications Systems. To be published in Part 8 of the 1958 IRE NATIONAL CONVENTION RECORD.

* Sponsored by the Professional Group on Instrumentation. To be published in Part 5 of the 1958 IRE NATIONAL CONVENTION RECORD.

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53.1. A Feedback Amplifier with Negative Output Resistance for Magnetic Measurements

W. P. Harris and I. L. Cooter, *National Bureau of Standards, Washington, D. C.*

Accurate core loss measurements at high-flux densities can be made by bridge methods if the power dissipated in the primary circuit at harmonic frequencies is measured and subtracted from the apparent power dissipated in the ferromagnetic material at fundamental frequency. The determination of this harmonic power term is inconvenient and must be done with greater accuracy than that required in the final result. An amplifier having negative output resistance was devised and is used in a manner which automatically allows accurate compensation for the harmonic power dissipation.

53.2. Millimicrosecond, Wide-Aperture, Electro-Optical Shutter

J. A. Hull, *Avco Manufacturing Corp., Lawrence, Mass.*

A transmission line pulse generator has been employed to drive a Kerr cell electro-optical shutter having an aperture of 2×2 inches. Reflected light photographs on Polaroid type 400 film have been obtained with this shutter at an exposure time of 10^{-8} seconds. The Kerr cell and pulse generator are packaged in a sealed unit which is 12×12×5 inches. The shutter requires an external power supply of 0–50 kv and a low-voltage trigger source. The pulse generator is actuated by a specially designed spark-gap which has a jitter of less than 0.1 μsec. The spark-gap may be modified to provide a pulsed, point light source which is fully synchronized with the opening of the Kerr cell. This combination is particularly useful in shadowgraph or Schlieren photographic stations. A technique for producing multiple exposures with a single Kerr cell using the above pulse generator has been accomplished. The Kerr cell shutter will produce high-resolution, high-speed photographs which are free from optical distortion.

53.3. A Quartz Servo Oscillator

N. Lea, *Marconi's Wireless Telegraph Co., Ltd., Chelmsford, Eng.*

The paper covers development of 5-mc oscillators with instabilities due to tube and circuit effects about 100 times less than in conventional crystal oscillators of the same frequency.

A single precision crystal gives independent but simultaneous stabilization of frequency by high Q resonant loop and by servobalance of an rf crystal bridge.

The servosystem has a correction sensitivity of 3 in 10^{11} and the oscillator an over-all weekly instability of 2 in 10^{10}.

The oscillator has outputs at 10, 1.0, and 0.1 mc and is fitted with cro monitor to facilitate servicing.

53.4. A New Method to Simplify Bridge Type Measurements on Quartz Crystal Units

E. Hafner, *U. S. Army Signal Eng. Labs., Fort Monmouth, N. J.*

The proper evaluation of the performance characteristics of quartz crystal units as used for frequency stabilization purposes and filter applications becomes increasingly difficult as the frequency of operation extends much beyond 50 mc. To facilitate bridge type measurements, which are inherently capable of giving the most complete information in a rather direct way, a method was devised whereby the information contained in the output signal from the bridge is used to control automatically the frequency of the generator feeding the bridge. This at once eliminates the need to continually adjust the generator frequency and leaves only one of the two bridge controls as independent variable. The control of the generator frequency is so effective that stabilities of a few parts in 10^9 were observed, even at 200 mc, over extended periods of time using an average variable frequency generator.

53.5. RF-Voltage Calibration Consoles

M. C. Selby, L. F. Behrent, and F. X. Ries, *National Bureau of Standards Boulder Labs., Boulder, Colo.*

Consoles to calibrate rf voltmeters for science and industry were developed at the National Bureau of Standards Boulder Laboratories. Very accurate calibrations will be performed in a fraction of the time heretofore required at the Bureau at any practical voltage level starting with 0.2 volt of twelve discrete frequencies between 30 kc and 700 mc. New type AT voltmeters, the most stable rf-voltage reference standards known to date, are used. These voltmeters can reproduce calibration data to ±1 per cent or better over a period of one year or longer. The major components, in addition to the reference voltmeters, are extremely stable rf sources, automatic and manual level controls, and protective and indicating circuits.

SESSION 54*

Thurs. 2:30–5:00 P.M.

New York Coliseum
Marconi Hall

ANTENNAS III—MICROWAVE ANTENNAS

Chairman: L. C. Van Atta, *Hughes Aircraft Co., Culver City, Calif.*

54.1. A Compact Dual-Purpose S-Band Beacon and VHF Telemetry Antenna

W. O. Puro, W. G. Scott, and W. A. Meyer, *Melpar, Inc., Falls Church, Va.*

* Sponsored by the Professional Group on Antennas and Propagation. To be published in Part 1 of the 1958 IRE National Convention Record.

The design and performance of a novel radiating element, capable of efficient performance at two frequencies separated by approximately four octaves, are described. The element is a rectangular cavity one-quarter wavelength long and one-tenth wavelength in cross section at the lowest operating frequency. One open face provides the radiating aperture. The element is capable of being used in many different flush-mounted configurations, either as an element of an array or singly, depending upon the desired pattern characteristics.

The element is designed for extreme mechanical, thermal, and low-pressure environmental conditions typical of many airborne vehicle conditions. Radiation patterns and impedance match data are also presented.

54.2. A Volumetric Electrically Scanned Two-Dimensional Microwave Antenna Array

J. L. Spradley, *Hughes Aircraft Co., Culver City, Calif.*

By proper interelement phase control, the beam of a stationary two-dimensional antenna array can be scanned throughout a volume in space. This is demonstrated with an array of radiating X-band waveguides fed by a unique arrangement which permits independent phase control of every element, as well as frequency scanning along one dimension of the array. Volumetric scanning is achieved by various combinations of these phase shifting schemes. The measured beam position corresponds very favorably with theory for any given phase distribution. The only adverse side lobes appearing in any of the measured radiation patterns correspond to systematic errors caused by mutual coupling. Methods for suppressing these systematic errors are discussed in detail and corroborated by encouraging experimental results.

54.3. Closely Spaced Polyrod Arrays

L. W. Mickey and G. G. Chadwick, *Melpar, Inc., Falls Church, Va.*

In an effort to reduce the existing low side lobes of microwave arrays an investigation of the element pattern influence was conducted. It was estimated that side lobe levels, beyond the first lobe region, of 50 db below the main beam gain could be achieved if end-fire elements with half-power beamwidths of 20° magnitude were employed. An additional consideration was that interelement spacings on the order of a half wavelength were required.

From previous study programs conducted elsewhere, it was known that end-fire elements suffered from severe pattern deterioration when mounted in closely spaced arrays. Experimental work therefore was concentrated on high dielectric constant materials in which the field may be concentrated about a radiator of extremely small head-on cross section. Experimental data on polyrods of dielectric constants ranging from 8.6 to 25.0 indicated that the basic end-fire characteristic changed only slightly. Furthermore, any noted changes led to the improvement of the element pattern. Data indicating the normalized resonant conductance of a rod in the narrow face of a standard X-band waveguide were obtained. This information allows the design of high-dielectric constant polyrod arrays using standard techniques, once the desired aperture distribution is selected.

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54.4. Waveguide Loaded Surface Wave Antenna

R. F. Hyneman and R. W. Hougardy, *Hughes Aircraft Co., Culver City, Calif.*

A new type of surface wave structure is considered in which the required reactive surface impedance is provided by a parallel array of air-filled waveguides with narrow walls adjacent. Each guide has a broad wall containing a series ot closely-spaced, nonresonant, transverse slots. By choosing an operating frequency such that the broad dimension of each guide is less than one-half free-space wavelength, all propagating modes are of the surface wave type.

An approximate solution for the propagation constant, γ, of the dominant mode has been obtained by means of a variational technique. Experimental results are shown to be in reasonable agreement with the theory.

A related application involving the use of this principle in the dispersive loading of slotted waveguide antennas also is considered. Results are compared with those obtained with a conventional corrugated waveguide.

54.5. Dielectric Image Line Surface Wave Antenna

H. W. Cooper, M. Hoffman, and S. Isaacson, *Maryland Electronic Manufacturing Corp., College Park, Md.*

This paper describes a new antenna medium consisting of a two-dimensional array of resonant slots in a ground plane. The slots are excited by a surface wave guided over a ground plane by a thin ribbon of dielectric called the image line. Those properties of the array were investigated which are necessary for array design, including control of relative amplitude and phase, both along and transverse to the transmission line, and efficient utilization of the power in the mode by the array. The antenna is flush mountable, inexpensive, and can be made to have outstanding logistic capabilities.

54.6. A Dual Beam Planar Antenna for Janus Type Doppler Navigation Systems

H. Saltzman and G. Stavis, *General Precision Lab., Inc., Pleasantville, N. Y.*

A two-dimensional or planar dual beam antenna consisting of slotted waveguide linear arrays has many characteristics which are applicable to Janus type, or self-coherent, Doppler Navigational Systems. The dual beams are formed by a proper choice of spacing and phasing of the slotted radiators in each linear array. By a unique combination of the linear arrays, lobe switching can be accomplished very simply with an rf switch and hybrid. One of the advantages in system application of this antenna is the independence of the Doppler return from transmitter frequency variations.

SESSION 55*

Thurs. 2:30–5:00 P.M.

New York Coliseum
Faraday Hall

RADIO AND TELEVISION

Chairman: D. D. Israel, *Emerson Radio & Phonograph Corp., Jersey City, N. J.*

55.1. Design Problems in Transformerless Single Rectifier Television Receivers

D. Sillman, *Westinghouse Electric Corp., Metuchen, N. J.*

The use of transformerless single rectifier power supplies permits cost savings in television receivers, but also introduces problems in maintaining good receiver performance. The principal difficulties lie in obtaining adequate sweep and video signal amplitudes. The paper describes a 21-inch 90° receiver with such a power supply and the components and circuits used to overcome the difficulties.

55.2. Problems in Two-Dimensional Television Systems

R. M. Bowie, *Sylvania Electric Products Inc., Bayside, N. Y.*

The rise of electroluminescence has posed the intriguing prospect of a flat, two-dimensional display of pictorial information and eventually of television. The Sylvatron, the light amplifier, and similar devices already have demonstrated the feasibility by electroluminescent means. To extend these to greater detail and higher speed will require advances in methods of scan, light modulation, and in the attainment of adequate brightness such as by the storage principle. It is concluded that some major breakthroughs still are needed.

55.3. A New High-Power Horizontal-Output Tube and Deflection System for Color Television

J. P. Wolff and R. G. Rauth, *Radio Corp. of America, Harrison, N. J.*

This paper describes a developmental high-power horizontal-deflection and high-voltage system for color television receivers, which employs the RCA-6DQ5, a new horizontal-output tube having unique design features. This system is analyzed specifically in relation to the ratings and electrical characteristics of the output tube such as cutoff and zero-bias plate current. Consideration is given to circuit parameters such as drive, scan, and power output. The load line of a typical operating system is discussed and the methods used to measure some of the tube parameters are briefly mentioned. For purposes of illustration, the over-all performance of an experimental complete operating high-power (25-watt) color-television horizontal-deflection system is examined.

* Sponsored by the Professional Group on Broadcast and Television Receivers. To be published in Part 7 of the 1958 IRE National Convention Record.

55.4. Improvements in Deflection Amplifier Design

C. Droppa, *Sylvania Electric Products Inc., Emporium, Pa.*

Extensive tv receiver life-test studies made by Sylvania's engineering laboratories show that the horizontal-deflection amplifier socket is one of the areas offering most room for improvement in life expectancy.

Some of the receiving tube improvements which would help increase the life expectancy of this socket include better screen dissipation capability to preclude screen emission, high plate-to-screen current ratio to insure high-scan efficiency and reduce screen dissipation, good uniformity of cutoff to minimize variations in drive requirements, low-knee curves to preclude the generation of "snivets," and a low microstructure to preclude "venetian-blind" effects.

A radical new approach to receiving tube design and manufacture appeared mandatory if all the above mentioned objectives were to be encompassed.

Sylvania's engineers adapted devices, heretofore only used on special higher cost tube types, to high-volume production techniques, and designed receiving tube types of competitive price for horizontal-deflection service. Details of one such design are described, and the manner in which this radical new tube structure overcomes current problems and actual test results attained with this tube in receivers are stated.

55.5. AGC Design Considerations for TV Receivers

R. H. Overdeer, *Philco Corp., Philadelphia, Pa.*

An AGC circuit cannot be isolated from the remaining circuitry in the tv receiver. The tuner and IF design, as well as the video and sync separator circuits, determine the environment into which the AGC system must be integrated. The signal conditions to which the set will be subjected impose another set of requirements. The price range of the set is still another restricting factor.

Starting with these boundary conditions numerous solutions have evolved over a period of years. The advantages and disadvantages of several systems will be discussed and the Philco solution will be explained in detail.

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IRE News and Radio Notes

Calendar of Coming Events and Authors' Deadlines*

1958

Nuclear Eng. and Science Congress, Palmer House, Chicago, Ill., Mar. 16–21

IRE Nat'l Convention, N. Y. Coliseum and Waldorf-Astoria Hotel, New York City, Mar. 24–27

Instruments & Regulators Conf., Univ. of Del., Newark, Del., March 31–Apr. 2

Conf. on Automatic Optimization, Univ. of Del., Newark, Del., Apr. 2–4

Symp. on Electronic Waveguides, Eng. Soc. Bldg., New York City, Apr. 8–10

SW Regional Conf. & Show, Mun. Audit., and St. Anthony Hotel, San Antonio, Tex., Apr. 10–12

Conf. on Automatic Techniques, Statler Hotel, Detroit, Mich., Apr. 14–16

Spring Tech. Conf. on TV and Transistors, Eng. Soc. Bldg., Cincinnati, Ohio, Apr. 18–19

Conf. on Radio Noise Spectrum, Harvard College Observatory, Cambridge, Mass., Apr. 22

Elec. Components Symp., Ambassador Hotel, Los Angeles, Calif., Apr. 22–24

URSI Spring Mtg., Willard Hotel, Wash., D. C., Apr. 24–26

Semiconductor Symposium of Electrochemical Society, Statler Hotel, New York City, Apr. 30–May 1

Seventh Region Conf. & Show, Sacramento, Calif., Apr. 30–May 2

PGMTT Symp., Stanford Univ., Stanford, Calif., May 5–7

Western Joint Computer Conf., Ambassador Hotel, Los Angeles, Calif., May 6–8

Nat'l Aero. & Nav. Elec. Conf., Dayton, Ohio, May 12–14

IEE Convention on Microwave Values, Savoy Place, London, England, May, 10–23

PGME Mtg., Wash., D. C., May

PGPT Symp., Hotel New Yorker, New York City, June 5–6

PGMIL Convention, Sheraton-Park Hotel, Wash., D. C., June 16–18

Spec. Tech. Conf. on Nonlinear Mag. and Mag. Amplifiers, Los Angeles, Calif., Aug. 6–8

Elec. Radio & Standards Conf., Univ. of Colo. and NBS, Boulder, Colo., Aug. 13–15

WESCON, Ambassador Hotel and Pan-Pacific Audit., Los Angeles, Calif., Aug. 19–22 (DL*: May 1, R. C. Hansen, Microwave Lab., Hughes Aircraft Co., Culver City, Calif.)

Int'l Conf. for Analog Computations, Strasbourg, Germany, Sept. 1–7

Nat'l Symp. on Telemetering, Americana Hotel, Miami Beach, Fla., Sept. 22–24

Indus. Elec. Conf., Rackham Mem. Audit., Detroit, Mich., Sept. 24–25

* DL = Deadline for submitting abstracts.

(Continued on page 659)

Joint Automatic Techniques Conference Set for Detroit

The first Joint Conference on Automatic Control Techniques is scheduled to convene in Detroit, Mich. at the Hotel Statler April 14–16, 1958. Under the co-direction of G. W. Heumann, Chairman, AIEE Joint Division Committee on Automation and Data Processing, and T. H. Belcher, Michigan Section AIEE, the forum will include four technical sessions, a panel discussion and an inspection trip.

The meeting will bring together, for the first time, members of the principal professional engineering societies who are making use of electrical, electronic, mechanical, hydraulic and pneumatic media to solve automatic control and data handling problems for industrial production. During the three-day meeting nineteen papers in three categories: automatic control and test techniques for flow process, manufacturing and utility operations; engineering, design and business data computation and processing; and utilization of control components will be delivered.

The program committee is composed of W. R. Harris, AIEE, Chairman; F. D. Snyder, ASME; C. A. Priest, IRE; G. W. Knapp, AIEE; and A. R. Satullo, AIEE, Michigan Section.

TV Conference Lists Chairmen

The annual Spring Technical Conference on Television and Transistors, for the twelfth year, will be held on April 18–19, at the Engineering Society Building, 1349 E. McMillan St., Cincinnati, Ohio. Information concerning advance registration or hotel reservations can be obtained from E. J. Emmerling, Cincinnati Gas & Electric Co., Main and 4th Sts., Cincinnati 2, Ohio. In charge of papers for the conference is J. R. Ebbeler, Avco Mfg. Corp., Crosley Div., 1329 Arlington St., Cincinnati 25, Ohio. Advertising and exhibition privileges are obtainable upon request from C. F. Winder, Baldwin Piano Co., 1801 Gilbert Ave., Cincinnati 25, Ohio.

WESCON Papers Deadline Set for May 1

Authors wishing to present papers at the 1958 WESCON Convention to be held in Los Angeles August 19–22 should send 100-word abstracts, and either the complete text or a detailed summary to the Technical Program Committee Chairman:

Dr. Robert C. Hansen
Microwave Laboratory
Hughes Aircraft Co.
Culver City, Calif.

There will again be a WESCON Convention Record. Authors will be notified of acceptance or rejection by June 1.

IRE Appoints 1958 Officers

The IRE Board of Directors, at its January meeting, appointed six members to the Board for 1958.

Reappointed as Treasurer of the IRE was W. R. G. Baker, Vice-President for Research of Syracuse University, Syracuse, N. Y. and former Vice-President of Electronics, General Electric Company. Haraden Pratt was appointed to his sixteenth term as IRE Secretary. J. D. Ryder, Dean of Engineering, Michigan State University, East Lansing, Mich., was appointed Editor of the IRE to succeed Donald G. Fink, Director of Research of the Philco Corp.

Appointed as Directors were A. N. Goldsmith, Consulting Engineer and Editor Emeritus of the IRE; D. B. Sinclair, Vice-President of Engineering, General Radio Corporation, Cambridge, Mass.; and Ernst Weber, President of the Polytechnic Institute of Brooklyn and President of Polytechnic Research and Development Company, Brooklyn, N. Y.

Additions to IRE Approved

On January 6, the Executive Committee approved the establishment of the Santa Ana Subsection of the Los Angeles Section.

On the following day, the IRE Board of Directors approved the formation of the Erie and Western Michigan Sections. The establishment of the Colombian Section was also approved. In addition to Colombia, IRE Sections are now operating in Tokyo, Buenos Aires, Rio de Janeiro, Israel, Egypt, twelve Canadian cities, and Hawaii. This makes a total of nineteen IRE Sections operating outside the continental limits of the United States.

Three Speakers Will Appear at Elec. Components Meeting

Three luncheon and dinner speakers for the 1958 Electronic Components Conference to be held April 22–24 at the Ambassador Hotel, Los Angeles, Calif., have been announced by E. E. Brewer, technical program chairman. The theme of the conference is "Reliable Application of Component Parts."

Lt. General C. S. Irvine, U. S. Air Force Materiel Department, will discuss the economic aspects of electronic component selection on April 24. J. M. Bridges, Director of Electronics for the Office of the Assistant Secretary of Defense, will speak on certain aspects of electronic component reliability and usage on April 22.

The principal speaker at the conference dinner meeting April 23 will be A. M. Zarem, president of Electro-Optical Systems, Inc., Pasadena, Calif., who will speak on "Whither Away? . . . Or . . . Wither Away!"

Session chairmen for the Electronic Components Conference include C. G. Killen, Sprague Electric Co.; W. H. Forster, Philco Corporation; K. F. McCready, Northrop Aircraft, Inc.; and C. E. Coon, Tung Sol Electric Co.

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WJCC Lists Panel Discussions

The first national symposium on modern computer design methods and application techniques, at which authorities in the field will argue the relative merits of various systems and components, will take place beginning May 6 at the Ambassador Hotel, Los Angeles, Calif.

Highlight of the 1958 Western Joint Computer Conference, of which W. H. Ware is chairman, the unusual panel discussion, under the general theme of "Contrasts in Computers," will devote three full days to an exhaustive debate of six controversial subjects in the field of computer design and application.

Among the topics scheduled for discussion are "Logical Design Methods," with Gene Amdahl as chairman, in which block diagrams and logical equations will be contrasted; "Active Elements for the Machine," with Ralph Meagher as chairman, and four experts arguing the case for magnetic cores, crytrons, transistors and special vacuum tubes; "Logical Circuitry for Transistor Computers," with J. H. Felker as chairman, in which the comparative values of resistor-transistor, direct-coupled, symmetrical and STRETCH logic will be presented; and "Solution of Ordinary Differential Equations," with Wesley Melahn as chairman, which will consider the relative advantages, in this respect, of general-purpose computers, analog computers, combination digital and analog, and digital differential analyzers. Also on the agenda will be sessions on "Command Structure" and "Very Large Files," presided over respectively by J. W. Carr and A. J. Perlis.

The 1958 Western Joint Computer Conference is co-sponsored by the IRE, the American Institute of Electrical Engineers and the Association for Computing Machinery. In addition to the panel discussions, individual papers will be presented during the course of the conference, scheduled May 6–8. In addition, the following day, May 9, will be given over to "Small Computers—A Report from the Manufacturers," sponsored by the Los Angeles Chapter of the ACM.

IRE Picks Hayes Medal Winner

B. F. C. Cooper (M'47) has been awarded the 1957 Norman W. V. Hayes Memorial Medal for the most meritorious paper published during 1956 in the *Proceedings of the Institution of Radio Engineers Australia*. The IRE acted as adjudicators this year.

Mr. Cooper's paper was entitled "The Application of Transistors to AM Broadcast Receivers" and was published in the October, 1956 issue of the *Proceedings of the I.R.E Australia*.

He is a member of the Research Staff of the Radiophysics Laboratory, Commonwealth Scientific and Industrial Research Organization and has been engaged in the development of electronic instrumentation for rain physics research and on a magnetic drum storage unit for the C.S.I.R.O.Mk. I Computer.

A graduate of the University of Sydney, he did previous work on airborne measuring equipment.

Nat'l Simulation Conference Calls for Papers by June 25

The 1958 National Simulation Conference, sponsored by the IRE Professional Group on Electronic Computers and the IRE Dallas Section, will be held at the Statler-Hilton Hotel, Dallas, Texas, October 23–25, 1958.

Technical papers in the general field of simulation are solicited. One hundred-word abstracts and 500-word summaries to aid in paper selection should be transmitted in duplicate to the Technical Program Chairman: D. J. Simmons, Route 8, Box 447, Fort Worth, Texas. The deadline is June 25, 1958.

In addition to topics such as the analog and/or digital simulation of mathematical, physical, logistic, economic, biological, and chemical systems, papers are desired covering advances in analog computer system and component design, techniques, and applications, as well as new methods of determining and improving the accuracy of analog solutions.

No "proceedings" volume will be published as such, but authors of papers accepted for the conference will be expected to submit the full text of their papers to the IRE Transactions on Electronic Computers for possible publication in that journal.

URSI Spring Meeting Slated

The U.S.A. National Committee of the International Scientific Radio Union and the IRE will co-sponsor a Spring Meeting at the Willard Hotel, Washington, D. C., April 24–26, 1958. A technical session of interest to all participants will be held on Thursday morning, April 24, followed by sessions in the following fields: Commission 1—Radio Measurement Methods and Standards; Commission 2—Tropospheric Radio Propagation; Commission 3—Ionospheric Radio Propagation; Commission 4—Radio Noise of Terrestrial Origin; Commission 5—Radio Astronomy; Commission 6—Radio Waves and Circuits; Commission 7—Radio Electronics.

The four IRE Professional Groups participating are the PG's on Antennas & Propagation, Circuit Theory, Instrumentation, and Microwave Theory & Techniques.

MARS Radio Releases Program

The Air Force MARS Eastern Technical Net, which broadcasts over the air every Sunday afternoon at 2 P.M. (EST) on 3295, 7540 and 7635 kc, announces the following program: March 2—"The Nike Missile," Charles Willhite; March 9—"Telephone Toll Switching Systems," Oscar Myers; March 16 and 23—"Airways Electronic Symposium," by representatives of major airlines with W. T. Carnes and Frank White; March 30—"Electronic Test Equipment for the Blind," Robert Gunderson; April 6—"Radio Teletype," Sal Barone; April 13—"The Modern Commercial Transmitting Station," W. J. McCambridge; April 20—"The Modern Commercial Message Center," W. S. Sparks.

Calendar of Coming Events and Authors' Deadlines*

(Continued)

PGEWS Symp., New York City, Oct. 2–3

Nat'l Electronics Conf., Hotel Sherman, Chicago, Ill., Oct. 13–15

IRE Canadian Convention, Toronto, Can., Oct. 8–10

PGCS Symp. on Aero. Communications, Hotel Utica, Utica, N. Y., Oct. 20–22

Nat'l Simulation Conf., Dallas, Tex., Oct. 23–25 (DL*: June 25, D. J. Simmons, Rte. 8, Box 447, Ft. Worth, Tex.)

EIA-IRE Radio Fall Meeting, Sheraton Hotel, Rochester, N. Y., Oct. 27–29

East Coast Aero. & Nav. Elec. Conf., Lord Baltimore Hotel and 7th Regiment Armory, Baltimore, Md., Oct. 27–29

PGED Meeting, Shoreham Hotel, Washington, D. C., Oct. 30–Nov. 1

Atlanta Section Conf., Atlanta-Biltmore Hotel, Atlanta, Ga., Nov. 17–19

NEREM, Mechanics' Bldg., Boston, Mass., Nov. 19–20

Acoustical Soc. of Amer. Mtg., Chicago, Ill., Nov. 20–22

Elect. Computer Exhibition, Olympia, London, Eng., Nov. 29–Dec. 4

Eastern Joint Computer Conf., Bellevue-Stratford Hotel, Philadelphia, Pa., Dec. 3–5

Nat'l Symp. on Global Comm., St. Petersburg, Fla., Dec. 3–5 (DL* Aug. 1, M. R. Donaldson, Elec. Commun. Inc., 1501-72 St. N., St. Petersburg, Fla.)

PGVC Annual Mtg., Chicago, Ill., Dec. 4–5

Mid-Amer. Elec. Convention, Mun. Audit., Kansas City, Mo., Dec. 9–11 (DL* Aug. 1, Wilbert O'Neal, The Vendo Co., 7400 E. 12th St., Kansas City, Mo.)

1959

IRE Region 7 Conf., Albuquerque, N. M., Feb. 2–4

Transistor-Solid State Circuits Conf., Univ. of Pa., Philadelphia, Pa., Feb. 12–13

IRE Nat'l Convention, New York City, Mar. 23–26

SW Regional Conf., Dallas, Tex., Apr. 16–18

Nat'l Aero & Nav. Elec. Conf., Dayton, Ohio, May 11–13

WESCON, San Francisco, Calif., Aug. 18–21

Nat'l Electronics Conf., Chicago, Ill., Oct. 12–14

East Coast Aero. & Nav. Conf., Baltimore, Md., Oct. 26–28

Nat'l Automatic Control Conf., Hotel Sheraton, Dallas, Tex., Oct. 28–30

PGED Meeting, Shoreham Hotel, Washington, D. C., Oct. 29–31

Radio Fall Meeting, Syracuse, N. Y., Nov. 9–11

Eastern Joint Computer Conf., Hotel Statler, Boston, Mass., Nov. 30–Dec. 3

* DL = Deadline for submitting abstracts.

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Books

Transistor Circuits and Applications by J. M. Carroll

Published (1957) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 16, N. Y. 280 pages+3 index pages+ix pages. Illus. 11¼×8¾. $7.50.

This book consists of 106 technical articles on transistor circuits and applications which have appeared in *Electronics* magazine during the years 1950 to 1956. The material describes various applications of transistors in broadcasting, home-entertainment, military, communications, computing, control, and industrial equipment. The majority of the reproduced articles contain a bibliography which is helpful to further investigation of the subject. The body of the text is well indexed.

There are eight chapters as follows: 1. Principles of Circuit Design; 2. Design of Transistor Amplifiers; 3. Transistor Oscillators; 4. Design of Transistor Pulse Circuits; 5. Broadcast and Home Entertainment Applications; 6. Military and Communications Equipment; 7. Computers and Servomechanisms; 8. Industrial, Scientific, and Medical Devices.

This book, like most others on transistors, suffers to a minor and unavoidable extent from the rapid progress of the art. A few of the transistors and circuit arrangements shown are already archaic. Aside from this slight defect, the manual will be useful to engineers, technicians, and students as a reference work and as a collection of specific designs and techniques.

R. P. Burr
Circuit Research Co.
Cold Spring Harbor, N. Y.

Introduction to Transistor Circuits by E. H. Cooke-Yarborough

Published (1957) by Interscience Publishers, 250 Fifth Ave., N. Y. 1, N. Y. 147 pages+5 index pages+xii pages. Illus. 5¾×9. $2.55.

In this little volume E. H. Cooke-Yarborough, Head of the Electronics Division of the Atomic Energy Research Establishment at Harwell, presents a concise and personal statement of the important basic concepts underlying the design of selected types of transistor circuits. The book opens with a brief and generally lucid discussion of transistor action, distinguished by the introduction of only those essential mathematical relationships that bear directly on circuit problems. (The price paid for this approach is occasional obscurity or inexactness, as shown near the bottom of page 13.) Then the classic T-equivalent circuit and low-frequency amplifier configurations are described, and a simple but effective attack on the problem of cascaded amplifiers is begun. The treatments of d-c operating conditions and of high-frequency performance are disappointingly superficial, and in spots obscure.

The best parts of the book are the all-too-brief chapters on pulse circuits and applications. Bistables and other basic switching and timing circuits are introduced with clarity, and a particularly good brief treatment of carrier storage is given. Qualitative but illuminating discussions of "high-tension" converters, counting-rate circuits, scaling circuits, and signal switches are presented. The chapter on transistors in computers, which closes the book, is inconsequential.

Emphasis throughout the book is on a physical understanding of the topics discussed, with simple formulas and physical reasoning replacing exhaustive mathematical analysis. Such methods occasionally make demands on the physical intuition and circuit design experience of the reader—a good result.

Brief and selective bibliographies are appended to each chapter.

Perspicuous distinct symbols for junction and for point-contact transitors are employed in accordance with the practice at several British laboratories.

On the whole, this modest book at a modest price is modestly recommended for the reader with some electronic circuit-design background who desires a concise introduction to transistor circuits, for the possibly tarnished expert who can benefit from a brief re-exposure to selected fundamentals of his art, and for the teacher seeking new points of view with which to enlighten (or confound) his students. But do not expect to find the whole story here; that was not the author's objective.

H. E. Tompkins
Moore School of Electrical Engineering
Philadelphia, Pa.

Receiving Aerial Systems by I. A. Davidson

Published (1957) by Philosophical Library, 15 E. 40 St., N. Y. 16, N. Y. 144 pages+5 pages of appendix+2 index pages+vii pages. 69 figures. 7¼×5. $4.75.

This book gives a general treatment of receiving antennas for TV and FM with particular emphasis on vertical polarization as used in Great Britain. The treatment is for the most part qualitative in nature and should be of great value to technicians dealing with design, fabrication and installation problems.

The book is divided into thirteen chapters which covers in general terms; radiation and impedance properties, various types in common use and how to make a choice, mechanical features and installation practice, and finally some discussion on methods of making characteristic measurements. The book concludes with a short prediction concerning future developments.

The language is typical English, using such terms as aerial for antenna, earth for ground, valve for vacuum tube and crank in connection with proper location of the antenna. Because of the language used and the emphasis on FM and TV bands as used in Great Britain it will be most useful to technicians in this environment.

Electromagnetic waves are stated to have a direction of polarization of either vertical or horizontal on page 6. Actually the wave in general will be elliptically polarized and may have a diagonal major axis.

Very few typographical errors were noted. The radiation resistance formula on page 31 does not have an = sign. The type size is quite small resulting in much more wordage than would normally be expected in a book of this size.

In general the descriptions are quite clear. This is perhaps due to the considerable amount of discussion on each subject.

Finally, in summary, it appears that this book will be appreciated by the technician dealing with TV and FM receiving antenna systems. Since the book treats the subject qualitatively it will be of little use to engineers.

C. E. Smith
Cleveland Inst. of Radio Electronics
Cleveland, Ohio

An Introduction to Probability Theory and Its Applications, Vol. I, 2nd ed., by William Feller

Published (1957) by John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 436 pages+13 appendix pages+11 index pages+xv pages. 9¼×6¼. $10.75.

The second edition of Professor Feller's book will be welcomed by all radio engineers interested in statistical communication theory. Quoting from the flyleaf of the book, "Like its predecessor, the second edition of *An Introduction to Probability Theory and Its Applications* serves a dual purpose. It develops probability theory rigorously as a mathematical discipline, and, at the same time, illustrates the broad variety of practical problems with modern techniques used in their solution. For his illustrative material and examples, the author draws from a great many fields, including engineering, genetics, physics, and statistics. He includes new results concerning fluctuation theory developed by elementary methods.

"In addition to new material in many of the chapters, the text includes two new chapters. One of these covers by elementary methods surprising phenomena of random walks and general fluctuation theory. An extended treatment of compound distributions and branching processes is offered in the other new chapter.

"In order to present the intuitive background, and the basic concepts of probability theory unhampered by analytical formalism, the first volume is restricted to discrete sample spaces. The restricted coverage allows the author to treat many typical problems in great detail and to explain the probabilistic approach to them."

The first edition of Professor Feller's book has been extremely useful to this reviewer, particularly his careful and illuminating discussion of Markov chains. The second edition appears to be a further improvement of an already excellent text. Perhaps the greatest merit of the volume is that it succeeds in providing a mathematically accurate presentation of probability theory while at the same time building up the intuitive "physical" understanding of the subject matter. The author is a mathe-

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matician and the subject is presented from a mathematician's point of view. This point of view and its advantages are carefully emphasized and explained throughout the book so that the reader will not only learn mathematical techniques but will also get an insight into a mathematician's approach to a mathematical theory. Because of its educational as well as its technical value, Dr. Feller's book is an excellent text for an introductory course on probability theory; furthermore, it deserves to be in the personal library of every student of statistical communication theory.

The only regret that this reviewer has in seeing the appearance of this second edition is that he would have preferred, if a choice was possible, to see the appearance of the first edition of the second volume. While this reviewer strongly agrees with Professor Feller on the wisdom of developing the basic concepts of probability in the context of discrete sample spaces, he is anxiously awaiting his promised elucidation of the analytical difficulties arising in connection with continuous random processes.

R. M. Fano
Massachusetts Institute of Technology
Cambridge 39, Mass

Electronic Components Handbook ed. by Keith Henney and Craig Walsh

Published (1957) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 224 pages+6 index pages+27 appendix pages. Illus. 11¼×8¾. $9.00.

Reviewing a "handbook" is not an easy assignment, but it happens that one of the authors of this book has proposed an excellent guide for reviewers* that states, in part, that a review should show "what the book is supposed to do,—how well it does it,—the purpose of the author,—the audience aimed at,—and what the book has to offer the reader." A handbook, according to the usual concepts, should contain much general data about the subjects dealt with, here four items only: resistors, capacitors, relays and switches. This it does. A specific objective of the authors was to give information to electronic equipment designers that would increase the operating reliability of any equipments using these items. All material was gathered and prepared under a grant from the Air Force (WADC).

"The audience aimed at"—here the answer is less evident. A general thumbing through the pages shows that probably more than a quarter of the space covers elementary details of these basic components, the same material found in high school physics books and in hundreds of other books as well. Do present day designers need to start at this level? Why, after more than a score of years of intensive development of the most complex equipments in the world, must today's designers have to look up what a resistor or a capacitor is and does?

Much of the book deals with the good and bad points of the various types of components, and where they should be used and where not, and the effects of various environments. It would be nice if such information could be sorted in this way. We grant that some designers lack a background of experience from developments as far back as World War II, and furthermore have had little apprenticeship under competent designers who know how to impart useful information. And there is little opportunity for them to evaluate good designs, with the scarcity of the feedback of failure data based on similar equipments in the field, and so some sort of information assistance must be provided somehow. After an evaluation of the text to find out if this book is a real "designers' bible" this reviewer feels that much is missing.

An attempt to list some of the surprising ideas noted was abandoned, since taking statements out of context is as misleading as is the arriving at conclusions as to the reliability of isolated components considered apart from their equipments. An underlying feeling of "caveat emptor" throughout seems to imply that everyone is out to "get" the designer. The engineering departments of component manufacturers are not all dumb but they do go all out to supply items that meet their customers' requests, no matter how foolish their specifications might be in relation to the real needs of the project. To prepare purchase specifications from the data appearing in this handbook would certainly result in over-engineering and promote specialties that are outside of the producer's normal procedures and test programs in many cases. A special item, even an expensive one, is not necessarily a better one and might even be inferior in applications of some types to the regular items made in quantities.

Designers with enough experience and common sense to be able to judge which of the statements apply and which do not from among the ideas listed should not need most of the material supplied in the book. However, it might well serve its purpose if it could be censored by experienced designers in a company (and it may be found that the deleted ideas may differ in each organization).

The authors have undertaken an unusually difficult (if not impossible) problem: that of defining component reliability in a realistic manner per se, without bringing usage factors into the picture. A certain component may last forever in nine locations in a piece of equipment and fail consistently in a tenth position. The book does emphasize the fact that even the simplest component, often taken for granted, does require careful consideration in each application. Many pages of excellent advice (such as p. 15) and excellent résumés of the roles of commercial and military standards in the selection of parts appear. The appendix summarizes the test procedures for resistors and capacitors that are covered by military specifications and makes excellent reading. The injunctions about seeking counsel from suppliers in questionable situations could well have been capitalized.

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* Proc. IRE., Feb., 1957; p. 251

Selection and Application of Metallic Rectifiers by S. P. Jackson

Published (1957) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 291 pages+29 appendix pages+2 bibliography pages+2 index pages +xiv pages. Illus. 9×6. $8.00.

The book, in the opinion of the reviewer, is a worthy contribution to the metallic rectifier field.

In Section I a good treatment of rectifier circuits, complex loads and filter is given. The simple mathematics should be of great aid to the design engineer faced with filter problems. Filter systems such as capacitor, inductance, L-type, multiple-L and π-type are well treated.

General background, construction, characteristics, and limitations of the many types of metallic rectifiers are discussed in Section II. Rectifiers such as copper oxide, magnesium copper sulphide and titanium dioxide, as well as the more common types such as selenium, germanium and silicon are covered. Chapter IV in this section gives a comparison of the significant properties of the rectifier types. This is a ready reference for the design engineer to select the type of rectifier best suited for his particular application. Subsequent chapters in this section cover the various rectifiers in more detail. A selection and application of metallic rectifiers chart is included in this section. Although it is considered to be of value to the design engineer, a graphical comparison would be of greater value and easier to use.

The design data in Section III should be of considerable value to equipment design engineers. Various equipment and applications are discussed such as battery chargers, electroplating, welders and magnetic amplifiers. The treatment on magnetic amplifiers is general and of a basic nature.

As indicated above, this book is directed primarily to the DC equipment design engineer. However, it should prove of great value to sales and sales application engineers.

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Impulse und Schaltvorgange in der Nachrichtentechnik by Heinrich Kaden

Published (1957) by R. Oldenbourg Verlag, Munich, Germany. 298 pages+6 index pages. 192 figures. 9¼×6¼. 32 DM.

This textbook on impulses and transients in communication engineering is an enlarged publication of a series of lectures given by Dr. Kaden at the Munich Technical Institute. The material of the book is very well organized, and it is presented in an excellent, didactic manner. The book should be of real help to educators anywhere presenting this subject.

The book begins with an excellent introduction to Fourier methods, followed by a chapter on statistical methods which presents the autocorrelation and the power spectrum with a smooth transition after lucid treatment of the Fourier integrals.

The next chapter treats the application of Fourier integrals in transient analysis. This chapter includes a treatment of the Heaviside Unit Step function method and a general treatment of the transmission of the television signals through cables. This is followed by a six-page exposure to the Laplace transform which appears adequately handled in this context.

The following three chapters cover the classification, the distortions, and some of the peculiarities of various transmission systems. The classical analysis of AM and

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FM sidebands is presented in a compact and lucid manner. The transmission of television signals is covered from their fundamental requirements through the considerations of phase distortion in light of the methods discussed in the earlier chapters to the transmission of the signals with single sideband.

The last chapter on pulses covers the sampling theory, the use of pulses for the testing of channels, and finally it describes pulse code modulation in three pages.

While the book covers its material very adequately and in an excellent order, it cannot be recommended as a reference book for the communication specialist. If the author would have had such purpose in mind, he would have included a much larger number of references and at least a short table of Laplace transforms. This reviewer missed particularly a reference to the elegant presentation of the relationship between the transient and the amplitude and phase responses of networks by Bedford and Fredendall ("Analysis, Synthesis, and Evaluation of the Transient Response of Television Apparatus," PROC. IRE, October, 1942) and Gabor's basic work on "The Theory of Communication" (*Journal of IEE*, September, 1946).

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Engineering Electronics by J. D. Ryder

Published (1957) by McGraw-Hill Book Co., 330 W. 42 St., N. Y. 36, N. Y. 655 pages+10 index pages +x pages. Illus. 9¼×6¼. $9.50.

This is one of the most complete texts on the subject of electronics. It is the author's intention to omit the radio phase of electronics, and confine the book to the industrial end more commonly associated with low frequency signals. The tremendous amount of material contained between the covers might indicate at first glance that the text could not be sufficiently complete to enable the reader proper comprehension. This is not so. Dean Ryder has, through excellent organization and the omission of unnecessary repetition, presented his topics in a very concise manner. There is adequate material available for a basic knowledge of any of the subjects presented.

In his analysis of circuits and applications he is very careful to start from a fundamental level. He exhibits a particular facility for logical classification of the material relating to the subject under discussion through his choice of sequence of presentation. To help the reader maintain proper perspective with a minimum of effort, he frequently offers introductory paragraphs outlining the broad aspects of the subject.

The book is well suited as a text for the electrical engineering student in his fourth year or the technical institute student with sufficient mathematics background. It is especially desirable to the engineer in the field to help him become familiar with phases of electronics in which he is not directly engaged.

The early chapters in the book are primarily concerned with fundamentals of the vacuum tube and its use as a circuit element. To the practicing engineer or electrical engineering student this would merely serve as a review and could be omitted, but to the newcomer to electronics this is necessary for an understanding of the ensuing material.

The small signal or linear region concerning the three most frequently used types of amplifiers are discussed in detail. The RC-coupled grounded-cathode circuit is analyzed for frequency response to sinusoidal waveforms and transient response to nonsinusoidals or pulses.

The nonlinear region is investigated when the power amplifier is introduced. Analysis of Class A and its limitations, investigation into Classes AB and B, and the design of the Class B amplifier are the highlights in this area.

A chapter on feedback in amplifiers follows. The first few pages are similar to the conventional approach in this type of text, but the gain in momentum is noticed when stability is discussed and the Laplace-transform makes its first appearance.

The direct-coupled amplifier includes a discussion of a number of important circuits, particularly the multistage bridge amplifier. Drift stability is given proper consideration. The dc amplifier is then employed in the analog computer to perform mathematical operations. A brief discussion of the analog computer follows in which Lagrange's equation is shown to be of help in writing equations preparatory for solution on the computer.

Fundamentals of the tube as a switching device and properties of RC and RL circuits are studied previous to their application in the digital computer. Power supplies, oscillators and Class C amplifiers together with applications in industrial electronics are the subjects for two chapters.

The principles of solid state theory are considered before placing the transistor in the circuit. A comparison is then made with the vacuum tube, and the parameters of the transistor are discussed.

A chapter is devoted to photo-electric devices and applications followed by chapters on power rectification, power control and inversion. Relays, timers and resistance welding control, and electronic motor control are the titles of the next two chapters. Fundamental principles of the servomechanism system, transfer function analysis and stability criteria are adequately covered to introduce the reader to the subject in the final chapter.

The abundance of problems at the end of each chapter are especially helpful for a better comprehension of the subjects discussed.

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Fundamentals of Electron Devices by K. R. Spangenberg

Published (1957) by McGraw-Hill Book Co., 330 W. 42 St., N. Y. 36, N. Y. 436 pages+29 pages of appendix+17 pages of problems+12 pages of bibliography+9 index pages+xii pages. Illus. 9¼×6¼. $10.00.

This book deserves the same general acceptance as Professor Spangenberg's earlier book *Vacuum Tubes*. The text material has been carefully assembled, and it is evident that the author has spent many hours in its preparation. Most of the illustrations are refreshingly new. The contents of this book are divided up approximately as follows according to general subjects: 162 pages devoted to basic physical theory with electron tubes and semiconductor theory developed in parallel; 139 pages devoted to devices including electron tubes, transistors, and photoelectric devices; 121 pages devoted to equivalent circuits, small and large-signal operation; and 14 pages devoted to noise. All bibliographical material is combined at the end of the book rather than as footnotes on individual pages where it would be most useful.

This book is intended primarily as a textbook for the third or fourth year college course in engineering. It will however also be valuable to those who are working closely with electron devices who wish to bring themselves up to date on transistors and related solid-state devices. The solid-state-physics development in this book represents about the minimum that present-day device engineers should be familiar with. Some of the more analytical developments are contained in thirteen appendices.

The development of the subject material is intended to unify electron tubes and solid-state devices. Thus, whereas in former books on electron tubes the preliminary development is predominantly on electron and ion motion in vacuum, in the present book this development is followed by material on electron and hole motion in semiconductors. Electron emission, an important aspect of electron tubes, is then developed on the basis of semiconduction theory. There is no doubt that unification of specialized subjects will become more important as greater emphasis is placed upon basic subjects in engineering curriculums. It is the reviewer's opinion—in the absence of teaching experience—that the treatment of several subjects in this book is still too specialized. Thus, as an example, diffusion current is not introduced until Chapter 7, Junction Effects, whereas for a more generalized approach it would be best introduced in Chapter 2, Electrons and Ions, where the more conventional formulation of current is presented.

Instructors contemplating the use of this book in a college course will probably want to introduce some of their own material on their pet topics, but generally the coverage is such that as a text it will fit in well with the usual college course.

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Mathematics and Computers by G. R. Stibitz and J. A. Larrivee

Published (1957) by McGraw-Hill Book Co., 330 W. 42 St., N. Y. 36, N. Y. 208 pages+12 pages of appendix+5 index pages+vi pages. Illus. 9¼×6¼. $5.00.

Recently, an article appeared in the *New York Times Book Review* section that had to do with popular books on science. Mr. John Pfeiffer made the point that until recently such books were rather scarce but such developments as radar and the atomic

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bomb have aroused great public interest and given authors the necessary encouragement. The recent Russian successes with satellites will result in more books, many on unrelated fields of science. Mr. Pfeiffer then goes on to describe how most of these books deal with "hardware" or applied research but very, very few with basic research or the scientific method. The authors of *Mathematics and Computers* appear to have reached the same conclusions at the time they started their book. As they point out in the preface:

"It is not possible for a layman without an intensive training in a particular scientific field to acquire any competence in the technical complexities of that field, but it seems not impossible and far more important that he should acquire an understanding and a sympathy with the spirit of scientific effort.

"It is with the hope of contributing a little of such an understanding that the authors have written this book. The field we have tried to map for the reader is broadly that of applied mathematics, with particular attention to those spectacular developments of the past few years, the automatic digital computers."

Here then is a book which covers computers and applied mathematics in a way which allows the layman or non-specialist to discover the how and why as well as the what. In the reviewer's opinion, the authors have succeeded admirably.

The book proper treats first of mathematics, problems and computers quite generally. This is followed by a number of chapters devoted to computers—primarily digital—a chapter on the Monte Carlo method, one on errors, and a rather short one on computers at work.

The first chapter, "Mathematics, Computers, and Problems" and the second, "Applied Mathematics and Solutions" are extremely elementary and obviously provide some background for the layman. Chapter 2, however, sets forth the change in attitude towards what constitutes a solution and contrasts the analytical method with that of numerical analysis; the latter having been put to such good use since the advent of the digital computer. The reader is now ready for a discussion of "Kinds of Problems and Where They Come From," the subject of Chapter 3. The discussions in this chapter are limited to descriptions of how various problems arise. The treatment is necessarily superficial, each type being described in a few sentences.

Chapter 4, "History of Computers" traces the histories of both analog and digital computers. The digital devices are taken up first and start (as has become traditional) with the abacus. The simple machines of Leibnitz and Pascal are treated next, followed by Babbage's work, Hollerith's, and finally the automatic digital devices. Analog computers are not covered in as much detail, but function computers, integrating computers and differential analyzers are all described, as well as counting computers.

Chapter 5, "Numerical Analysis," is an excellent introduction to the subject, although introduction may infer the existence of more material than appears in the chapter. It is written in such a way, however, as to make the average reader at whom the book is directed want to find out more about the subject which is all an author can really ask for. "Classical" numerical analysis, exact versus approximate solutions, successive approximations, including Bernoulli's and the Newton-Raphson methods, integration and interpolation are all described, as well as solving differential equations. The authors do very well in their choice of examples, both here and in the remainder of the book.

The next chapter, "Digital Computer Components," returns to equipment and opens with why the components do what they do. This is followed rapidly by sections on components for recording, storing and transferring digits and a simple explanation of addition and accumulation. For their example of binary addition the authors pick the unusual but familiar two-light switch problem—either switch being capable of turning a lamp on or off regardless of the position of the other switch. Input-output equipment and analog-digital converters are then described, thus concluding the chapter.

Chapter 7, "Logical Design of Digital Computers," describes how the simple components of the last chapter are tied together to give them the ability to solve problems. Such topics as analogies, table hunting, commands or instructions, automatic programs, memories, floating point, checking, serial versus parallel computers, and branching programs or sequences are covered in twenty pages, obviously not allowing for adequate engineering explanations; but again, keeping the authors' point of view in mind, is adequate. However, the reviewer did find this chapter somewhat weak in that the authors attempted to cover too much and did not adequately differentiate between what a logical designer must know and what he does. An example of how logical design proceeds would have been exceedingly illuminating.

In the next chapter we return to a discussion of analog computers and simulators. Function generating components are treated in enough detail to be interesting to the engineering reader, as are components for integrating and differentiating. This is followed by treatments of power boosters, logical design of analog computers and concludes with a further description of differential analyzers. In discussing the above, the authors are careful to point out the greater versatility, precision and cost of the digital devices.

Chapter 9, "Computing with Random Numbers," is an excellent account of the Monte Carlo method. It is well enough written so that it was easily understandable by a high school senior and a college freshman who immediately wanted more information on the subject.

The book concludes with chapters on "Computer Errors" and "Computers at Work." The former describes how errors arise, are propagated, and must be accounted for. The latter mentions some well-known computers, as well as applications including language translation, the playing of games and the control of machines. This is followed by a good bibliography covering equipment and applications.

It is the reviewer's opinion that the authors have succeeded in their intent as set forth in the preface. In addition, their book should be of interest to engineers and scientists far removed from computers who wish to gain some speaking acquaintance with that field.

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Recent Books

Buckles, R. A., *Ideas, Inventions and Patents.* John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $5.95.

Canning, R. G., *Installing Electronic Data Processing Systems.* John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $6.00.

Chapin, Ned, *An Introduction to Automatic Computers*, D. Van Nostrand Co., 257 Fourth Ave., N. Y. 10, N. Y. $8.75.

Chute, G. M., *Electronics in Industry*, 2nd ed. McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $7.50.

Colebrook, F. M., and Head, J. W., *Mathematics for Radio and Electronics.* Philosophical Library, 15 E. 40 St., N. Y. 16, N. Y. $6.00.

Forbes, G. F., *Digital Differential Analyzers*, 4th ed. G. F. Forbes, 10117 Bartee Ave., Pacoima, Calif. $5.00.

Goldberg, M. S., *Economics of Atomic Energy.* Philosophical Library Inc., 15 E. 40 St., N. Y. 16, N. Y. $6.00.

Goudet, G. and Meuleau, C., *Semiconductors—Their Theory and Practice*, translated from the French by G. King. Previously reviewed in its original edition in the Proc. IRE, July, 1957, p. 1031. Essential Books, Inc., 16-00 Pollitt Dr., Fair Lawn, N. J. $18.90.

Hunter, J. L., *Acoustics.* Prentice-Hall, Inc., 70 Fifth Ave., N. Y. 11, N. Y. $8.50.

Lanczos, Cornelius, *Applied Analysis.* Prentice-Hall, Inc., 70 Fifth Ave., N. Y. 11, N. Y. $9.00.

Llewellyn-Jones, F., *Ionization and Breakdown in Gases.* John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $3.50.

Nuclear Engineering, ed. by C. F. Bonilla. McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $12.50.

Rodenhuis, E., *Dry-Battery Receivers with Miniature Valves.* Philips' Technical Library book, obtainable from Elsevier Press, 2330 W. Holcombe Blvd., Houston 25, Tex. $4.95.

Smythe, D. W., *The Structure and Policy of Electronic Communications.* University of Illinois Bulletin Series No. 82. Bureau of Economic and Business Research, Univ. of Illinois, Urbana, Ill. $1.50.

Solid State Physics, Vol. II, ed. by F. Seitz and D. Turnbull. Academic Press, Inc., 111 Fifth Ave., N. Y., 3 N. Y. $10.00.

Stockman, Harry, *Distributed Amplification*, 2nd ed. SER Co., 543 Lexington St., Waltham, Mass. $2.90.

Stockman, Harry, *The $j\omega$ or Symbolic Method.* SER Co., 543 Lexington St., Waltham, Mass. $3.50.

Weiler, H. D., *High Fidelity Simplified*, 3rd ed. John F. Rider, Inc., 116 W. 14 St., N. Y. 11, N. Y. $2.50.

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Aeronautical & Navigational Electronics—Joseph General, 6019 Highgate Dr., Baltimore 15, Md.
Antennas & Propagation—J. I. Bohnert, Code 5200, Naval Research Lab., Washington 25, D. C.
Audio—Dr. H. F. Olson, RCA Labs., Princeton, N. J.
Automatic Control—E. M. Grabbe, Ramo-Wooldridge Corp., Box 45067, Airport Station, Los Angeles 45, Calif.
Broadcast & Television Receivers—H. A. Bass, Avco Mfg. Corp., Arlington, Cincinnati, Ohio.
Broadcast Transmission Systems—C. H. Owen, 7 W. 66th St., N. Y. 23, N. Y.
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Engineering Writing and Speech—D. J. McNamara, Sperry Gyroscope Co., Great Neck, L. I., N. Y.
Human Factors in Electronics—H. P. Birmingham, U. S. Naval Research Lab., Washington 25, D. C. (acting)
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Medical Electronics—L. B. Lusted, M.D., Clinical Center, National Institute of Health, Bethesda 14, Md.
Microwave Theory and Techniques—W. L. Pritchard, Raytheon Mfg. Co., Newton, Mass.
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* Numerals in parentheses following section designate region number. First name designates Chairman, second name, Secretary.

(Cont'd)

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Hawaii (7)—Vaughn Kelly, 99-1215 Aiea Hgts. Dr., Aiea, Hawaii; D. L. Grubb, 236 Paiko Dr., Honolulu, Hawaii.
Houston (6)—M. A. Arthur, Humble Oil & Refining Co., P.O. Box 2180, Houston 1, Tex.; C. G. Turner, Communications Engineering Co., P.O. Box 12325, Houston 2, Tex.
Huntsville (3)—W. O. Frost, P.O. Box 694, Huntsville, Ala.; J. E. Douma, 804 Carmelian St., S.E., Huntsville, Ala.
Indianapolis (5)—N. G. Drilling, 3002 Ashland Ave., Indianapolis 26, Ind.; R. B. Flint, 4038 Evelyn St., Indianapolis 24, Ind.
Israel—E. H. Frei, Weizman Inst. of Science, Rehovoth, Israel; Moshe Tkatsh, Kyriat Motzkin POB 1, Israel.
Ithaca (1)—W. E. Gordon, Phillips Hall, Cornell Univ., Ithaca, N. Y.; C. E. Ingalls, 106 Sheldon Rd., Ithaca, N. Y.
Kansas City (6)—P. C. Constant, Jr., 3014 E. Meyer Blvd., Kansas City 30, Mo.; N. E. Vilander, 2509 W. 83rd St., Kansas City 15, Mo.
Little Rock (6)—J. D. Reid, 2210 Summit, Little Rock, Ark.; J. P. McRae, Route 1, Box 27, Scott, Ark.
London (8)—J. C. Warder, 513 Princess Ave., London, Ont., Canada; J. D. B. Moore, 27 McClary Ave., London, Ont., Canada.
Long Island (2)—E. G. Fubini, Airborne Instrument Labs., 160 Old Country Rd., Mineola, L. I., N. Y.; E. K. Stodola, 118 Stanton St., Northport, N. Y.
Los Angeles (7)—J. K. Gossland, 318 E. Calaveras St., Altadena, Calif.; R. G. Kuck, 914 Arroyo Dr., So. Pasadena, Calif.
Louisville (5)—M. C. Probst, 5067 Poplar Level Rd., Louisville, Ky.; W. J. Ryan, 207 E. Broadway, Louisville 1, Ky.
Lubbock (6)—E. W. Jenkins, Jr., Shell Oil Co., Admin. Dept., Box 1509, Midland, Tex.; J. J. Criswell, 511 50th St., Lubbock, Tex.
Miami (3)—Capt. C. M. Ryan, 1330 S. Greenway Dr., Coral Gables, Fla.; E. W. Hannum, 5920 S. W. 5 Terrace, Miami 4, Fla.
Milwaukee (5)—J. E. Jacobs, 6230 Hale Park Dr., Hales Corners, Wis.; F. J. Lofy, Milwaukee School of Engrg., 1025 N. Milwaukee St., Milwaukee 2, Wis.
Montreal (8)—R. E. Penton, 6695 Fielding Ave., Apt. 29, Montreal, Que., Can.; R. Lumsden, 1680 Lepine St., St. Laurent, Montreal 9, Quebec, Canada.
Newfoundland (8)—J. B. Austin, Jr., Hq. 1805th AACS Wing, APO 862, c/o PM, New York, N. Y.; J. A. Willis, Canadian Marconi Co., Ltd., Box 994, St. John's Newfoundland, Canada.
New Orleans (6)—M. F. Chapin, 8116 Hampson St., New Orleans, La.; G. A. Hero, 1102 Lowerline St., New Orleans 18, La.
New York (2)—J. S. Smith, 3717 Clarendon Rd., Brooklyn, N. Y.; Joseph Reed, 52 Hillcrest Ave., New Rochelle, N. Y.
North Carolina (3)—H. H. Arnold, 548 S. Westview Dr., Winston-Salem, N. C.; C. A. Norwood, 830 Gales Ave., Winston-Salem, N. C.
Northern Alberta (8)—D. L. MacDonald, 10330—84 Ave., Edmonton, Alta., Can.; Frank Hollingsworth, 9619—85th St., Edmonton, Alberta, Canada.
Northern New Jersey (2)—T. N. Anderson, 1539 Deer Path, Mountainside, N. J.; G. D. Hulst, 73 Mt. Prospect Ave., Verona, N. J.
Northwest Florida (3)—G. C. Fleming, 579 E. Gardner Dr., Fort Walton Beach, Fla.; W. F. Kirlin, 67 Laurie Dr., Fort Walton Beach, Fla.
Oklahoma City (6)—Nicholas Battenburg, 2004 N.W. 30th St., Oklahoma City 6, Okla.; E. W. Foster, 5905 N.W. 42 St., Oklahoma City 12, Okla.
Omaha-Lincoln (5)—J. S. Petrik, c/o KETV, 27 & Douglas Sts., Omaha, Neb.; H. W. Becker, 1214 N. 34 St., Omaha 3, Neb.
Ottawa (8)—L. H. Doherty, 227 Barclay Rd., Ottawa 2, Ont., Canada; R. S. Thain, 54 Rossland Ave., Box 474, City View, Ont., Canada.
Philadelphia (3)—Nels Johnson, Philco Corp., 4700 Wissahickon Ave., Philadelphia 44, Pa.; W. A. Howard, WRCV, 1619 Walnut St., Philadelphia 4, Pa.
Phoenix (7)—G. L. McClanathan, 509 E. San Juan Cove, Phoenix, Ariz.; E. S. Shepard, 5716 N. 19 St., Phoenix, Ariz.
Pittsburgh (4)—J. B. Woodford, Jr., Box 369, Carnegie Tech. P.O., Pittsburgh 13, Pa.; A. E. Anderson, 1124 Olympic Hts. Dr., Pittsburgh 35, Pa.
Portland (7)—D. C. Strain, 7325 S.W. 35 Ave., Portland 19, Ore.; W. E. Marsh, 6110 S.W. Brugger St., Portland 19, Ore.
Princeton (2)—L. L. Burns, Jr., RCA Labs., Princeton, N. J.; Sylvan Fich, College of Eng., Rutgers Univ., New Brunswick, N. J.
Regina (8)—J. A. Funk, Saskatchewan Gov't, Regina, Sask., Can.; E. C. Odling, 1121 Minto St., Regina, Saskatchewan, Canada.
Rio de Janeiro, Brazil—M. P. De Britto Pereira, Caixa Postal 562, Rio de Janeiro, Brazil; Maj. H. C. de Mattos, Rua Nascimento Silva 391, Rio de Janeiro, Brazil.
Rochester (1)—B. L. McArdle, Box 54, Brighton Station, Rochester 10, N. Y.; L. D. Smith, 27 Landing Park, Rochester, N. Y.
Rome-Utica (1)—E. R. Orear, 36 Herthum Rd., Harts Hill Heights, Whitesboro, N. Y.; R. A. Zachary, Jr., 11 Arbor Drive, New Hartford, N. Y.
Sacramento (7)—R. A. Poucher, Jr., 3021 Mountain View Ave., Sacramento 21, Calif.; X. W. Godfrey, 3220 Fulton Ave., Sacramento 21, Calif.
St. Louis (6)—C. E. Mosley, 8622 St. Charles Rock Rd., Overland 14, Mo.; R. D. Rodenroth, 7701 Delmont, Affton 23, Mo.
Salt Lake City (7)—J. S. Hooper, 1936 Hubbard Ave., Salt Lake City 5, Utah; A. L. Gunderson, 3906 Parkview Dr., Salt Lake City 17, Utah.
San Antonio (6)—F. A. Brogan, 433 Brees Blvd., San Antonio 9, Tex.; W. L. Donaldson, 129 El Cerrito, San Antonio, Tex.
San Diego (7)—A. H. Drayner, 4520 62nd St., San Diego 15, Calif.; R. E. Honer, 5462 Mary Lane Dr., San Diego, Calif.
San Francisco (7)—Meyer Leifer, Electronic Defense Lab., Box 205, Mountain View, Calif.; V. B. Corey, 385 Gravatt Dr., Berkeley 5, Calif.
Schenectady (1)—A. E. Rankin, 833 Whitney Dr., Schenectady, N. Y.; Sec.-Treas. to be appointed later.
Seattle (7)—R. H. Hoglund, 1825 E. Lynn St., Seattle 2, Wash.; L. C. Perkins, Box 307, Des Moines, Wash.
Shreveport (6)—H. H. Moreland, Hq. 2nd Air Force, Barksdale AFB, La.; L. Hurley, 2736 Rosemont, Shreveport, La.
South Bend-Mishawaka (5)—R. D. Ewing, 3755 Terry Lane, Mishawaka, Ind.; H. E. Harry, Bendix Prod. Div.-Missiles, 400 S. Beiger St., Mishawaka, Ind.
Southern Alberta (8)—W. K. Allan, 2025 29th Ave., S.W., Calgary, Alta., Canada; R. W. H. Lamb, Radio Station CFCN, 12th Ave. and Sixth St., E., Calgary, Alberta, Canada.
Syracuse (1)—J. B. Russell, Salt Spring Rd., Fayetteville, N. Y.; R. N. Lothes, General Electric Co., Bldg. 3—Industrial Park, Syracuse, N. Y.
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Toledo (4)—H. L. Neuert, 3534 Woodmont Rd., Toledo 13, Ohio; K. P. Herrick, 2516 Fulton St., Toledo 10, Ohio.
Toronto (8)—H. W. Jackson, 352 Laird Dr., Toronto 17, Ont., Canada; R. J. A. Turner, 66 Gage Ave., Scarborough, Ont., Canada.
Tucson (7)—P. E. Russell, Elec. Engrg. Dept., Univ. of Ariz., Tucson, Ariz.; C. L. Becker, 4411 E. Sixth St., Tucson, Ariz.
Tulsa (6)—R. L. Atchison, 415 E. 14 Pl., Tulsa 20, Okla.; B. H. Keller, 1412 S. Winston, Tulsa 12, Okla.
Twin Cities (5)—E. W. Harding, 5325 Colfax Ave., S. Minneapolis, Minn.; S. W. Schulz, 3132 Fourth St., S.E., Minneapolis 14, Minn.
Vancouver (8)—R. A. Marsh, 3873 W. 23 Ave., Vancouver, B. C., Canada; T. G. Lynch, 739 Edgewood Rd., North Vancouver, B. C., Canada.
Virginia (3)—F. E. Brooks, Box 277, Colonial Ave., Colonial Beach, Va.; E. S. Busby, Jr., 1608 'B' St., Portsmouth, Va.
Washington (3)—A. H. Schooley, 3940 First St., S.W., Washington 24, D. C.; J. E. Durkovic, 10316 Colesville Rd., Silver Spring, Md.
Western Massachusetts (1)—R. P. Sheehan,

(Cont'd)

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Ballou Lane, Williamstown, Mass.; A. K. Hooks, Sprague Elec. Co., North Adams, Mass.

Western Michigan—R. R. Stevens, 1915 Lotus S. E., Grand Rapids, Mich.; R. V. Hammer, 1961 Leahy St., Muskegon, Mich.

Wichita (6)—W. K. Klatt, 2625 Garland, Wichita 4, Kan.; A. T. Murphy, Univ. of Wichita, Dept. of Elec. Engrg., Wichita 14, Kan.

Williamsport (4)—(No chairman at present); W. H. Bresee, 1666 Oak Ridge Pl., Williamsport, Pa.

Winnipeg (8)—C. J. Hopper, 332 Bronx Ave., Winnipeg 5, Man., Canada; T. J. White, Dept. of E.E., University of Manitoba, Winnipeg 9, Man., Can.

Subsections

Buenaventura (7)—M. H. Fields, 430 Roderick St., Oxnard, Calif.; D. J. Heron, 1171 Brunswick Lane, Ventura, Calif.

Burlington—Officers to be elected.

Charleston (3)—F. A. Smith, Rte. 4, Melrose, Box 572, Charleston, S. C.; Joseph Martin, 214 Victoria Ave., N. Charleston, S. C.

East Bay (7)—C. W. Park, 6035 Chabolyn Terr., Oakland 18, Calif.; Robert Rector, 1425 Edith St., Berkeley, Calif.

Eastern North Carolina (3)—H. Hulick, Station WPTF, Insurance Bldg., Raleigh, N.C.; M. C. Todd, Wendell, N. C.

Gainesville (3)—W. E. Lear, Dept. of Elec. Eng., Univ. of Fla., Gainesville, Fla. (Chairman)

Kitchener-Waterloo (8)—Jules Kadish, Raytheon Canada, Ltd., 61 Laurel St., Waterloo, Ont., Canada; G. C. Field, 48 Harber Ave., Kitchener, Ont., Canada.

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)—H. W. Haas, Box 236, State College, N.M.; M. Goldin, 1921 Calle de Suneos, Las Cruces, N. M.

Lehigh Valley (3)—F. W. Smith, E.E. Dept., Lafayette College, Alumni Hall of Engrg., Easton, Pa.; L. G. McCracken, Jr. 1774 W. Union Blvd., Bethlehem, Pa.

Memphis (3)—R. N. Clark, Box 227, Memphis State Coll., Memphis, Tenn. (Chmn)

Mid-Hudson (2)—Altman Lampe, Cramer Rd., R.D. 3, Poughkeepsie, N. Y.; M. R. Marshall, 208 Smith St., Poughkeepsie, N. Y.

Monmouth (2)—Edward Massell, Box 433, Locust, N. J.; Harrison Rowe, Box 107, Red Bank, N. J.

Nashville (3)—W. W. Stifler, Jr., Aladdin Electronics, Nashville 2, Tenn.; P. E. Dicker, Dept. of Elec. Engrg., Vanderbilt Univ., Nashville 5, Tenn.

New Hampshire (1)—M. R. Richmond, 55 Raymond St., Nashua, N. H.; R. O. Goodwin, 86 Broad St., Nashua, N. H.

Northern Vermont (1)—Charles Horvath, 15 Iby St., S. Burlington, Vt.; D. M. Wheatley, 14 Patrick St., S. Burlington, Vt.

Orange Belt (7)—R. E. Beekman, 113 N. Lillie Ave., Fullerton, Calif.; R. W. Thorpe, 2431 San Simeon, Riverside, Calif.

Palo Alto (7)—Wayne Abraham, 611 Hansen Way, c/o Varian Associates, Palo Alto, Calif.; W. E. Ayer, 150 Erica Way, Menlo Park, Calif.

Panama City (3)—C. B. Koesy, 1815 Moates Ave., St. Andrew Station, Panama City, Fla.; M. H. Naeseth, 1107 Buena Vista Blvd., Panama City, Fla.

Pasadena (7)—J. L. Stewart, Assoc. Prof. of Elec. Engrg., Calif. Inst. of Tech., Pasadena, Calif.; J. E. Ranks, 1270 Leonard Ave., Pasadena 8, Calif.

Quebec (8)—R. F. Chinnick, 500 La Visitation, Ste. Foye, Quebec City, Que., Can.; P. DuBerger, 1267 Villars Ave., Sillery, Que., Canada.

Richland (7)—R. E. Connally, 515 Cottonwood Dr., Richland, Wash.; R. R. Cone, 611 Thayer, Richland, Wash.

San Fernando (7)—J. J. Guarrera, 17160 Gresham Ave., Northridge, Calif.; C.C. Olsefsky, 8100 Aldea Ave., Van Nuys, Calif.

Santa Ana—Officers to be elected.

Santa Barbara (7)—Walter Hausz, 4520 Via Vistosa, Santa Barbara, Calif.; C. P. Hedges, 316 Coleman Ave., Santa Barbara, Calif.

USAFIT (4)—Lt. Cdr. E. M. Lipsey, 46 Spinning Rd., Dayton 3, Ohio; sec.-treas. to be appointed later.

Westchester County (2)—D. S. Kellogg, 9 Bradley Farms, Chappaqua, N. Y.; M. J. Lichtenstein, 52 Sprain Valley Rd., Scarsdale, 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.

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Scanning the Transactions

Character recognition devices are showing important signs of coming into their own. If and when they do, it will cause a revolution of major proportions in the automation of office machines. In San Francisco the Standard Oil Company of California has in operation a sensing system for automatically processing gasoline credit invoices. The system "reads" the customer's account number, which is stamped on the invoice card by his charge plate, and then automatically punches the number into the same invoice card for later processing by conventional accounting machines. The reading operation is accomplished by a slotted scanning disk, a photomultiplier behind it which converts the scanned image into electrical impulses, and a small computer which interprets the impulses. The system reads and punches 180 cards per minute.

As another example, Bell Labs. recently came up with another reading device in which numbers written with a special conductive-lead pencil are brought to bear against a plate that has seven sensitized radial lines on its face. Numbers are identified by the lines that they cross, a scheme that may prove tremendously valuable in processing the two billion long distance tickets that are used annually in the Bell System.

It seems evident that electronic machines, which already speak our language fluently at the ouput end, are at last learning to translate human language into machine language at the input end. (D. H. Shepard, P. F. Bargh, and C. C. Heasley, Jr., "A reliable character sensing system for documents prepared on conventional business devices," 1957 IRE WESCON Convention Record, Part 4.)

Automation is gaining headway in broadcasting. A recent survey of more than 30 per cent of the radio and television stations in the U. S. showed that 5 per cent are using automatic programming of tape recorded shows and announcements, and that another 35 per cent are contemplating automatic systems. A typical system consists of one tape reproducer for announcements and station identifications and either an automatic record player or a second tape reproducer for program material. Sub-audible 25 cps tones are recorded at the end of announcements and at appropriate points in the program tape to cue one reproducer to take over from the other automatically. This system can be programmed for up to 18 hours of automatic operation, but most stations use it primarily for their late-at-night programs.

Automatic programming is also being tried out in television control rooms to perform automatically the rather frantic and multitudinous switching operations that take place during station breaks. A typical 30-second station break between two network shows, involving say a 10-second slide with audio from the station announcer, followed by a 10-second sound film, followed by 3 slides and the station announcer for the last 10 seconds, requires no less than 22 manual switching operations—several of them simultaneously, all of them perfectly timed, and all crammed into just 30 seconds. The operating personnel then may have nothing at all to do for the next 29½ minutes, except perhaps to restore their shattered nerves. KRON-TV in San Francisco has installed an automatic sequential program switcher which has a panel of about 100 switches. The operator presets the switches during the 29½-minute lull and then merely pushes a button at the proper moment to set the entire station-break switching sequence into automatic operation. And now an improved system is under development in which the entire operation will be controlled by punch cards. These same cards will also be used to print up the daily program log and even bill the client. It seems evident that the machine is a more reliable timer and switch operator than the human, and that the days of amusing, as well as annoying, program switching errors are now numbered. (R. A. Isberg, "A survey of automation and the applications of tape recording in broadcasting and telecasting," IRE Trans. on Broadcast Transmission Systems, December, 1957. J. L. Berryhill, "Automation applied to television master control and film room," *loc. cit.*)

The new garnet materials discovered in the last two years seem destined to become as important to the microwave art as their ferromagnetic cousins, the ferrites. Although very few actual garnet devices have been built yet, enough has been learned now about the properties of these materials to be able to make some well-informed predictions about their capabilities. Garnets are composed of iron oxide in combination with either yttrium or rare earth oxide. Like conventional ferrites, they exhibit ferromagnetic resonance at microwave frequencies, a property associated with the spin of electrons and upon which all ferrite and garnet devices depend for their operation. The garnets, however, exhibit a much narrower resonant line width than ferrites, indicative of smaller damping. One important result of this is that garnets will operate at substantially lower frequencies than ferrites. Thus we can expect that the ferrite isolator and phase shifter, whose respective low-frequency limits are 1200 and 3000 mc, will be supplemented (but not replaced) by garnet versions operating at frequencies below 1000 mc.

Recent work has shown that ferrites can be used as microwave frequency doublers and detectors, and it now seems apparent that the garnets, too, are well suited to these tasks. Both the ferrites and garnets have an important advantage over conventional crystal doublers and detectors because their operation is based on a volume rather than a surface effect. It is expected, therefore, that they will prove superior to crystals in high-power applications and at frequencies above 50 kmc.

Even more recently it has been determined that at high signal levels certain nonlinear properties appear in ferromagnetic materials which can be utilized to amplify signals and also to absorb signals that exceed a certain threshold value. Yttrium garnet seems to be an excellent candidate for use in ferromagnetic amplifiers or limiters because its smaller damping factor serves to make the devices operable at lower threshold powers. These are but two of what may well be a whole new array of devices made possible through the nonlinear properties of ferromagnetic resonance and the small damping present in yttrium garnet. (G. P. Rodrigue, "Microwave properties and applications of garnet materials," 1957 IRE WESCON Convention Record, Part 3.)

An oscillator that uses a vibrating wire as a frequency generating and controlling element offers a simple and ingenious way of performing a wide variety of physical measurements. When placed in a transverse magnetic field, a wire can be kept in sustained oscillation by applying an ac voltage whose frequency is equal to the natural resonant frequency of the vibrating wire. The ac voltage, in turn, is obtained from the back emf generated by the movement of the wire in the magnetic field, after being amplified by a feedback amplifier. Any change in the tension on the wire will change its natural frequency, and hence, the frequency of the oscillator. This idea was put to use recently to measure, of all things, the pitch of ocean liners by detecting small changes in barometric pressure as the bow or stern rose and fell. One end of the vibrating wire was merely connected to a pressure sensitive diaphragm. Thus changes of elevation were translated into frequency changes and then converted into voltage changes by an fm converter—a sort of "rock-and-roll" recorder, as it

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were. (W. Gunkel, *et al.*, "Telemetering microbarometer for determination of vertical displacements," IRE TRANS. ON INSTRUMENTATION, December, 1957.)

Magnetic file cards offer an interesting new way of storing a great deal of information in a small volume, and yet making it readily accessible. A data handling system has been developed which can store over one billion bits in a space of only 2.5 cubic feet, using 300,000 1" by 3" magnetic file cards. The cards are searched and handled by means of high speed pneumatic card transport equipment, while reading and writing is done in much the same way as with magnetic drums. The use of individual cards as a storage medium permits rapid sorting, filing and processing of items in a way not feasible with other types of data handling systems, giving random access to a tremendous store of information in a few seconds. (R. M. Hayes and J. Wiener, "Magnacard—a new concept in data handling." 1957 IRE WESCON CONVENTION RECORD, Part 4.)

TV test transmissions during vertical blanking of the receiver is under consideration by the three networks, the Electronic Industries Association, and the FCC. By inserting test signals in the brief interval between the bottom of one picture and the top of the succeeding one, at which moment the home screen is blanked out, it will be possible to check such matters as amplitude levels, video frequency response, and differential phase and gain at frequent intervals while the station is on the air and without interfering with the program. Various proposed test signals are being field tested and comments will be filed with the FCC in June. Adoption of suitable test standards should materially improve television broadcasts, particularly by enabling the operating personnel to adhere more exactly to the standards prescribing the amplitude levels of various components of the composite signal. (R. M. Morris and J. Serafin, "Progress report on vertical interval television test signals," IRE TRANS. ON BROADCAST TRANSMISSION SYSTEMS, December, 1957.)

MOBIDIC. One is sometimes led to suspect that by far the weightiest problem that confronts the designers of a computer is the selection of a suitable name. Ground rules for this game require that the name be at once distinctive, representative of the nature of the machine, and preferably composed of initials that resemble a word. MOBIDIC easily qualifies on all counts. It stands for MOBIle DIgital Computer, and although the name suggests a nautical environment, it is being developed for the U. S. Army Signal Corps for use in a 26-foot trailer. Besides being intrigued by the title, we were also curious to know what use the Army would have for a really large-scale computer that had the rather unique feature of being transportable. The list of applications is impressive, to say the least, both as to length and variety: inventory control, service record processing, payroll processing, intelligence report processing, order of battle information, enemy movement reports, meteorological computations, aircraft loading plans, cryptoanalysis, air traffic control, fighter aircraft assignment and control, and anti-aircraft assignment and control. This all-transistor computer has a potential storage capacity of over 28,000 words in its large 1,000,000-bit memory, with a high-speed access time of 8 microseconds. It is rather remarkable that such a large and complex machine can be made available to armed forces in the field. From the standpoint of ruggedness alone, much less size, it is doubtful it would have been possible without the transistor. (J. Terzian, "System organization of MOBIDIC," 1957 IRE WESCON CONVENTION RECORD, Part 4.)

Electrophotography and triboelectricity are not words engineers are apt to run into every day, but they are reminding ful of the fact that in the last decade the engineer's attention has turned more and more to the electro-, magneto- and photo-properties of materials, and that terms similar to these are rapidly becoming commonplace in engineering literature. Electrophotography refers to a technique of capturing light images on an electrically charged surface which, upon exposure to light, becomes discharged just in the exposed areas. A dye consisting of small charged particles is then applied, and the dye adheres only to the discharged (exposed) areas to form the image. Triboelectricity refers to the charge on the particles and is simply a fancy name for friction electricity. The electrophotographic principle was discovered 15 years ago and has been developed to a high state by the Haloid Company under the name of Xerography. As a result of further developments on electrophotosensitive surfaces RCA announced three years ago an image recording system called Electrofax. This technique has now been successfully applied to the recording of oscilloscope patterns with a new instrument called an electrograph. Unlike conventional recording oscillographs, it offers for the first time a completely dry process, with the added advantages of being economical and fast. (R. A. Broding, J. D. Shroeder, and J. C. Westervelt, "The electrograph," IRE TRANS. ON INSTRUMENTATION, December, 1957.)

The impact of computers on network theory—II. It was noted here last month that computers are having an extremely important effect on the field of network theory by making it possible at last for the practical designer to use complex theoretical techniques he doesn't understand or which were formerly too laborious to work with. Computers are also leading to a broader design approach in which the whole system, instead of isolated components, is considered. A good example is suggested by a paper in the latest issue of Circuit Theory TRANSACTIONS dealing with the insertion loss method of design. Over the years, the conventional method of designing a multi-section network has been to design it section by section in such a way as to obtain an image match between sections, with relatively simple calculations. In the insertion loss method, on the other hand, the network is taken as a whole, and although the results are more attractive, the extensive and precise calculations that are required have discouraged its use. However, the increasing availability of computing machinery is now accelerating the widespread adoption of insertion loss techniques. One outstanding economy is appearing in that once a specific design has been evaluated and the results published, it has been done once and for all and other designers need not waste time computing the same problem over and over again. Thus, the author of this paper has used a computer to analyze Butterworth, Tchebycheff and Bessel filters with a wide variety of resistive terminations and choice of network configurations and element values, and has published the results for the use of all. (L. Weinberg, "Tables of networks whose reflection coefficients possess alternating zeros," IRE TRANS. ON CIRCUIT THEORY, December, 1957.)

A missile control system is often nonlinear because it must use simple, inexpensive components which are kept to a minimum. An example of this is in the roll stabilization of a missile by an auto pilot system which uses a relay servo to drive jet nozzle valves on the periphery of the missile. The analysis of the system operation indicates that a steady-state oscillation results. Although in aircraft operation almost any continuous oscillation is intolerable, in an unmanned missile it is possible that a steady-state oscillation confined to reasonable limits is entirely satisfactory. The magnitude of the oscillation is related to the relay hysteresis band, the system time delays, the gain of the control loop, the control force magnitude, and the missile geometry and weight. For satisfactory operation, it is necessary to minimize unbalanced torques, to maintain the frequency of the roll oscillation at a reasonably high value that will restrict the magnitude of the roll and the coupling with the roll control mode, and to minimize the hysteresis and time lags of the non-ideal components.

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(L. Atran, "Analysis of a nonlinear control system for stabilizing a missile," IRE TRANS. ON AUTOMATIC CONTROL, November, 1957.)

Ultrasonics delay lines are receiving more and more attention these days because of the increasing numbers of applications that call for storage of digital data for time intervals well above one microsecond. MTI radars, for example, which show only targets that are in motion, use delay-type memory circuits in order that signals from stationary targets may be subtracted out of the return signal. Many other examples may be found in computer, guidance, and telephone switching applications. Ultrasonic, rather than electrical, delay lines are used in these cases because mechanical disturbances in elastic media travel more slowly than electrical signals. Thus for longer delays an ultrasonic line makes a much more compact device.

Both major types of ultrasonic delay lines, magnetostrictive and piezoelectric, are discussed in a current issue of TRANSACTIONS. An investigation of the longitudinal propagation of pulses along a magnetostrictive rod has made it possible to determine how closely adjacent pulses can be stacked together without loss of resolution. The results of this investigation will be very helpful in determining the useful storage capacity of these lines. In another study the piezoelectric-type delay line has been analyzed by means of a new equivalent circuit representation that presents the essential impedance variations in a form that can be directly measured. This will be especially helpful to the circuit designer, who would like to know the electrical characteristics of the device in terms of practical design values. That the circuit analysts, as well as ultrasonics engineers, are becoming more involved in this field is pointed up by the fact that a session on delay lines is being jointly presented by the Professional Groups on Circuit Theory and Ultrasonics Engineering at the IRE National Convention this month. (A. Rothbart and L. Rosenberg, "A theory of pulse transmission along a magnetostrictive delay line," IRE TRANS. ON ULTRASONICS ENGINEERING, December, 1957. A. H. Meitzler, "Methods of measuring electrical characteristics of ultrasonic delay lines," *loc. cit.*)

Computers to assist human operators in the control of large chemical plants and oil refineries has been under consideration for some time. Because of the great number of variables, including human experience, that influence the desired optimum operation of the process, large computers are employed to store information, to make complex calculations, to make decisions, and to start or stop a process. The kind of application and the role assigned to the computer make a considerable difference in the characteristics that the control computer should have. Although much has been written about the subject, there has been little discussion of the relative merits of analog and digital computers for control applications. Engineers familiar with computers are aware of the advantages and disadvantages of analog and digital computers as tools for scientific investigations, but they may not have considered the problems involved in applying them to process control. Actually, each type of computer has characteristics which make it preferable for particular kinds of applications. Analog computers are fast, simple, and inexpensive for relatively small control jobs, while digital computers have accuracy, versatility, and flexibility which adapt them to more complex control jobs. (T. M. Stout, "Analog or digital computers for process control," IRE TRANS. ON AUTOMATIC CONTROL, November, 1957.)

Abstracts of IRE Transactions

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Automatic Control

PGAC-3, November, 1957

PGAC Transactions Policies (p. 1)

The Issue in Brief (p. 2)

Analog or Digital Computer for Process Control?—T. M. Stout (p. 3)

Although much has been written about the push-button or computer-controlled factory, there has been little discussion of the relative merits of analog and digital computers for control applications. Engineers familiar with computers are aware of the advantages and disadvantages of analog and digital computers as tools for scientific investigations. They have probably not considered their suitability for process control applications. In this paper, we will: 1) outline some of the tasks that will probably be assigned to a process control computer; 2) state some of the questions that must be answered before any computer is selected; 3) review some of the characteristics of analog and digital computers with special reference to process control requirements; 4) describe some systems already installed or about to be installed, in which computers are used.

Following this discussion, it will be evident, as might be anticipated in advance, that each type of computer has characteristics which make it preferable for particular kinds of applications. Analog computers are fast, simple, and inexpensive for relatively small control jobs, while digital computers have accuracy, versatility, and flexibility which adapt them to more complex control jobs.

Analysis of a Nonlinear Control System for Stabilizing a Missile—Leonard Atran (p. 8)

An autopilot with attitude and rate feedback, representative system lags, and a two-way relay servo with inherent hysteresis is considered for roll control of a missile with peripheral, tangentially operating jets.

This type of control system is shown to produce a steady-state oscillation. Missile dynamics in the presence of this hunting are developed and the relationships governing angu-

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lar position and rates are found to be functions of the oscillation frequency, control force magnitude, and missile constants (geometry and weight).

The describing function technique is utilized to determine graphically the relationship among frequency, hysteresis band, and system time delays. A comparison is made between the root locus and amplitude-phase presentation. An analog computer study of system behavior is presented to illustrate the agreement between the analysis and system performance.

Root-Locus Method of Pulse Transfer Function for Sampled-Data Control Systems—Masahiro Mori (p. 13)

The pulse transfer function is a powerful tool for analysis and synthesis of sampled-data systems. The author has extended the root-locus method to the region of the pulse transfer function so that we can analyze and synthesize the sampled-data systems by the analogous method for the continuous-data systems.

The results are as follows: the rules by which the root loci of the pulse transfer function can be plotted are identical to those of the continuous-data systems. The conditions for absolute and relative stabilities for the sampled-data systems on the z plane are conformal to those for the continuous-data systems on the s plane.

Existence of a finite settling time in a sampled-data control system has been pointed out by the root-locus method.

Input-Output Analysis of Multirate Feedback Systems—G. M. Kranc (p. 21)

A general analytical technique described in this paper permits the extension of Z-transform methods to sampled-data systems containing synchronized switches which do not operate with the same sampling rate. Sampling periods of each switch are first expressed in the form $T/p_1 \cdots T/p_n$ (where $p_1 \cdots p_n$ are integers not equal to zero) and then it is shown that each switch with a period T/p can be replaced by a system of switches and advance and delay elements where each switch operates with a sampling period T. In this way, the original sampled-data system can be represented by an equivalent system containing switches operating with the same sampling rate. The general solution of such equivalent systems is outlined in this paper.

Statistical Design and Analysis of Multiply-Instrumented Control Systems—R. M. Stewart (p. 29)

It is the purpose of this paper to show how Wiener's linear least-square filter theory may be applied to some common types of multiply-instrumented control systems. By multiply-instrumented control systems are meant those in which more than one sensing instrument is used.

A Time Domain Synthesis for Optimum Extrapolators—C. W. Steeg, Jr. (p. 32)

A direct method is presented for the solution of the integral equation for the optimum predictor in terms of the solution to the integral equation for the optimum filter. This method is a means for circumventing the practical difficulties encountered in the actual design of optimum predictors based upon the techniques derived by Wiener. The synthesis procedure is applied in order to simplify the design of extrapolators for use when the prediction interval is a non-negative function of time in contrast to the more conventional situation where prediction is made for a continuously varying instant that is always a fixed number of seconds in the future. The specific example utilized is prediction for a fixed instant that is a continuously decreasing number of seconds in the future. The discussion includes an explanation of a procedure for avoiding the solution of integral equations in the synthesis of optimum extrapolators.

PGAC Transactions Reviewers (p. 42)

Broadcast Transmission Systems

PGBTS-9, December, 1957

A Transistorized Intercom System—E. P. Vincent (p. 1)

Automation Applied to Television Master Control and Film Room—J. L. Berryhill (p. 11)

Recent Developments in TV Camera Tubes—F. S. Veith (p. 21)

Image orthicons and vidicons have been improved substantially to fulfill requirements for better TV broadcasting. The advantages of "micromesh" and "super-dynodes" are explained and the performance characteristics of a new image orthicon with very high photocathode sensitivity described. Recommendations concerning studio practices to obtain best black-and-white and color pictures are presented.

Vidicon characteristics are analyzed with respect to best performance of existing commercial types. Electron optical problems in vidicon chains are considered and operational information for optimum performance recommended. Characteristics of a new developmental vidicon with increased "effective sensitivity" is described.

A Transistor Regulated Power Supply for Video Circuits—R. H. Packard and M. G. Schorr (p. 32)

A high-efficiency all-transistor regulated power supply for 280 volts and 1.5 amperes is compared with conventional supplies using vacuum tubes. Test results show good characteristics for video use with size and weight reduced by a factor of two.

Maintenance of Directional Antennas—J. G. Rountree (p. 39)

A Simplified 5-Megawatt Antenna for the UHF Broadcaster—R. E. Fisk (p. 46)

A stacked arrangement of two UHF five-bay helical antennas is used to provide an antenna system with a power capacity of 120 kw and a power gain of 50. A power gain of 100 is achieved in the maximum direction with horizontal directivity.

The contoured pattern approach and available performance data is discussed.

Reduction of Image Retention in Image Orthicon Cameras—S. L. Bendell and K. Sadashige (p. 52)

In television cameras employing image orthicon tubes, the problem of picture sticking or image retentivity often limits the effective life of these tubes. Methods are discussed which minimize this problem by slowly rotating the image on the tube in a small circular orbit by either optical or electronic means for color or monochrome cameras.

Application of Automatic Gain Control Devices to Broadcast Audio Control—A. A. McGee (p. 59)

Present day broadcast program practices have brought about more complex operational requirements with the result that less effort can be devoted by the operator to maintaining a desirable constant audio level.

Recent developments in the audio equipment field have provided new automatic gain controlled amplifiers which can relieve the operator of many level control problems—and in many cases, can result in better over-all level control than was possible by manual means.

Progress Report on Vertical Interval Television Test Signals—R. M. Morris and John Serafin (p. 65)

The Television Allocations Study Organization—Its Objectives and Progress—G. R. Town (p. 70)

The Television Allocations Study Organization was established, at the request of the Federal Communications Commission, by five of the major associations in the television industry. The objective of TASO is to make a thorough study of the engineering factors affecting the allocation of channels for television broadcasting. To carry out this task, six engineering panels have been formed. Each is working on a specific phase of the over-all problem—(1) transmitting equipment, (2) receiving equipment, (3) field tests, (4) wave propagation, (5) analysis and theory, and (6) levels of television picture quality. A total of 174 engineers from 92 organizations from all branches of the television industry are serving on TASO panels. The work of the panels is progressing rapidly and it is expected that significant results in the form of reports for the use of the FCC and the television industry will be forthcoming by the middle of next year.

A Management View of TV Transmitter Operational Practices—R. N. Harmon (p. 76)

A Survey of Automation and the Applications of Tape Recording in Broadcasting and Telecasting—R. A. Isberg (p. 81)

The responses to a mail survey of 2735 broadcast and television stations were analyzed to determine the extent to which tape recording is presently used in routine broadcasting.

The average broadcaster has three or more tape recorders. The majority of the stations use tape extensively for delaying remote, local and network programs as well as for spot announcements, auditions, telephone calls, master recording, echo effects and sound effects.

Approximately 40 per cent of the stations evidenced interest in automatic program systems, and approximately 5 per cent of the stations reported that they are presently using automation. Reactions to the future use of automation in individual stations varied from enthusiastic to very negative, but the favorable comments prevailed. Comments from users of automation were all enthusiastic and indicated that automation permitted expansion of program service with the same size staff, better working conditions, and increased profits.

Six types of automation equipment are described, as well as an automatic one hour delay program system for network time zone application.

Circuit Theory

Vol. CT-4, No. 4, December, 1957

Abstracts (p. 296)

A New Editor Takes Over—W. R. Bennett (p. 297)

The Limits of Gain Attainable in Three-Terminal RC Networks with Two Capacitors—I. Cederbaum (p. 298)

In the paper some properties of the transfer function of a three-terminal RC network with two capacitors are deduced from a basic theorem concerning pure resistive four ports. It is shown that the zeros of such a transfer function may lie anywhere in the left half p plane (being, of course, conjugate if complex) independently of the position of the poles. However, the limit of the multiplicative constant depends on the position of both poles and zeros. Writing the transfer function in the form $T(p)=a(p+p_1')(p+p_2')/(p+p_1)(p+p_2)$, with $p_1<p_2$, we show that apart from the well-known limits 1 and $p_1p_2/p_1'p_2'$ the multiplicative constant a cannot exceed $(p_1'+p_2')p_2/p_1'p_2'$. Alternatively, putting $T(p)$ in the form $T(p)=ap^2+kp+bp_1p_2/(p+p_1)(p+p_2)$ we show that k is necessarily confined to the interval $abp_1\leq k \leq p_2+p_1(a+b-ab)$. A set of graphs based on these results is derived on which the limit of the gain attainable at infinity is plotted as a function of the location of the zeros in the complex frequency plane. A simple network configuration is given for which the multiplicative constant of the transfer function may approach arbitrarily near the above defined upper limit.

Transformations of Positive Real Functions—S. Seshu and N. Balabanian (p. 306)

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A positive real function is an analytic function of a complex variable which is regular in the right half plane and maps the right half plane into the right half plane and the real axis into the real axis. In this paper, methods of transforming one or more positive real functions with a positive real function, resulting from the transformation, are considered. In addition to collecting the known transformations of positive real functions, this paper presents a generalization of the well-known Richards' transformation and strengthens some of the known results on transformations of driving point impedance functions of two-element type networks.

Tables of Networks Whose Reflection Coefficients Possess Alternating Zeros—Louis Weinberg (p. 313)

In two preceding papers tables were presented for the design of three large classes of ladder networks. The zeros of the reflection coefficients of each of these networks were all in the left half plane or all on the imaginary axis. In this paper tables are presented for networks whose reflection coefficients possess zeros that alternate in the left and right half planes. The tables are classified on the basis of the parameter r, which is the input-to-output resistance or conductance ratio. They give the element values of normalized low-pass ladders with one of the following characteristics: maximally flat magnitude (Butterworth), equal-ripple magnitude (Tchebycheff), and maximally flat time delay (Bessel polynomial). By means of frequency transformations the networks given by the tabulated element values for the Butterworth and Tchebycheff networks may be converted to give high-pass, band-pass, and band-elimination filters. Thus the tables may be used as a handbook to give new sets of networks for the realization of these optimum characteristics.

The Pentode Gyrator—G. E. Sharpe (p. 321)

A gyrator may be constructed from four pentodes, no other network element being necessary. The method is based on a theory of ideal active elements recently proposed by the author. Two physically-distinct kinds of gyrator, electric-electric and magnetic-magnetic types, may be obtained. A modification to Telegen's gyrator symbol is proposed to distinguish these types. The lossless coupling property of ideal active elements is demonstrated and the principle stated, that ideal gyrators and transformers must be physically equivalent, whether they be passive constructed or active derived.

Theory of the Band-Centering AFC System—J. C. Samuels (p. 324)

A theory of the band-centering afc system for pulsed-carrier operated receivers is developed. The interaction between the afc system and the agc system of the IF amplifier is accounted for by introducing a so-called ideal agc action.

The open-loop characteristic of the afc system was evaluated in explicit form for arbitrary pulse shape and IF amplifier response. The results are given in terms of the pulse spectrum and the poles and zeros of the IF amplifier response function.

Specific open-loop characteristics are worked out for the rectangular pulse and 1) an IF amplifier with N synchronous single-tuned stages, 2) an IF amplifier made of a flat-staggered pair.

Some Properties of Three-Terminal Devices—S. J. Mason (p. 330)

A three-terminal device can be classified according to its deviation from three-way symmetry. Such classification offers a particularly compact and symmetrical expression of the passivity criterion and also relates the asymmetry of the device to the unilateral power gain obtainable with lossless bilateral coupling.

Reviews of Current Literature (p. 333)

Design Data for Symmetrical Darlington Filters—J. K. Skwirzynski and J. Zdunek. Reviewed by A. J. Grossman. (p. 333)

Abstracts of Articles on Circuit Theory—On a New Type of Variable Equalizer—J. Oswald (in French). (p. 334)

Two-Branch Filter Structures with Three Cut-Off Frequencies—J. E. Colin (in French). (p. 334)

On the Representation of Certain Classes of Signals by a Series of Samples—J. Bouzitat. (p. 334)

Minimum Noise Figure of Mismatched Amplifiers—H. Potzl (in German). (p. 334)

Regarding the Use of Transformation Matrices in the Investigation of Network Models—H. Edelmann (in German). (p. 334)

Mutually Coupled Degenerative Resistance Amplifiers—Kurt Franz (in German). (p. 334)

Geometric Representation of the General Series Connected Lossy Fourpole—J. De Buhr (in German). (p. 334)

A Practical Method for the Formulation of the Hurwitz Polynomial in Filter Synthesis—F. Bauhuber (in German). (p. 334)

Novel Method for the Realization of Response Curves for Two-Terminal Networks by Means of Canonical Circuits and Circuits without Coupling Impedances—R. Unbehauen (in German). (p. 335)

Crystal Filters with Sharp Cut-Offs and a Large Bandwidth for Application in Branching Filters—W. Poschenrieder (in German). (p. 335)

Correspondence (p. 336)

PGCT News (p. 342)

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Information Theory

Vol. IT-3, No. 4, December, 1957

W. B. Davenport, Jr. (p. 212)

Sputnik Et Cetera—W. B. Davenport (p. 213)

A Theory of Multilevel Information Channel with Gaussian Noise—Satosi Watanabe (p. 214)

The interval between the "0" level and "1" level of a binary channel with Gaussian noise is subdivided to provide $n=2^g$ levels per symbol. The channel capacity is computed as a function of g and of the signal-noise ratio D of the original binary channel. For a sufficiently large, fixed value of D, if we increase g indefinitely, the channel capacity approaches the logarithm of D, as can be expected from continuous channel. For a fixed value of g, if we increase D indefinitely, the channel capacity approaches g. For a given value of D, there is a certain value of g, beyond which the channel capacity does not appreciably increase any longer by increasing g. The problem is first solved by a simplifying model, and then the error introduced by this simplification is estimated.

A Generalization of a Method for the Solution of the Integral Equation Arising in Optimization of Time-Varying Linear Systems with Nonstationary Inputs—Marvin Shinbrot (p. 220)

A new method is presented for the solution of the integral equation which arises in the optimization of a system in the presence of noise when the inputs are not stationary. The method depends on the correlation functions satisfying a certain condition which, fortunately, is frequently satisfied in practical situations. A simple example is presented to illustrate the method.

On the Mean Square Noise Power of an Optimum Linear Discrete Filter Operating on Polynomial Plus White Noise Input—Marvin Blum (p. 225)

In a recent article Dr. K. R. Johnson presents an asymptotic formula for the output noise power of an optimum filter designed to make a zero-lag estimate of either the input or its derivatives. It is assumed that the input function consists of a non-random polynomial plus stationary uncorrelated noise.

It is the purpose of this paper to present an exact formula for the output noise power for the same input model. The formula presented is more general in that the estimation can be for any lag α with respect to the latest data point.

Tables and graphs of the root mean square error for the zero-lag estimation of the 0th, 1st, and 2nd derivative are presented as a function of the input polynomial up to degree 5 and memory spans up to 100 sample points. A comparison is made of the relative error in root mean square using the asymptotic formula derived by Johnson.

The Distribution of the Number of Crossings of a Gaussian Stochastic Process—C. W. Helstrom (p. 232)

It is shown how filtered Gaussian noise having a power spectrum which is a rational function of the square of the frequency can be represented as one component of a multidimensional Markov process. Methods are studied for obtaining the distribution of the number of times such a noise process crosses a given amplitude level in a fixed time interval. The generating function of this distribution is the solution of a Fokker-Planck type differential equation with appropriate boundary conditions. Integral equations are found for the generating function from which all the moments of the distribution can be calculated by iteration.

An Analysis of Coherent Integration and Its Application to Signal Detection—K. S. Miller and R. I. Bernstein (p. 237)

An important characteristic of coherent integrators is that their effective bandwidth decreases as the integration time increases. If it is only known that a weak signal occurs somewhere in a given frequency range, then the number of integration channels required to cover the specified range increases as the amount of coherent integration is increased. However, each integration channel can independently cause a false alarm, although only the particular channel in which the signal appears can cause a true alarm. The question arises therefore whether it is profitable to lengthen the coherent integration period to increase the signal-to-noise ratio when doing so requires an increase in the number of integration channels. This problem is investigated analytically. Numerical results appropriate for system design are presented as a series of graphs of missed-signal probability vs number of integration channels, with initial signal-to-noise ratio and over-all false alarm probability as parameters.

Also included is a detailed analysis of statistical properties of ideal and approximate ideal coherent integrators.

The Sequential Detection of a Sine-Wave Carrier of Arbitrary Duty Ratio in Gaussian Noise—H. Blasbalg (p. 248)

In this paper the Wald theory of sequential analysis is applied to the detection of a sine-wave carrier of arbitrary duty ratio in Gaussian noise. This is a generalization of a familiar problem. The detector law for the problem is obtained. In particular, it is specialized to the important cases: 1) arbitrary duty ratio and signal-to-noise ratio less than unity and 2) duty ratio much less than unity and peak-signal-to-noise ratio much greater than unity. For the latter case, it is shown that the best detector law goes over into a Bernoulli detector. In the former case it is shown that the only important parameter in detection is the average signal-power to noise-power ratio. For the case of unity duty ratio the detector law goes over into

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High-Energy Particle Nomograph—D. B. Hoisington (p. 62)

A nomograph is presented which has proved to be useful in calculations related to the design of high-energy particle accelerators. Rest mass is given in the range from 0.0001 to 100 atomic mass units with the energy equivalent of this mass in electron volts. The final velocity and mass of the particle are given for a range of energies from 5000 to 10^{15} electron volts. The equation for particle velocity is given in the form

$$v/c=[1-1/(1+E/E_0)^2]^{1/2},$$

where E is the particle energy and E_0 is the energy equivalent of its rest mass. The frequency required for operation of a proton synchrotron is determined to illustrate the use of the nomograph.

A Portable Scintillation Alpha Survey Instrument—W. G. Spear (p. 63)

A scintillation alpha counter utilizing a zinc sulfide fluor, a multiplier phototube, and a two-stage vacuum tube amplifier has been developed. A neon-bulb oscillator high-voltage supply is used to supply 900 volts to the multiplier phototube, and headphones are used to obtain an aural indication of counting.

The average geometry or counting efficiency over the 7.43-inch^2 probe area is approximately 13 per cent with variations from a maximum counting efficiency at probe center of 18 to 22 per cent down to approximately 7 to 8 per cent at the probe edges. The background counting rate is less than one count per minute, and the battery life is approximately 150 hours of continuous operation for the $3\frac{1}{2}$ pound instrument.

A Simple Monitor for Airborne Alpha Particles—R. D. Hiebert, R. N. Mitchell, and D. B. Tod (p. 66)

A simple continuous monitor for airborne alpha particles is described. The single-channel system can detect activity concentrations of ten times maximum permissible level for plutonium and gives indication of accumulated activity at 10-minute intervals. The instrument is designed with emphasis on simplicity, reliability, flexibility of use, and ease of decontamination.

Correction (p. 69)

News and Views (p. 70)

Ultrasonics Engineering

PGUE-6, December, 1957

Methods of Measuring Electrical Characteristics of Ultrasonic Delay Lines—A. H. Meitzler (p. 1)

This paper is concerned with methods of measuring useful electrical characteristics of ultrasonic delay lines employing piezoelectric transducers. The impedance characteristics of delay lines are discussed by means of an equivalent circuit representation, the advantage being that it presents the essential impedance variations of the line in terms of quantities that are directly measurable by an admittance bridge. Measurements of insertion loss and the determination of band-pass characteristics are carried out by the use of a conventional loss-measuring circuit. In regard to this type of measurement, the recent development and use of short delay lines (delay times of 50 μsec or less) with ceramic transducers requires the consideration of certain measuring problems resulting from the low loss in the line and the high electromechanical coupling of these transducers. It is shown that the measurement of unwanted signals arising from the sonic pulse making multiple path traversals requires special care because the input as well as the output termination affects the measurement. Sample results obtained for various types of measurements are given using data from experimental delay lines having ceramic transducers bonded to fused quartz. However, all the measuring techniques discussed may be applied to lines using quartz crystal or other piezoelectric transducers.

Ultrasonic Output Power Measurements in Liquids—G. E. Henry (p. 17)

"Gross Acoustic Power Transfer" is defined as the time rate of delivery of acoustic energy by a transducer to a selected liquid load. This quantity is related to, but should not be confused with, the power density, the intensity, the energy density, and the gross acoustic output from the transducer.

To determine gross acoustic power transfer, one can (1) make a sound pressure plot, utilizing a suitable hydrophone in a free field, convert to power, and integrate; (2) determine the rate at which sound energy is degraded to heat energy; or (3) measure acoustic radiation force. Of these, the last is the preferred method if the lateral dimensions of the transducer are large compared to the wavelength. If the sound is beamed vertically upward, one can utilize a radiation pressure float which balances the upthrust of acoustic radiation force against the downward force of gravity. Such a float assumes an equilibrium position of partial immersion in the liquid. The preferred design eliminates any need for external guides or constraints: the float can be made self-centering. Care is needed in interpreting results, but the method is basically reliable and practical.

A Theory of Pulse Transmission Along a Magnetostrictive Delay Line—A. Rothbart and L. Rosenberg (p. 32)

Biographical Notes on the Authors (p. 59)

IRE Professional Group on Ultrasonics Engineering Membership Directory (as of October 1, 1957) (p. 60)

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

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The Index to the Abstracts and References published in the PROC. IRE from February, 1956 through January, 1957 is published by the PROC. IRE, May, 1957, Part II. It is also published by *Electronic and Radio Engineer*, incorporating *Wireless Engineer*, and included in the March, 1957 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

534.1.087:621.395.616 **332**
The Condenser Microphone as a Displacement Detector Calibrator—W. Koidan. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 813–816; July, 1957.) The absolute calibration of a variable capacitance-type displacement detector was performed by using the diaphragm of a condenser microphone as a reference moving surface. Measured values of the detector response are plotted from 10 cps to 40 kc. See also 3011 of 1954.

534.2-8-16 **333**
Ultrasonic Attenuation in Metals at Low Temperatures in the Normal and in the Superconducting State—G. Kurtze. (*Naturwiss.*, vol. 44, pp. 368–370; July, 1957.) Report of measurements on Cu, Pb, and Sn single crystals for longitudinal and transverse waves.

534.2-8-16:538.221 **334**
Change of Ultrasonic Absorption in Ferrites by an External Magnetic Field—G. Uhlig. (*Nachr. Tech.*, vol. 7, p. 221; May, 1957.) Preliminary note on experiments with a ferrite rod. A strong axial magnetic field produces up to 100 per cent reduction in the ultrasonic absorption of the rod at 3 mc.

534.213-8 **335**
Dispersion Effects in Ultrasonic Waveguides and their Importance in the Measurement of Attenuation—M. Redwood. (*Proc. Phys. Soc.*, vol. 70, pp. 721–737; August 1, 1957.) The possible sources of error are investigated which arise from the use of the pulse technique in determining absorption at 10–100 mc in low-loss solid materials. Estimated errors of attenuation measurements are tabulated and compared with the intrinsic absorption in fused silica and single-crystal Ge. See also 366 of 1957 (Redwood and Lamb).

534.23:621.396.677.3 **336**
Theory of Time-Averaged-Product Arrays—A. Berman and C. S. Clay. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 805–812; July, 1957.) Mathematical analysis in polynomial form of the directional characteristics of linear additive arrays. It is shown that the same directional characteristics may be obtained from multiplicative arrays having a small number of detectors as with an additive array with many elements. See also 983 of 1956 (Fernandez Huerta).

534.232:534.64 **337**
Self-Reciprocity Transducer Calibration in a Solid Medium—R. M. White. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 834–836; July, 1957.) The application of the self-reciprocity technique to the calibration of a reversible compressional-wave transducer is described.

534.25/.26-14:534.213.4 **338**
Acoustic Refraction and Scattering with Compliant Elements—W. J. Toulis. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 1021–1033; September, 1957.) Various techniques of using mechanical structures, such as compliant tubes, for effectively increasing the compressibility of water are described; the results of measurements are analyzed.

534.75 **339**
Localization of High-Frequency Tones—W. D. Feddersen, T. T. Sandel, D. C. Teas, and L. A. Jeffress. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 988–991; September, 1957.) Continuation of earlier work (649 of 1956).

534.76:534.78 **340**
Mechanism of Binaural Fusion in the Hearing of Speech—B. M. Sayers and E. C. Cherry. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 973–987; September, 1957.) The mechanism is discussed as a form of statistical operation based upon the brain's execution of running cross correlation of the two ear signals.

534.846 **341**
Acoustics of Large Orchestral Studios and Concert Halls—K. L. Rao, T. Somerville, and C. L. S. Gilford. (*Proc. IEE*, pt. B, vol. 104, p. 597; November, 1957.) Comment on 2017 of 1957 and authors' reply.

621.395.625.6:534.862.3 **342**
Improved Light Valve for the Photographic Recording of Vibration—G. Menon Skreekantath. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 1034–1035; September, 1957.) A beam of light is modulated by passing through two similar coarse gratings, one of which vibrates, and the modulations are recorded on a moving photographic film. The process may be applied to sound recording.

ANTENNAS AND TRANSMISSION LINES

621.315.212 **343**
Multiple Screening of Flexible Coaxial Cables—L. Krugel. (*Telefunken Ztg*, vol. 30, pp. 207–214; September, 1957. English summary, pp. 218–219.) The screening effect of double layers of braiding with and without intermediate magnetic shields is investigated; results are given in graphical form. See also 2334 of 1957.

621.315.212.095.3:621.372.2 **344**
An Interation Method for Computing Electromagnetic Fields—A. Redhardt. (*Arch. elekt. Übertragung*, vol. 11, pp. 227–230; June, 1957.) The conditions at a discontinuity inside a coaxial line are analyzed and the resulting field is computed.

621.372.2 **345**
Initial Value Problems and Time-Periodic Solutions for a Nonlinear Wave Equation—F. A. Ficken and B. A. Fleishman. (*Commun. pure appl. Math.*, vol. 10, pp. 331–356; August, 1957.) A discussion of an equation governing the displacement of a taut string with restraining forces and applicable to the voltage on a uniform transmission line with certain nonlinear characteristics.

621.372.2 **346**
Return Loss: Part 2—T. Roddam. (*Wireless World*, vol. 63, pp. 583–588; December, 1957.) The relation between return loss and circuit response characteristics is discussed with particular reference to distortion in a television picture. Methods of measuring return loss are described. Part 1: January, 16.

621.372.22.09 **347**
Theory of Nonuniform Lines—R. Codelupi. (*Alta Frequenza*, vol. 26, pp. 226–282; August, 1957.) Mathematical functions are derived for use in determining the input impedances and voltages along the line and at its terminals. Examples of analyses for multiple reflections are given.

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621.372.51:621.396.674 348
A 200-Watt Balun Coupler for Centre-Fed Antennas—J. M. Shulman. (*QST*, vol. 41, pp. 26–28; June, 1957.) Commercial-type wide-band balun coils are used to provide either 75Ω/75Ω or 75Ω/300Ω unbalance/balance transformation.

621.372.51.029.6:621.317.335.3 349
The Dielectric Disk used as Transformation Quadripole for the Magnification of the Node Displacement on Measuring Lines—L. Breitenhuber. (*Arch. elekt. Übertragung*, vol. 11, pp. 223–226; June, 1957.) The maximum attainable node displacement is derived as a function of the thickness and the dielectric constant of the disk. The maximum is obtained when the disk shifts the phase by 45°; in this case and for very high dielectric constants the magnification of the node displacement is directly proportional to the dielectric constant.

621.372.8 350
Characteristics of some Ferrous and Non-ferrous Waveguides at 27 Gc/s—J. Allison, F. A. Benson, and M. S. Seaman. (*Proc. IEE*, pt. B, vol. 104, pp. 599–602; November, 1957.)

621.372.8.001.2 351
An Application of Sturm-Liouville Theory to a Class of Two-Part Boundary-Value Problems—S. N. Karp. (*Proc. Camb. phil. Soc.*, vol. 53, pt. 2, pp. 368–381; April, 1957.) "A simple solution of a general problem involving a bifurcated waveguide is presented. The purpose of the work is to explain a new and simple method of solving such problems and to exhibit an organic connexion between Sturm-Liouville theory and the theory of two-part boundary-value problems."

621.372.8.002.1 352
Waveguide Design for Die-Casting—P. Humphreys. (*Electronic Radio Eng.*, vol. 34, pp. 441–447; December, 1957.) "This article explains how components which have been designed in normal rectangular waveguide may be easily modified on a theoretical basis to make them suitable for die-casting manufacturing methods. The theory is applicable to cases where the waveguide can be manufactured by splitting it along the length of the central E plane and the unit is therefore cast in two halves."

621.372.822 353
Theory of the Helical Waveguide of Rectangular Cross-Section—R. A. Waldron. (*J. Brit. IRE*, vol. 17, pp. 577–592; October, 1957.) A mathematical treatment of the properties of a helical waveguide, regarded as an equivalent circular waveguide in which points differing in azimuth by 2π, radians are not equivalent.

621.372.832.6 354
A New Form of Hybrid Junction for Microwave Frequencies—L. Young, P. D. Lomer, and J. W. Crompton. (*Proc. IEE*, pt. B, vol. 104, p. 586; November, 1957.) Comment on 2339 of 1957 and authors' reply.

621.372.832.8 355
A Broad-Band Microwave Circulator—E. A. Ohm. (*Bell Lab. Rec.*, vol. 35, pp. 293–297; August, 1957.) A device for routing microwave energy over a variety of waveguide transmission paths is described. By surrounding the ferrite insert with material of high dielectric constant the rotation/frequency characteristic is flattened and a wide-band device is obtained.

621.372.852.22 356
Wide-Band Isolator at 4 kMc/s—S. Kawazu, K. Matsumaru, and H. Ishii. (*Rep. elect. Commun. Lab., Japan*, vol. 5, pp. 15–17; March, 1957.) The isolator which incorporates a magnetized ferrite insert covered by dielectric material has forward and reverse losses of 1.5 and 20 db, respectively, over a 530-mc band.

621.396.677:621.372.092:621.318.134 357
A New Technique in Ferrite Phase Shifting for Beam Scanning of Microwave Antennas—Reggia and Spencer. (See 387.)

621.396.677:621.396.11 358
Antenna/Propagation Mismatch—R. J. Hitchcock. (*Wireless World*, vol. 63, pp. 599–602; December, 1957.) Many point-to-point hf circuits in operation are inefficient because antenna systems are not matched to the predominant modes of propagation. Recent investigations show that on medium- and long-distance circuits low angles of arrival predominate both by day and night. Appropriate improvements to antennas and sites are discussed.

621.396.677.3 359
The Effect of the Mutual Impedance due to the Neighbouring Elements on the Driving Point-Impedances of a Linear Array—R. Parthasarathy. (*J. Inst. Telecommun. Eng., India*, vol. 3, pp. 242–247; June, 1957.) The driving point impedance of a four-element array is independent of the beam direction, provided the element spacing is not closer than $\lambda/4$.

621.396.677.3 360
The New Antenna Installation of the Vatican Short-Wave Broadcasting Services at Santa Maria di Galeria—W. Berndt. (*Telefunken Ztg*, vol. 30, pp. 174–184; September, 1957. English summary, p. 217.) Facilities are provided for changing the angles of main radiation in both the horizontal and vertical planes.

621.396.677.3:534.23 361
Theory of Time-Averaged-Product Arrays—Berman and Clay. (See 336.)

621.396.677.5 362
The "Quad" Antenna—F. B. Singleton. (*Wireless World*, vol. 63, pp. 607–608; December, 1957.) The advantages of these antennas for indoor use on bands I and II are outlined.

621.396.677.833.1 363
An Experimental Wide-Band Parabolic Aerial for the 2000-Mc/s Band—N. Ganapathy and P. E. G. T. Hopkins. (*Marconi Rev.*, vol. 20, pp. 134–152; 4th Quarter, 1957.) A 10-foot-diameter horn-fed paraboloid for multichannel links, combining high gain with good impedance match over a wide band, is described; performance data are quoted.

AUTOMATIC COMPUTERS

681.142 364
Odd Binary Asynchronous Counters—J. F. Robertson. (IRE TRANS, vol. EC-5, pp. 12–15; March, 1956. Abstract, PROC. IRE, vol. 44, pt. 1, p. 832; June, 1956.)

681.142 365
A One-Microsecond Adder using One-Megacycle Circuitry—A. Weinberger and J. L. Smith. (IRE TRANS, vol. EC-5, pp. 65–73; June, 1956. Abstract, PROC. IRE, vol. 44, p. 1083; August, 1956.)

681.142 366
High-Speed Computer stores 2.5 Megabits—W. N. Papian. (*Electronics*, vol. 30, pp. 162–167; October 1, 1957.) Design and performance details of the Lincoln TXT2 computer are given.

681.142 367
An Iterative Analogue Computer for Use with Resistance Network Analogues—I. C. Hutcheon. (*Brit. J. Appl. Phys.*, vol. 8, pp. 370–373; September, 1957.)

681.142 368
Circuitry and Characteristics of a Repeating Electronic Analogue Computer—W. Dhen. (*Elektrotech. Z., Edn A*, vol. 78, pp. 490–494; July 11, 1957.) Description of a computer for solving engineering problems. A hyperbolic-field tube [2924 of 1956 (Schmidt)] is used as multiplier.

681.142 369
The Error Effect of the Operation Amplifier in Analogue Computers—A. Kley. (*Telefunken Ztg*, vol. 30, pp. 136–141; June, 1957. English summary, p. 153.) The various errors and their sources are discussed and some methods of automatic correction are described.

681.142 370
A Multiplier based on the Two-Parabola Method—W. Schneider. (*Telefunken Ztg*, vol. 30, pp. 141–145; June, 1957. English summary, p. 153.) The method of operation is described and the accuracy obtainable is shown by some examples of calculations.

681.142:538.244 371
The Utilization of Domain Wall Viscosity in Data-Handling Devices—V. L. Newhouse. (PROC. IRE, vol. 45, pp. 1484–1492; November, 1957.)

681.142:621.314.7:621.375.4 372
Transistors in Current-Analogue Computing—B. P. Kerfoot. (IRE TRANS., vol. EC-5, pp. 86–93; June, 1956.) A comparison of variable-current and variable-voltage analog computing techniques shows that transistors are particularly suitable for the former. Low-power low-frequency transistors in direct-coupled amplifiers which may be used in computers are described.

681.142:621.318.57 373
Electric Correlators—K. Steinbuch and H. Endres. (*Nachr. Tech. Z.*, vol. 10, pp. 277–287; June, 1957.) Typical applications of electric correlating devices are discussed, and the operation of various types of these is described. Details are given of a static-type translator using transistors and crystal diodes as nonlinear switching elements.

681.142:621.374.33 374
A Time-Division Multiplier—M. L. Lilamand. (IRE TRANS., vol. EC-5, pp. 26–34. March, 1956. Abstract, PROC. IRE, vol. 44, pt. 1, p. 832; June, 1956.)

681.142:621.385.5 375
Analogue Multipliers and Squarers using a Multigrid Modulator—R. L. Sydnor, T. R. O'Meara, and J. Strathman. (IRE TRANS., vol. EC-5, pp. 82–85. June, 1956. Abstract, PROC. IRE, vol. 44, p. 1084; August, 1956.)

681.142:621.395.625.3 376
A Small Coincident-Current Magnetic Memory—W. J. Bartik and T. H. Bonn. (IRE TRANS., vol. EC-5, pp. 73–78; June, 1956. Abstract, PROC. IRE, vol. 44, p. 1083; August, 1956.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.049.75 377
The Role of Printed Wiring in High Fidelity—N. H. Crowhurst. (*Audio*, vol. 41, pp. 17–20, 78; May, 1957.) The advantages of printed-wiring techniques are summarized.

621.314.22 378
A Design Method for Wide-Band Balanced and Screened Transformers in the Range 0.1–200 Mc/s—M. M. Maddox and J. D. Storer. (*Electronic Eng.*, vol. 29, pp. 524–531; November, 1957.) Leakages and winding capacitances

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are accurately and independently controlled, and a high degree of winding balance is attained with an insertion loss of about 0.5 db.

621.314.22 379
Simplified Pulse Transformer Design—J. H. Smith. (*Electronic Eng.*, vol. 29, pp. 551–555; November, 1957.) An outline of design techniques for low- and high-power pulse transformers.

621.316.726:621.396.62 380
The Frequency-Lock A.F.C. Circuit—R. Leek. (*Proc. IEE*, pt. B, vol. 104, pp. 587–597; November, 1957.) Noise-free operation of the system is analyzed and its performance at low snr is considered. A theory of the system operation under the latter conditions is developed from an examination of the tracking errors resulting from noise and changes in signal frequency. Practical results obtained are in agreement with this theory.

621.318.57 381
A Novel Electronic Transmit-Receive Switch—A. Sabaroff. (*QST*, vol. 41, pp. 24–25, 162; June, 1957.) A Type 6AH6 tube, capable of withstanding 250 v between grid and cathode, isolates the receiver from the transmitter tank circuit.

621.318.57:621.314.7 382
Stabilization of Current-Operated Transistor Switching Circuits—N. W. Morgalla. (*A.T.E.J.*, vol. 13, pp. 192–200; July, 1957.) A method is given for calculating the parameters of simple compensating circuits.

621.318.57:621.314.7:621.395.3 383
The Transistor as a Speech-Path Switch—R. C. N. Mundy. (*A.T.E.J.*, vol. 13, pp. 227–235; July, 1957.) The suitability of transistors is discussed and some methods of using them for this purpose in telephone systems are described.

621.318.57.01:681.142 384
Complexity in Electronic Switching Circuits—D. E. Muller. (IRE Trans., vol. EC-5, pp. 15–17. March, 1956. Abstract, Proc. IRE, vol. 44, pt. 1, p. 832; June, 1956.)

621.37.049:[537.3+538.6 385
Future Circuit Aspects of Solid-State Phenomena—E. W. Herold. (Proc. IRE, vol. 45, pp. 1463–1474; November, 1957.) Several phenomena such as superconductivity, molecular amplification, magnetic effects in semiconductors, and the nonlinear capacitance of *p-n* junctions are discussed from the point of view of the circuit designer. New devices will, to an increasing extent, be based on controlled inhomogeneity.

621.372:538.565.5 386
The Influence of Electromagnetically Coupled Systems—W. Dahlke. (*Arch. elekt. Übertragung*, vol. 11, pp. 231–238; June, 1957.) Compound systems consisting of a primary electromagnetically coupled to a secondary system are considered. The equations obtained are applied to examples of magnetic coupling and a conducting diode.

621.372.092:621.318.134:621.396.677 387
A New Technique in Ferrite Phase Shifting for Beam Scanning of Microwave Antennas—F. Reggia and E. G. Spencer. (Proc. IRE, vol. 45, pp. 1510–1517; November, 1957.) In the device described a longitudinal magnetic field is applied to a ferrite rod in a rectangular waveguide excited in the TE_{10} mode. Phase shifts of over 250° per inch and variations in transmitted power of less than ±0.2 db have been obtained with a field of 60 oersteds. See also 44 of 1957 (Scharfman).

621.372.412.011.2 388
Computation of Crystal Admittance—W. J. Lucas and P. B. Barber. (*Electronic Radio Eng.*, vol. 34, pp. 454–548; December, 1957.) The results of a digital-computer program designed to calculate the admittance of a coaxial crystal for various values of video resistance, spreading resistance, and barrier capacitance over the range 2–18 kmc are given. Comparisons with available measurements are made.

621.372.413 389
Method of Obtaining Pressure- and Temperature-Insensitive Microwave Cavity Resonators—C. M. Crain and C. E. Williams. (*Rev. Sci. Instr.*, vol. 28, pp. 620–623; August, 1957.) Methods using invar walls with steel or brass end plates for TE_{011} cavities are unsatisfactory due to hysteresis effects. Improved techniques are described for fixed-frequency resonators which have temperature coefficients less than ±0.2 in 10^6 per °C. The pressure coefficient is less than 0.003 in 10^6 per millibar.

621.372.413:621.318.134 390
Retardation Effects Caused by Ferrite Sample Size on the Frequency Shift of a Resonant Cavity—J. E. Tompkins and E. G. Spencer. (*J. Appl. Phys.*, vol. 28, pp. 969–974; September, 1957.) Expressions are derived, using perturbation theory, for the frequency shift of a circularly polarized resonant microwave cavity due to insertion of a small ferrite sample.

61.372.51 391
Simplified Design of Impedance-Matching Networks—G. Grammer. (*QST*, vol. 41, pp. 38–42, 32–35, and 29–34; March–May, 1957.) A step-by-step method of impedance transformation is described. The formation of Π and T networks from the basic L section and some complex matching networks for particular applications are discussed.

621.372.543.2 392
Super Selectivity with Crystals—R. F. Burns. (*Radio Ü Telev. News*, vol. 58, pp. 52–53, . . . curve; July, 1957.) Constructional details of a lattice-type IF filter with two crystals; the response curve is symmetrical.

621.372.56.029.6:621.372.8 393
A Frequency-Independent Microwave Attenuator with Largely Constant Phase Shift—R. Steinhart. (*Nachr. Tech. Z.*, vol. 10, pp. 294–297; June, 1957.) A waveguide attenuator of the rotary-vane type is described and calibration and error curves are given. Close agreement of measured attenuation with theoretical values was obtained in attenuators for the range 3300–4200 mc.

621.372.57 394
The Transactor—A. W. Keen. (*Electronic Radio Eng.*, vol. 34, pp. 459–461; December, 1957.) The two-terminal constant-current and constant-voltage generators used in equivalent circuits of active networks are replaced by transmission-type active elements called transactors. The relations between variants of this element are established.

621.372.6:621.3.018.1 395
Theory of Two-Phase Networks—G. Wunsch. (*Nachr. Tech.*, vol. 7, pp. 200–205; May, 1957.) Phase-splitting networks as used in SSB modulation systems are discussed and calculations by approximation methods are described.

621.373.4.029.4:621.396.963.5 396
Phase-Shift Oscillator Indicates Radar Range—R. C. Barritt. (*Electronics*, vol. 30, pp. 160–161; October 1, 1957.) Firing range information is conveyed by a variable af note.

621.373.4.029.64 397
High-Order Harmonics for X-Band Oscillator Stabilization—M. C. Thompson and J. V. Cateora. (*Rev. Sci. Instr.*, vol. 28, p. 656; August, 1957.) A simplified phase-locking stabilization technique for microwave oscillators. Harmonic orders up to about 300 are used.

621.373.421 398
The Simultaneous Generation of Two Oscillations in One Oscillator and the Stability of the Difference Frequency—W. Feist. (*Nachr. Tech. Z.*, vol. 10, pp. 215–222; May, 1957.) The operating conditions are determined which are required for the simultaneous generation of two different frequencies in a tube feedback oscillator. [See also 80 of 1952 (Herzog).] Experimental results are discussed.

621.373.43 399
"Grid-Diode" Sawtooth Generator—T. A. Mendes. (*Wireless World*, vol. 63, pp. 603–606; December, 1957.) Description of a simple time base circuit with wide frequency range.

537.311.33:538.632:621.373.5 400
Experimental and Theoretical Investigation of Semiconductor Hall-Effect Generators—Strutt and Sun. (See 492.)

621.373.52 401
Silicon Transistor Crystal Oscillators have High Temperature Stability—E. G. Homer. (*Electronics*, vol. 30, pp. 218–222; October 1, 1957.) Outline of circuit design considerations and discussion of experimental results.

621.374.32 402
Difference Counters—A. F. Fischmann. (*Electronic Eng.*, vol. 29, pp. 546–550; November, 1957.) Two examples are given with maxima counting rates 1) 2×10^6 pulses per second and 2) 10^7 pulses per second. The design of a decimal difference counter is outlined.

621.375.2.024 403
Direct-Coupled Amplifiers—D. J. R. Martin. (*Electronic Radio Eng.*, vol. 34, pp. 438–441; December, 1957.) "A method is described of artificially matching tubes to obtain improved mutual compensation for the effects of normal heater-supply voltage changes. Adjustment is easier than selecting naturally matched pairs of tubes, and considerably better balance is obtained."

621.375.221.029.62 404
A 70-Mc/s I.F. Amplifier for Wide-Band Microwave Links—L. J. Herbst and G. R. Shoubridge. (*A.T.E.J.*, vol. 13, pp. 184–191; July, 1957.)

621.375.4 405
Direct-Coupled Amplifiers with Junction Transistors—S. Giustini. (*Alta Frequenza*, vol. 26, pp. 196–225; August, 1957.) A balanced dc amplifier circuit is analyzed and the theory of its operation is developed on the basis of the work by Ebers and Moll (884 of 1955). The problem of balancing the circuit with ideal and with commercial-type transistors is considered and practical design formulas are derived.

621.375.9:538.569.4 406
Masers and Reactance Amplifiers—Basic Power Relations—B. Salzberg. (Proc. IRE, vol. 45, pp. 1544–1545; November, 1957.) An alternative derivation, based on circuit theory, of the expressions of Manley and Rowe (2988 of 1956). The power relations involve only the source and load frequencies and are independent of the specific characteristics of the nonlinear device.

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621.375.9:538.569.4:621.317.3.029.64 407
Maser Noise Measurement—Helmer. (See 547.)

621.375.9.029.63/.64:538.221 408
New Ferrite Microwave Amplifier—(*Bell Lab. Rec.*, vol. 35, pp. 316–317; August, 1957.) An experimental solid-state amplifier using a ferrite material as the active element is described. The device is suitable as an amplifier of very weak microwave signals. See also 3076 of 1957 (Suhl).

621.375.9.029.64:538.569.4 409
The Reaction Field and its Use in Some Solid-State Amplifiers—P. W. Anderson. (*J. Appl. Phys.*, vol. 28, pp. 1049–1053; September, 1957.) The theory of the maser is presented in terms of the radiation "reaction" field produced in a cavity or waveguide by the presence of the electromagnetic moment of the sample being investigated. Two amplifiers are discussed which use the reaction field in different ways from the usual solid-state maser.

621.376.32:621.314.63 410
The *PN*-Junction on a Variable-Reactance Device for F.M. Production—D. C. Brown and F. Henderson. (*Electronic Eng.*, vol. 29, pp. 556–557; November, 1957.) A transistor modulator for a 10-mc carrier frequency is described which gives good fm with negligible AM. 0.15-v change in modulating signal produces a deviation of 100 kc.

GENERAL PHYSICS

537.226.2 411
The Influence of a Strong Magnetic Field on the Dielectric Constant of a Diamagnetic Fluid—A. D. Buckingham. (*Proc. Phys. Soc.*, vol. 70, pp. 753–760; August 1, 1957.)

537.226.2 412
Formulae for Dielectric Constant of Mixtures—J. A. Raynolds and J. M. Hough. (*Proc. Phys. Soc.*, vol. 70, pp. 769–775; August 1, 1957.) Fifteen formulas are tabulated with references and shown to be special cases of a general fundamental formula. An exception is the Lichtenecker type of formula, possibly owing to incorrect assumptions.

537.311 413
The Second-Order Effect of Free Electrons on Lattice Conduction—I. C. Pyle. (*Proc. Camb. phil. Soc.*, vol. 53, pt. 2, pp. 508–513; April, 1957.) "Second-order perturbation theory, in conjunction with the usual treatment of electron-phonon interaction, allows us to calculate the correction to the first-order result for the scattering of phonons by electrons. It is shown that the second-order term is much smaller and, therefore, negligible. This justifies the use of the first-order theory in the treatment of the interaction in metals and semiconductors." See also 2017 of 1956 (Ziman).

537.311.1 414
A Method of Calculation of Electrical Conductivity—H. Nakano. (*Progr. Theoret. Phys.*, vol. 17, pp. 145–161; February, 1957.) Using relaxation theory a general solution is obtained for the electric current density; irreversibility is then introduced so that practical cases may be considered. A perturbation approximation gives the same result as that based on the Bloch theory.

537.312.62 415
A Note on the Energy-Gap Model of Superconductivity—M. J. Buckingham. (*Nuovo Cim.*, vol. 5, pp. 1763–1765; June 1, 1957. In English.)

537.312.62 416
Gauge Invariance and the Energy-Gap Model of Superconductivity—J. Bardeen. (*Nuovo Cim.*, vol. 5, pp. 1766–1768; June 1, 1957. In English.) See also 415 above.

537.321 417
On the Elementary Theory of Thermoelectric Phenomena—R. Stratton. (*Brit. J. Appl. Phys.*, vol. 8, pp. 315–321; August, 1957.) All the relations between the thermoelectric parameters are deduced from physical properties rather than by means of mathematical transformations. The sources of the heat and emf developed are indicated.

537.525:538.56 418
The Plasma Resonator—A. Dattner. (*Ericsson Tech.*, vol. 13, no. 2, pp. 309–350; 1957.) The interaction between an ionized gas column and an electromagnetic wave in a waveguide is studied at $\lambda = 6$ cm. Effects examined include the influence of the discharge current, gas collisions, decay times of electron density, and modulation of the rf power.

537.525:621.387.032.435.4 419
New Effects of an Auxiliary Electrode on a Discharge at Low Pressure—Thong Saw Pak. (*J. Electronics Control*, vol. 3, pp. 471–480; November, 1957.) An auxiliary electrode close to the cathode provides a means of sensitive control of the discharge. Optimum operational conditions were found for a range of discharge parameters. Low-frequency noise may be suppressed by an electrode of wire gauze.

537.533.8 420
Theory of Secondary Emission—R. G. Lye and A. J. Dekker. (*Phys. Rev.*, vol. 107, pp. 977–981; August 15, 1957.) The elementary theory of secondary electron emission has been generalized and modified to incorporate results of recent measurements of the range energy relation and the dissipation of energy by slow electrons in solids. These modifications give considerably improved agreement between the theoretical and experimental "universal" reduced yield curves.

538.1 421
"Satellite-Electron" Theory of Ferromagnetism, Antiferromagnetism and Related Phenomena—L. Singh. (*Naturwiss.*, vol. 44, pp. 417–418; August, 1957. In English.) A theory based on classical concepts is developed to show the electronic nature of ferromagnetism. Magnetic properties of transition-metal oxides are briefly discussed.

538.11:538.124 422
A Note on the Ground State of Antiferromagnetism—H. Taketa and T. Nakamura. (*J. Phys. Soc. Japan*, vol. 11, pp. 919–923; September, 1956.) Kasteleijn's method (2199 of 1952) of obtaining the lowest energy state is generalized for the cases of two- and three-dimensional lattices.

538.3:52 423
On the Reflection and Refraction of Magnetohydrodynamic Waves—S. Prakash and J. N. Tandon. (*Proc. Nat. Inst. Sci. India*, pt. A, vol. 23, pp. 264–273; July 26, 1957.)

538.566:535.42 424
Asymptotic Formulas for Diffraction by Parabolic Surfaces—H. Hochstadt. (*Commun. pure appl. Math.*, vol. 10, pp. 311–329; August, 1957.) Asymptotic solutions to the wave equation are found at a boundary consisting of a paraboloid of revolution or a parabolic cylinder. The results are applied to the reflection at such surfaces of incoming plane waves or diverent waves from a point or line on the axis of the surface.

538.569.4 425
Double Modulation System for Narrowing Electron Resonance Absorption Lines—R. R. Unterberger, J. L. Garcia de Quevedo, and A. E. Stoddard. (*Rev. Sci. Instr.*, vol. 28, pp. 616–619; August, 1957.) A method for observing electron-resonance absorption on an oscilloscope screen by modulating both the magnetic field and the klystron frequency. The apparent line width can be reduced by a factor of 17 relative to the true line width.

538.569.4 426
Microwave Studies of the Internal Motion and the Structure of Methyl Amine—T. Nishikawa. (*J. Phys. Soc. Japan*, vol. 12, pp. 668–680; June, 1957.)

538.569.4 427
Potential Barrier and Molecular Structure of Methyl Mercaptan from its Microwave Spectra—T. Kojima and T. Nishikawa. (*J. Phys. Soc. Japan*, vol. 12, pp. 680–686; June, 1957.)

538.569.4 428
Microwave Spectrum of BrCN and Dependence of Quadrupole Coupling Constant on the Vibrational State—T. Oka and H. Hirakawa. (*J. Phys. Soc. Japan*, vol. 12, pp. 820–823; July, 1957.)

538.569.4:621.375.9 429
Proposal for a Nuclear Quadrupole Maser—R. Braunstein. (*Phys. Rev.*, vol. 107, p. 1195–1196; August 15, 1957.) It is suggested that certain substances exhibiting pure quadrupole transitions have the requisite properties for maser operation. The differences between masers based on these substances and paramagnetic solids are discussed.

538.569.4:621.375.9.029.6 430
Fluctuations in Amplification of Quanta with Application to Maser Amplifiers—K. Shimoda, H. Takahasi, and C. H. Townes. (*J. Phys. Soc. Japan*, vol. 12, pp. 686–700; June, 1957.) Expressions for the probability distribution of quanta for the average values and for the fractional fluctuation are developed and applied to maser-type amplifiers.

538.569.4:621.375.9.029.64 431
The Reaction Field and its Use in Some Solid-State Amplifiers—Anderson. (See 409.)

538.615 432
Radio-Frequency Zeeman Effect in O_2—J. M. Hendrie and P. Kusch. (*Phys. Rev.*, vol. 107, pp. 716–723; August 1, 1957.) By the molecular-beam magnetic-resonance method the ratio of the g value of the rotational magnetic moment to that of the electron-spin moment is found to be $(6.08 \pm 0.74) \times 10^{-5}$; the ratio of the g value of the unpaired-electron spin moments in O_2 to that of the free-electron spin moment is $1 = (190 \pm 13) \times 10^{-6}$.

539.16.08:621.372.413 433
The Cloud Chamber—(*Electronic Radio Eng.*, vol. 34, pp. 447–450; December, 1957.) The theory and application of the Wilson cloud chamber are outlined, and its modification by Gabor and Hampton (*Nature, London*, vol. 180, pp. 746–749; October 12, 1957.) is briefly described.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16:523.4 434
Sources of Radio Noise on the Planet Jupiter—C. H. Barrow, T. D. Carr, and A. G. Smith. (*Nature, London*, vol. 180, p. 381; August 24, 1957.) Daily records made from December 31, 1956 to March 8, 1957 at Gainesville, Florida at 18 mc suggest that there are two localized sources on Jupiter and that Jupiter is surrounded by an ionosphere of electron density 10^8/cm^3, *i.e.*, comparable with

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that of the earth's ionosphere. See 2710 of 1956 (Shain).

523.16:523.64 435
Radio Emission from the Comet 1956 h—H. G. Müller, W. Priester, and G. Fischer. (*Naturwiss.*, vol. 44, pp. 392–393; July, 1957.) Report of observations carried out at the radio observatory Stockert.

523.16:551.510.535 436
A Consideration of Radio-Star Scintillations as Caused by Interstellar Particles Entering the Ionosphere: Part 3—The Kind, Number and Apparent Radiant of the Incoming Particles—G. A. Harrower. (*Can. J. Phys.*, vol. 35, pp. 792–798; July, 1957.) Discussion of results given earlier (106 of 1958) indicates that the interstellar particles must be hydrogen atoms.

523.7 437
The Proportion of Umbra in Large Sunspots, 1878–1954—A. W. F. Edwards. (*Observatory*, vol. 77, pp. 69–70; April, 1957.) The ratio of umbral area to whole-spot area is evaluated and shown to vary significantly with the position of the sunspot in the solar cycle.

523.7:621.396.822.029.66 438
Detection of Submillimetre Solar Radiation—H. A. Gebbie. (*Phys. Rev.*, vol. 107, pp. 1194–1195; August 15, 1957.) Preliminary results are presented which indicate regions of transmission in the wavelength range 1 mm–300 μ.

523.72:621.396.822 439
Short Time Transients in Solar Noise—T. de Groot. (*Nature, London*, vol. 180, p. 382; August 24, 1957.) Histograms have been made from observations at Dwingeloo, Holland, at 400 mc. Reber's theory (see 1637 of 1955) that there is a linear relation between wavelength and duration of "pips" is not confirmed at this frequency.

523.746 440
A Review of Recent Investigations into Sunspot Cycles—N. A. Huttly. (*Marconi Rev.*, vol. 20, pp. 117–129; 4th Quarter, 1957.) Summaries of and comments on selected published papers and CCIR documents relating to predictions of solar activity for use in ionospheric work.

523.75 441
Occultation of a Radio Source by the Solar Corona—O. B. Slee. (*Observatory*, vol. 76, pp. 228–231; December, 1956.) Measurements of radio noise flux from Taurus A, when its angular separation from the sun is very small, suggest that partial wide-angle scattering has occurred due to electron density irregularities in the solar corona.

523.75 442
A Solar Flare on 1956 November 7—P. A. Wayman. (*Observatory*, vol. 77, pp. 24–26; February, 1957.) Visual and radio observations during the flare are presented. Anomalies in the radio observations are accounted for by small regions of the flare remaining enhanced for several hours after the main effect had subsided.

523.78 443
Centimetre-Wave Observations of the Solar Eclipse of 1954 June 30—J. S. Hey and V. A. Hughes. (*Observatory*. vol. 76, pp. 226–228; December, 1956.) Radiation measurements at 10.5 cm λ disclose an ellipticity in the brightness distribution of the sun's disk with the greatest extent in the equatorial plane.

550.3:519.24 444
Proper and Improper Use of Statistics in Geophysics—B. Kinsman. (*Tellus*, vol. 9, pp. 408–418; August, 1957.) A discussion of some of the peculiarities of the geophysical sciences together with their implications for statistical methods, particularly for the correlation coefficient, is given. The discussion is illustrated by an analysis of a recently published paper. In conclusion some advice on the use of correlations is offered.

550.384 445
Hydromagnetics and the Earth's Inner Core—G. H. A. Cole. (*Observatory*, vol. 77, pp. 17–19; February, 1957.) It is suggested that the terrestrial magnetic field is produced in an inner liquid core permeated by a magnetic field of $\sim 10^5$ G. Secondary motions, partly hydromagnetic, occur in an outer surrounding liquid region and the field emerging from the complete core is the small field measured at the earth's surface.

550.385.1:523.7 446
Method of Magnetic-Storm Forecasting from the Activities of Flares Accompanied by the Solar Radio Noise Outbursts—K. Sinno. (*J. Radio Res. Labs., Japan*, vol. 4, pp. 267–276; July, 1957.) A statistical examination has been made of the occurrence of magnetic storms, solar flares, and radio noise outbursts on 200 mc. The correlations discovered are applied to the problem of forecasting magnetic storms during the IGY.

550.389.2:551.510.535(54) 447
Ionospheric Studies in India during the International Geophysical Year—K. R. Ramanathan. (*J. Inst. Telecommun. Eng. India*, vol. 3, pp. 193–197; June, 1957.)

550.389.2:621.396.11 448
Propagation and the International Geophysical Year—G. W. Slack. (*R.S.G.B. Bull.*, vol. 33, pp. 8–11; July, 1957.) Summary of phenomena which affect hf and vhf propagation. Some methods of observation are suggested to the radio amateur.

550.389.2:629.19 449
Artificial Satellites of the Earth—(*Wireless World*, vol. 63, pp. 574–578; December, 1957.) Extracts are given from *Radio, Moskow* published before the launching of the U.S.S.R. satellites [see, *e.g.*, 3860 of 1957 (Vakhnin)] and results of observations made at various stations in the U.K. are summarized.

550.389.2:629.19:681.142 450
Tracking the Man-Made Satellite—M. Gunther. (*Radio Ü Telev. News*, vol. 58, pp. 31–33; July, 1957.) An outline of the preparations for launching and observing the U.S. satellite, including the use of an electronic computer for calculating the orbit from observational data.

551.501.8:621.396.11 451
The Use of Surface Weather Observations to Predict the Total Atmospheric Bending of Radio Rays at Small Elevation Angles—Bean and Cahoon. (See 578.)

551.508.7 452
The Detection and Measurement of Water Droplets—H. F. Liddell and N. W. Wootten. (*Quart. J. Roy Meteorol. Soc.*, vol. 83, pp. 263–266; April, 1957.) A method of obtaining a permanent record of drop size is described; it can be used for droplets of less than 1-micron diameter.

551.510.5 453
The Threefold Structure of the Atmosphere and the Characteristics of the Tropopause—F. Defant and H. Taba. (*Tellus*, vol. 9, pp. 259–274; August, 1957.) The atmosphere in the northern hemisphere is divisible into three regions, north of the polar-front jet, south of the subtropical jet and the region between the jets. Each region has a typical vertical temperature structure in the stratosphere and troposphere and a characteristic tropopause height.

551.510.535 454
The Effect of the Equatorial Electrojet on the Ionospheric E_s and F_2 Layers—N. J. Skinner and R. W. Wright. (*Proc. phys. Soc.*, vol. 70, pp. 833–839; September 1, 1957.) The occurrence of low values of fE_s at Ibadan is correlated with variations in the horizontal and vertical components of the earth's field and with F_2-layer parameters. An immediate relation is indicated among the electrojet, the production of equatorial E_s, and vertical drifts in the F_2 layer.

551.510.535 455
Dynamical Structure of the Ionospheric F_2 Layer—T. Shimazaki. (*J. Radio Res. Labs., Japan*, vol. 4, pp. 309–332; July, 1957.) Formulas giving diffusion velocity and its divergence are derived for the electron-ion gas in the ionosphere on the assumption of a linear temperature gradient. The change in electron density distribution is then calculated, using these formulas for different models of the ionosphere. Comparisons of the calculated and observed changes in the F_2 layer show that Bradbury's rather than Chapman's theory is more appropriate.

551.510.535(52) 456
On the Occurrence of the $F_{1.5}$ Layer in Japan: Part 1—I. Kasuya. (*J. Radio Res. Labs., Japan*, vol. 4, pp. 291–300; July, 1957.) A statistical analysis based on vertical incidence soundings over a sunspot cycle shows annual seasonal and diuranal variations. Good correlation is demonstrated between frequency of occurrence and sunspot number.

551.524.7+551.571.7 457
Some Further Observations from Aircraft of Frost Point and Temperature up to 50000 ft—N. C. Helliwell, J. K. Mackenzie, and M. J. Kerley. (*Quart. J. Roy Meterol. Soc.*, vol. 83, pp. 257–262; April, 1957.) Measurements at various altitudes taken in 46 flights over Southern England in 1955 are tabulated. Good general agreement is obtained with 1954 measurements.

551.578.13:551.501.81 458
Variation with Height of Rainfall below the Melting Level—W. G. Harper. (*Quart. J. Roy. Meterol. Soc.*, vol. 83, pp. 368–371; July, 1957.) 3-cm radar measurements in warm-front conditions suggest that there is uniformity of drop size distribution and rate of rainfall with height once meeting is complete and terminal velocity has been attained.

551.59 459
On a Special Aspect of the Condensation Process and its Importance in the Treatment of Cloud Particle Growth—C. Rooth. (*Tellus*, vol. 9, pp. 372–377; August, 1957.) The distribution of water vapor around a droplet depends upon the ratio of droplet size to a factor determined by the equilibrium rate of exchange of molecules between the water surface and the water vapor.

551.59 460
Electric Charge Separation in Subfreezing Cumuli—S. Twomey. (*Tellus*, vol. 9, pp. 384–393; August, 1957.) A theory of the electrification of thunderclouds based on charge separation occurring when supercooled droplets impinge on ice particles. The theory can explain observed charge separations without assuming the presence of large hailstones.

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551.594.21 461
The Measurement of World-Wide Thunderstorm Activity at a Single Locality—G. A. Isted. (*Marconi Rev.*, vol. 20, pp. 130–132; 4th Quarter, 1957.) Counts of atmospheric impulses received in England in the 0.3–12-kc band are found to be similar to records of integrated noise in California in the band 25–130 cps [see 2070 of 1956 (Holzer and Deal)]. It is inferred that world-wide thunderstorm activity was recorded in both experiments.

551.594.221 462
Preliminary Discharge Processes in Lightning Flashes to Ground—N. D. Clarence and D. J. Malan. (*Quart. J. Roy. Meterol. Soc.*, vol. 83, pp. 161–172; April, 1957.) The first return stroke is usually preceded by three successive and distinct discharge processes lasting several hundred milliseconds. The characteristics of the field changes in these preliminary stages are analyzed in terms of the distance of the source. Probable mechanisms are discussed.

551.594.6 463
The Accuracy of the Determination of Reflection Heights and Distances of Atmospherics on the basis of their Waveform—G. Skeib. (*Z. Met.*, vol. 11, pp. 129–135; May/June, 1957.) The errors in calculating reflection height H and distance D of atmospherics are investigated as a function of the ratio H/D and of the reflection modes. A recorded waveform is analyzed and results are tabulated. See also 1762 of 1957 and 122 of 1956 (Horner and Clarke).

551.594.6:621.396.029.4 464
Diurnal Variation in the Occurrence of "Dawn Chorus"—J. H. Pope. (*Nature, London*, vol. 180, p. 433; August 31, 1957.) A histogram of observations made at frequencies between 1 and 10 kc, on 116 days between January and July, 1956, at College, Alaska, shows a maximum at 1400 hours local time which appears to be related to geomagnetic latitude.

LOCATION AND AIDS TO NAVIGATION

621.396.9:621.396.11.029.55 465
Direct-Vision-Type Direction Finder for High Frequency—K. Miya, T. Sasaki, M. Ishikawa, and S. Matsushita. (*Rep. Ionosphere Res. Japan*, vol. 11, pp. 1–10; March, 1957.) The bearing of an hf signal is indicated directly on a cr tube by a bright radial line at the corresponding angle. A sensitivity 40 db higher than that of conventional equipment is achieved by deriving the cr tube deflection voltages from the changes in the rectified output of a receiver when different antenna combinations are connected to its input in sequence. The high sensitivity permits directional observations on weak scattered signals. Circuit diagrams are given and sources of error are discussed.

621.396.962.33 466
Two Doppler Navigators—(*Brit. Commun. Electronics*, vol. 4, pp. 551–553; September, 1957.) Brief details of airborne equipments for use by military and commercial aircraft.

621.396.962.33 467
Radar Drift Measurement using Doppler Techniques—P. L. Stride. (*Brit. Commun. Electronics*, vol. 4, pp. 554–557; September, 1957.) A drift indicator is described which includes all the control and display circuits necessary to adapt a conventional search radar to measure drifts. Some results of recent trials are discussed.

621.396.963.5:621.373.4.029.4 468
Phase-Shift Oscillator Indicates Radar Range—Barritt. (See 397.)

621.396.969.33 469
Errors in Radar Navigation—T. Stuland. (*J. Inst. Nav.*, vol. 10, pp. 390–396; October, 1957.) The possible maximum errors are considered which may be contributory factors to collisions between ships. Alteration in course may not, by itself, be sufficient and should be combined with speed reduction so that there is more time in which information is obtained.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.583:621.385.032.14 470
The Oxidation of Evaporated Barium Films (Getters)—R. N. Bloomer. (*Brit. J. Appl. Phys.*, vol. 8, pp. 321–329; August, 1957.) Mott's theory of oxidation (see, *e.g.*, 439 of 1948) is applied to measurements of the main quantities important to the gettering process. The formation of the first monolayer of oxide film is explained. Below a critical temperature of 40°C a protective oxide film grows to a thickness of about 50Å only, while above 40°C Ba films are oxidized right through. See also 2140 of 1957.

533.583:621.385.032.14 471
Barium Getters and Carbon Monoxide—R. N. Bloomer. (*Brit. J. Appl. Phys.*, vol. 8, pp. 352–355; September, 1957.) An experimental study of the absorption of CO by evaporated Ba films indicates that the CO is dissociated at the surface of the Ba film, forming an oxide layer containing free carbon. The critical temperature above which the oxide layer ceases to be protective is 80°C.

535.215:537.533 472
Photoelectric Emission from Barium Oxide—H. R. Philipp. (*Phys. Rev.*, vol. 107, pp. 687–693; August 1, 1957.) Photoelectric emission measurements were made on sprayed coatings of BaO in several states of thermionic activity and at different temperatures. The spectral distribution of the photoelectric yield shows a rise at a quantum energy of 3.8 ev ascribed to exciton-induced emission, and another rise at 5 ev attributed to electrons ejected from the filled band. Analysis of the energy distribution of emitted electrons shows that different emission mechanisms are operative at low and high incident quantum energies.

535.37:546.472.21 473
The Measurement of the Optical Properties of Zinc Sulphide—C. K. Coogan. (*Proc. phys. Soc.*, vol. 70, pp. 845–861; September 1, 1957.) A comprehensive description of apparatus and technique and a discussion of results. A comparison is made with other work on the optical properties of ZnS. A band at about 330 mμ is found in the absorption spectrum, particularly at liquid air temperatures, together with a corresponding region of anomalous dispersion.

535.37:546.472.21 474
Models for Different Types of Traps in Zinc Sulphide Phosphors. Thermal and Optical Liberation of Trapped Electrons—D. Curie. (*J. Phys. Radium*, vol. 18, pp. 214–222; April, 1957.) Models for shallow and deep traps are considered and activation energies are calculated. Forty references.

535.376:546.472.21 475
Mechanism of Electroluminescence of Zinc Sulphide—R. Goffaux. (*J. Phys. Radium*, vol. 18, pp. 1–4; January, 1957.)

537.226/.227:[546.431.824-31+546.42.824-31 476
Solid-State Reaction between Barium Titanate and Strontium Titanate—S. Nomura. (*J. Phys. Soc. Japan*, vol. 11, pp. 924–929; September, 1956.) The reaction is induced by sintering. Measurements were made of the variations with sintering conditions of the permittivity/temperature characteristics and of the crystal structure of 1:1 mixtures.

537.226/.227:546.431.824-31 477
Some Aspects of Wedge-Shaped Domains in $BaTiO_3$ Crystal—T. Nakamura, W. Kinase, and Y. Kato. (*J. Phys. Soc. Japan*, vol. 12, pp. 836–837; July, 1957.) At 20°C wedge-shaped domains were observed which were presumed to have formed during the transition from cubic to tetragonal structure in cooling through the 120°C Curie point. On cooling below 5°C a marked nonreversible change in the pattern occurred.

537.226:537.52 478
Dipoles and Electric Breakdown—J. A. Kok and M. M. G. Corbey. (*Appl. sci. Res.*, vol. B6, no. 6, pp. 449–455; 1957.) The breakdown of dielectric material may be caused by induced as well as permanent dipoles gathering at a place of maximum stress to form a bridge.

537.227 479
Laboratories Announce New Ferroelectric—(*Bell Lab. Rec.*, vol. 35, p. 271; July, 1957.) Some of the properties of triglycine sulphate are described. These properties make it a promising material for switching circuits and storage devices.

537.228.1 480
Coupling Coefficient and the Electromechanical Efficiency of Piezoelectric Materials—Y. Le Corre. (*J. Phys. Radium*, vol. 18, pp. 51–58; January, 1957.) A mathematical analysis is given defining the electromechanical coupling coefficients and efficiencies of unidimensional and orthorhombic crystals. A hollow cylindrical piezoelectric detector is also considered.

537.311.33 481
Vasileff's Calculation of Electronic Self-Energy in Semiconductors—E. N. Adams. (*Phys. Rev.*, vol. 107, p. 671; August 1, 1957.) Vasileff's theory (2178 of 1957) fails for semiconductors of the usual type because of the predominance of processes involving virtual phonons of large wave number.

537.311.33 482
The Mass Action Laws for the Reactions between Free Carriers and Impurities in Semiconductors considering the Electron Spin—F. W. G. Rose. (*Proc. phys. Soc.*, vol. 70, pp. 801–803; August 1, 1957.) A general rule is quoted which does not require a knowledge of the partition function. The reaction between free electrons and group-V donors in Si or Ge is treated to illustrate the use of the general rule.

537.311.33 483
Low-High-Conductivity Junctions in Semiconductors—L. W. Davies. (*Proc. phys. Soc.*, vol. 70, pp. 885–889; September 1, 1957.) A theoretical study including the effect of diffusion of charge carriers near the junction. A quantitative description of the "accumulation" of minority charge carriers at low-high junctions in filaments is given.

537.311.33 484
Semiconductor Lifetime as a Function of Recombination State Density—D. H. Clarke. (*J. Electronics Control*, vol. 3, pp. 375–386; October, 1957.) "The analysis of the Shockley-Read model is extended to describe the transient behavior. The results are compared with those already given by Fan (1797 of 1954) and by Rittner, and it is concluded that the latter ceases to be valid at high recombination state densities. The relation between steady-state and transient measurements of photocon-

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ductivity is examined, and some numerical examples are presented."

537.311.33 485

Cooling of Hot Electrons by Acoustic Scattering in Degenerate Semiconductors—R. F. Greene. (*J. Electronics Control*, vol. 3, pp. 387–390; October, 1957.) "Shockley's calculation of the energy loss rate of hot electrons by acoustic scattering in semiconductors is extended to include the case of degenerate electron statistics. A simple formula is given for the loss rate in terms of the acoustic mean free time, valid for any degree of degeneracy. A relaxation time for the electron energy is expressed in terms of the acoustic mean free time, showing the effect of degeneracy."

537.311.33 486

Quasi-electric and Quasi-magnetic Fields in Nonuniform Semiconductors—H. Kroemer. (*RCA Rev.*, vol. 18, pp. 332–342; September, 1957.) In a nonuniform semiconductor, *e.g.*, one with nonuniform elastic strains or a semiconductor alloy of varying composition, the width of the energy gap varies throughout the material. This produces gradients of the band edges which differ for the conduction and valence bands, and act as "quasi-electric" fields which are not the same for holes and electrons. Examples are given to show how transistor performance can be improved by the incorporation of such fields. A description is also given of how "quasi-magnetic" fields arise when an inhomogeneity causes a shift of the location within the Brillouin zone of the energy minimum of the band.

537.311.33:[537.32+536.21 487

Thermoelectric and Thermal Properties of Semiconductors—A. Joffe. (*J. Phys. Radium*, vol. 18, pp. 209–213; April, 1957.) Discussion with reference to the application of semiconductor thermocouples for power supply.

537.311.33:537.32 488

A Simple Derivation of the Thermoelectric Voltage in a Nondegenerate Semiconductor—F. W. G. Rose, E. Billig, and J. E. Parrott. (*J. Electronics Control*, vol. 3, pp. 481–486; November, 1957.) Only concepts such as Fermi levels, field current, diffusion current, etc., are used in the derivation.

537.311.33:537.533 489

Thermionic and Semiconducting Properties of [Ag]-Cs_2O, Ag, Cs—J. E. Davey. (*J. Appl. Phys.*, vol. 28, pp. 1031–1034; September, 1957.) A description of measurements on [Ag]-Cs_2O as a thermionic emitter. Results are given for work function, conductivity, and activation energy over a wide range of temperatures.

537.311.33:538.632 490

Determination of the Impurity Concentrations in a Semiconductor from Hall Coefficient Measurements—P. A. Lee. (*Brit. J. Appl. Phys.*, vol. 8, pp. 340–343; August, 1957.) Impurity concentrations are found from an equation relating carrier concentration to temperature and acceptor and donor concentrations. The experimental method is applied to *p*-type Si.

537.311.33:538.632 491

The Relationship between the Hall Coefficient and the Resistivity of Semiconductors, taking Various Scattering Mechanisms of the Charge Carriers into Account—T. Fukuroi and C. Yamanouchi. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, vol. 9, pp. 267–272; August, 1957.) In addition to lattice and ionized-impurity scattering, neutral impurity and dislocation scattering become appreciable at low temperatures. The relation between Hall coefficient and resistivity is calculated for five cases in which two of these four mechanisms are combined. The method is similar to that used by Jones (1357 of 1951) for combined lattice and ionized-impurity scattering.

537.311.33:538.632:621.373.5 492

Experimental and Theoretical Investigation of Semiconductor Hall-Effect Generators—M. J. O. Strutt and S. F. Sun. (*Arch. elekt. Übertragung*, vol. 11, pp. 261–265; June, 1957.) DC amplification was obtained using an experimental feedback circuit incorporating a Hall element of InSb at room temperature. An oscillator circuit covering the range 13–330 cps with an output of about 12 mw is described. Results of measurements agree with theory; the effect of temperature on oscillator performance is discussed.

537.311.33:539.23 493

Frequency Dependence of the A.C. Resistance of Thin Semiconducting Films—M. Lax and R. Sachs. (*Phys. Rev.*, vol. 107, pp. 650–655; August 1, 1957.) A modification of Howe's theory (see *Wireless Eng.*, vol. 17, pp. 471–477; November, 1940) is put forward in which the assumption of constant capacitance per unit length is dropped. The theoretical results agree with the experiments of Broudy and Levinstein (2696 of 1954); the agreement of the Howe theory with earlier experiments is shown to be partially fortuitous.

537.311.33:[546.28+546.289 494

Anisotropic Diffusion Lengths in Germanium and Silicon Crystals containing Parallel Arrays of Edge of Dislocations—R. L. Bell and C. A. Hogarth. (*J. Electronics Control*, vol. 3, pp. 455–470; November, 1957.) Techniques are described for introducing parallel arrays of edge dislocations into Ge and Si by plastic bending. Ge crystals with similar dislocations, but with a higher mean minority-carrier lifetime than those subjected to plastic bending, may be grown by pulling from the melt, using a suitably orientated dislocated seed crystal. The diffusion length is greater when measured along the dislocations than across them. A model suggests that the anisotropies observed will be more marked for crystals in which the dislocations are highly polygonized.

537.311.33:546.28 495

The Effect of Heat Treatment on the Bulk Lifetime of Excess Charge Carriers in Silicon—L. M. Nijland and L. J. van der Pauw. (*J. Electronics Control*, vol. 3, pp. 391–395; October, 1957.) Bulk lifetime in both *p*- and *n*-type crystals made by the floating zone technique and in *n*-type crystals made by the Czochralski technique is increased by annealing at 300°C to 700°C. Annealing at >700°C decreased lifetimes for *p*- and *n*-type crystals made by either method.

537.311.33:546.28 496

Radiochemical Analysis of Silicon—J. A. James and D. M. Richards. (*J. Electronics Control*, vol. 3, pp. 500–506; November, 1957.) "Results of radioactivation analyses for some 12 elements are given for silicon from two sources. The results of segregation coefficient measurements using radiochemical methods are quoted for P, Fe, Co, W, and Au."

537.311.33:546.28:535.343 497

Infrared Absorption of Oxygen in Silicon—H. J. Hrostowski and R. H. Kaiser. (*Phys. Rev.*, vol. 107, pp. 966–972; August 15, 1957.) Three infrared absorption bands have been correlated with the oxygen concentration of Si. Results obtained are explained by a model in which interstitial oxygen is bonded to two adjacent Si atoms in a nonlinear Si-O-Si unit.

537.311.33:546.28:538.63 498

Galvanomagnetic Effects in *p*-Type Silicon—D. Long. (*Phys. Rev.*, vol. 107, pp. 672–677; August 1, 1957.) Measurements are reported of the dependence on temperature and field strength of the resistance and Hall effect for several samples of *p*-type Si. The results are discussed with reference to the usual valence-band model to show that there is some inconsistency.

537.311.33:546.289 499

Recombination in Plastically Deformed Germanium—G. K. Wertheim and G. L. Pearson. (*Phys. Rev.*, vol. 107, pp. 694–698; August 1, 1957.) "Lifetimes in plastically deformed *n*- and *p*-type Ge have been measured as a function of the amount of deformation and as a function of temperature. The results indicate that the dislocations have an electron capture radius of 3.4×10^{-8} cm. The lifetime in high-purity crystals containing 10^3 to 10^4 dislocations per cm^2 may consequently be limited by recombination at dislocations."

537.311.33:546.289 500

Recombination Centres and Fast States on Unstable Germanium Surfaces—S. Wang and G. Wallis. (*Phys. Rev.*, vol. 107, pp. 947–953; August 15, 1957.) The results of measurements of surface conductance, photoconductance, dark field effect, and field effect under illumination are presented for freshly etched samples of *n*- and *p*-type Ge exposed to the Brattain-Bardeen ambient cycle. Analysis shows that the recombination centers and "fast" centers are identical.

537.311.33:546.289 501

Modulation of Light Reflected from Germanium by Injected Current Carriers—I. Filiński. (*Phys. Rev.*, vol. 107, p. 1193; August 15, 1957.) Preliminary experimental results are reported of the modulation of reflected light from the surfaces of single-crystal diodes, which arises when alternating voltages are applied to them.

537.311.33:546.289 502

Contribution of Current Carriers in the Reflection of Light from Semiconductors—L. Sosnowski. (*Phys. Rev.*, vol. 107, pp. 1193–1194; Aug. 15, 1957.) An interpretation is given of the electro-optical phenomena reported by Filiński (501 above).

537.311.33:546.289 503

On the Absorption by Free Carriers in Semiconductors—B. Donoven and N. H. March. (*Proc. phys. Soc.*, vol. 70, pp. 883–885; September 1, 1957.) The range of validity is assessed of some theoretical treatments of the properties of free carriers in alternating fields [see, *e.g.*, 2110 of 1956 (Fan, *et al*)], by examining infrared absorption in *n*-type Ge.

537.311.33:546.289 504

Field Dependence of Mobility in *p*-Type Germanium—K. S. Mendelson and R. Bray. (*Proc. phys. Soc.*, vol. 70, pp. 899–900; September 1, 1957.) A short description of results obtained on *p*-type Ge at 78°K and 195°K, using samples with carrier content between 3×10^{12} and $4\times10^{16}cm^{-3}$. A graph of conductivity as a function of electric field is shown.

537.311.33:546.289 505

Effect of Edge Dislocations on the Alloying of Indium to Germanium—J. I. Pankove. (*J. Appl. Phys.*, vol. 28, pp. 1054–1057; September, 1957.) Observations on the effects of crystal imperfections on the alloying process show that edge dislocations and other crystal disturbances enhance the dissolution of Ge in In. An interpretation of this and other effects is given.

537.311 33:546.289 506

Spiral Etch-Pits in Germanium—R. G. Rhodes, K. O. Batsford, and D. J. Dane-

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Thomas. (*J. Electronics Control*, vol. 3, pp. 403–408; October, 1957.) Discussion illustrated with photographs of spiral "terraced" pits produced by etching the {111} or {100} crystal surfaces with iodine solution. The density is 10^5—10^6 cm^{-2}; association with impurities is suggested.

537.311.33:546.289 507
On the Carrier Recombination through Nickel Impurities in Germanium—J. Okada. (*J. Phys. Soc. Japan*, vol. 12, p. 741; June, 1957.) The increase of the photoconductivity to a saturation value as the temperature decreases may be explained by assuming that Ni introduces two acceptor levels and recombination takes place through multiple levels of constant capture cross section.

537.311.33:546.289 508
Some Experiments on the Surface Field Effects in Germanium Single Crystals—M. Kikuchi. (*J. Phys. Soc. Japan*, vol. 12, pp. 756–762; July, 1957.) Measurements of the field effect on surface conductance and on surface recombination velocity show that the difference between the Fermi level and the middle of the energy gap at the surface drifts to a negative value when the surface is kept in room air after CP4 etching. The drift is accompanied by a gradual growth of an oxide layer on the surface.

537.311.33:546.289 509
A Simpler Method for Removing Copper from Germanium—M. Kikuchi and S. Izima. (*J. Phys. Soc. Japan*, vol. 12, p. 824; July, 1957.) A simple treatment involving immersion in concentrated nitric acid followed by heating in A is described, which removes about 45 per cent of the Cu atoms originally present with a density of about 4–7×10^{-14}cm^{-3}.

537.311.33:546.289:538.63 510
Galvanomagnetic Effects in Germanium—G. C. Della Pergola and D. Sette. (*Nuovo Cim.*, vol. 5, pp. 1670–1678; June 1, 1957.) Results of measurements on *n*- and *p*-type single-crystal Ge of magnetoresistance and Hall effect (see also 3444 of 1956) as a function of magnetic induction are compared with results derived theoretically from a knowledge of the valence band structure. This comparison is wholly satisfactory only for *n*-type material.

537.311.33:546.289:548.73 511
Fine Structures of X-Ray Absorption Spectra of Crystalline and Amorphous Germanium—T. Shiraiwa, T. Ishimura, and M. Sawada. (*J. Phys. Soc. Japan*, vol. 12, pp. 788–792; July, 1957.) A comparison of the spectra suggests that amorphous Ge has a structure similar to the diamond lattice of the crystalline Ge but with a larger spacing.

537.311.33:546.48.86 512
Electrical Properties of Cd-Sb—T. Miyauchi and H. Kimura. (*J. Phys. Soc. Japan*, vol. 11, pp. 1013–1014; September, 1956.) From conductivity and Hall coefficient measurements it is found that hole mobility at room temperature is 1100 ± 200 cm^2/volts per second and the width of the forbidden energy gap is 0.3 ± 0.05 ev.

537.311.33:546.561-31 513
Linear and Quadratic Zeeman Effects and Diagmagnetism of the Exciton in Cuprous Oxide Crystals—E. F. Gross and B. P. Zakhartohenia. (*J. Phys. Radium*, vol. 18, pp. 68–71; January, 1957.) Cu_2O films 100 microns thick were studied at a temperature of 13°K in a magnetic field of 30,000 oersteds by means of a spectrograph of 1.5Å/mm dispersion.

537.311.33:546.682.19:538.63 514
The Transverse Magnetoresistance Effect in Indium Arsenide—C. H. Champness and R. P. Chasmar. (*J. Electronics Control*, vol. 3, pp. 494–499; November, 1957.) "Measurements were made on samples of various shapes. For a long thin sample the magnitude of the effect is considerably smaller than expected for acoustic lattice scattering. It is suggested that this result may be explained by the presence of some impurity scattering. In approximately square samples the magnitude of the magnetoresistive changes are large being 100 per cent or more at room temperature in a field of 10^4 G."

537.311.33:[546.682.86+546.681.86 515
Nuclear Magnetic Resonance in Semiconductors: Part 2—Quadrupole Broadening of Nuclear Magnetic Resonance Lines by Elastic Axial Deformation—R. G. Shulman, B. J. Wyluda, and P. W. Anderson. (*Phys. Rev.*, vol. 107, pp. 953–958; August, 1957.) By applying stresses to samples of InSb and GaSb, having a high degree of crystalline perfection, it has been possible to destroy the crystalline symmetry reversibly, thereby producing quadrupole broadening of the nuclear-magnetic-resonance lines. Strains of less than 10^{-4} have been detected and the resulting field gradients measured. Part 1: 1107 of 1956 (Shulman, *et al*).

537.311.33:546.682.86 516
Electrical Properties of *P*-Type Indium Antimonide—T. Fukuroi and C. Yamanouchi. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, vol. 9, pp. 262–266; August, 1957.) Report of measurements of resistance, Hall coefficient, and magnetoresistance on polycrystalline specimens in the temperature range 289°–1.4°K.

537.311.33:546.682.86 517
Absorption and Dispersion of Indium Antimonide—T. S. Moss, S. D. Smith, and T. D. F. Hawkins. (*Proc. phys. Soc.*, vol. 70, pp. 776–784; August 1, 1957.)

537.311.33:546.682.86:535.343 518
Infrared Resonant Absorption from Bound Landau Levels in InSb—W. S. Boyle and A. D. Brailsford. (*Phys. Rev.*, vol. 107, pp. 903–904; August 1, 1957.) Measurements of light transmission as a function of magnetic field on thin *n*-type samples show a field-dependent resonant absorption which is ascribed primarily to Landau levels bound to donor impurities.

537.311.33:546.682.86:538.63 519
Oscillatory Galvanomagnetic Effects in *n*-type Indium Antimonide—Y. Kanai and W. Sasaki. (*J. Phys. Soc. Japan*, vol. 11, pp. 1017–1018; September, 1956.) Measurements were made of magnetoresistance and Hall coefficient in the temperature range 1.3–300°K. At 1.3°K an oscillatory dependence on magnetic field strength was observed. See also 477 of 1956 (Kanai).

537.311.33:546.873.241 520
Electrical Properties of Bi_2Te_2—S. Shigetomi and S. Mori. (*J. Phys. Soc. Japan*, vol. 11, pp. 915–919; September, 1956.) The conductivity, Hall coefficient, and thermoelectric power have been measured in the temperature range 100–750°K. The conductivity is *p* type at room temperature. Hole mobility varies with temperature as $1.2\times10^8T^{-2.3}$, and the gap width is 0.21 ev at 0°K.

537.311.33:546.873.241:537.322.1 521
Performance of Composite Peltier Junctions of Bi_2Te_3—T. S. Shilliday. (*J. Appl. Phys.*, vol. 28, pp. 1035–1042; September, 1957.) An experimental Peltier-type refrigerator using *p*- and *n*-type Bi_2Te_3 as thermoelements of annular shape is described. Under no-load conditions a maximum temperature difference of 49°C was obtained. The experimental results agree with the theoretical predictions. See also 3224 of 1956 (Stil'bans, *et al*).

537.311.33:548.73 522
The Crystalline Perfection of some Semiconductor Single Crystals—R. L. Bell. (*J. Electronics Control*, vol. 3, pp. 487–493; November, 1957.) "Single crystals of Ge, Si, InSb, and HgTe were examined by the Guinier-Tennevin X-ray technique. The degree of perfection as shown by the width of the focused Laue image was compared with the dislocation density as revealed by etch pits."

537.311.33.096 523
Graphical Analysis of the Temperature Dependence of the Electronic Population in Semiconductors—G. Brouwer. (*Philips Res. Rep.*, vol. 12, pp. 415–422; October, 1957.) "The distribution of electrons and holes in semiconductors and its variation with temperature may be derived by means of a graphical approximation presented in this paper. The method is based directly on the electronic reaction equations in the equilibrium."

537.32:546.562-31 524
Thermoelectric Effects Shown by some Oxides. Investigation of Cupric Oxide—G. Péri, M. Perrot, and J. Robert. (*J. Phys. Radium*, vol. 18, pp. 282–283; April, 1957.) See 3111 of 1956 (Perrot, *et al*).

538.22 525
Theory of the Magnetic Properties of Ferrous and Cobaltous Oxides: Part 1 and 2—J. Kanamori. (*Progr. Theoret. Phys.*, vol. 17, pp. 177–222; February, 1957.)

538.22:538.569.4 526
High-Field Antiferromagnetic Resonance in MnF_2 using Pulsed Fields and Millimetre Wavelengths—S. Foner. (*Phys. Rev.*, vol. 107, pp. 683–685; August 1, 1957.) Measurements from 4.2°K up to the Néel temperature (68°K) are reported; they agree with the molecular-field approximation of antiferromagnetic resonance theory.

538.22:538.569.4:539.15 527
Observation of Nuclear Magnetic Resonance in Antiferromagnetic $Mn(F^{19})_2$—V. Jaccarino and R. G. Shulman. (*Phys. Rev.*, vol. 107, pp. 1196–1197; August 15, 1957.) Magnetic resonances have been observed at frequencies in the range 152–168 mc, in external magnetic fields between 300 and 3000 oersteds and at temperatures between 1.3° and 20.4°K.

538.22:546.3-1'711'681'26 528
Spontaneous Magnetization in Mn-Ga-C Alloys—H. P. Myers. (*Can. J. Phys.*, vol. 35, pp. 819–822; July, 1957.) In the alloy systems Mn-Al-C, Mn-Zn-C, Mn-Sn-C, and Mn-In-C, there is a distinctive face-centered cubic ternary phase which has pronounced spontaneous magnetization. A similar phase has been found in the Mn-Ga-C system. See also 3924 of 1957 (Brockhouse and Myers).

538.221 529
Magnetic Influence on the Recrystallized Grain Texture of a Ferromagnetic Alloy—B. Sawyer and R. Smoluchowski. (*J. Appl. Phys.*, vol. 28, pp. 1069–1070; September, 1957.) A note on the effect of a magnetic field on the texture of an Fe-Co alloy in the form of rolled sheet.

538.221 530
Temperature Dependence of Anisotropy Energy of Ferromagnetics—T. Kasuya. (*J. Phys. Soc. Japan*, vol. 11, pp. 944–947; September, 1956.) Calculations are made using the spin-wave method. The anisotropy constant is shown to depend on $(M/M_o)^{16}$ where M_o and

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M are, respectively, magnetizations at absolute zero and at the finite temperature considered.

538.221:538.24:538.652 531
Magnetization, Magnetostriction and Relaxation Phenomena during Isothermal Magnetic Annealing at High Temperatures in Ni-Co Alloys—H. Masumoto, H. Saito, and M. Takahashi. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, vol. 9, pp. 293–308; August, 1957.) The magnetization and magnetostriction as a function of time were investigated at constant temperature after the application of a magnetic field to thermally demagnetized Ni-Co alloys. In a temperature range between about 300°C and the Curie point a relaxation effect was observed which is attributed to displacements of impurities or lattice defects due to the magnetostriction at high temperatures.

538.221:621.318.1 532
The Unusual Magnetic Properties of Quenched Alcomax III—M. McCaig. (*Proc. phys. Soc.*, vol. 70, pp. 823–826; September 1, 1957.) Quenching of alcomax III from a high temperature results in an unusual soft magnetic material. The permeability is almost constant up to flux densities of 5000 G, and the ratio of resonance to saturation intensity is less than 0.1.

538.221:621.318.122 533
Processes occurring during the Heat Treatment of Alcomax—A. G. Clegg and M. McCaig. (*Proc. phys. Soc.*, vol. 70, pp. 817–822; September 1, 1957.) The dependence of saturation intensity and coercivity on temperature from room temperature to 900°C is shown. Maxima in the coercivity curves are associated with subsidiary Curie points.

538.221:621.318.124 534
Rotational Model of Flux Reversal in Square-Loop Ferrites—E. M. Gyorgy. (*J. Appl. Phys.*, vol. 28, pp. 1011–1015; September, 1957.) A discussion of the mechanism of magnetic flux reversal using fields much higher than those encountered in quasi-static processes. Detailed results for the switching coefficients are given.

538.221:621.318.134 535
The K_1- and True g-values of Polycrystalline Ferrites—J. Snieder. (*Appl. sci. Res.*, vol. B6, no. 6, pp. 471–473; 1957.) Measurements are reported which give a true g value of 2.11 and a K_1 value at room temperature of -5.8×10^3 J/m³ for Ni-Zn ferrites with 50 per cent NiO.

538.221:621.318.134 536
Magnetic Annealing Effect in Iron-Nickel Ferrites—Y. Aiyama, H. Sekizawa, and S. Iida. (*J. Phys. Soc. Japan*, vol. 12, p. 742 June, 1957.)

538.221:621.318.134 537
Structural Study of Iron Selenides $FeSe_x$: Part 1—Ordered Arrangement of Defects of Fe Atoms—A. Okazaki and K. Hirakawa. (*J. Phys. Soc. Japan*, vol. 11, pp. 930–936; September, 1956.)

538.221:621.318.134 538
Ferrimagnetic Resonance in Gadolinium Iron Garnet—B. A. Calhoun, J. Overmeyer, and W. V. Smith. (*Phys. Rev.*, vol. 107, pp. 993–994; August 15, 1957.) The apparent g factor and line width have their maximum values at the compensation point (13°C), which implies that the g factors of the Gd^{3+} and Fe^{3+} ions are equal.

538.221:621.318.134 539
Antiferromagnetism of Zn Ferrite—M. Tachiki and K. Yosida. (*Progr. Theoret. Phys.*, vol. 17, pp. 223–240; February, 1957.) Magnetic dipole-dipole spin interaction probably causes the main anisotropy energy below the Néel temperature. Above this temperature calculations are made for susceptibility and specific heat taking account of short-range order.

538.222:538.569.4 540
Paramagnetic Resonance of Nickel Fluosilicate under High Hydrostatic Pressure—W. M. Walsh, Jr. and N. Bloembergen. (*Phys. Rev.*, vol. 107, pp. 904–905; August 1, 1957.)

621.315.616.9 541
Dielectric Strength of Polyethylene at 3300 Mc/s—H. Farber and J. W. E. Griemsmann. (*J. Appl. Phys.*, vol. 28, pp. 1002–1005; September, 1957.) The dielectric strength at 3300 mc was measured to be 16.8, 13.1, and 6.0 kv/10^{-3} inches at 26°C, 57°C, and 95°C, respectively. The values are the same as those obtained for a dc field. The specially designed electrode system is described.

621.315.616.95:621.317.335.3.029.64:538.569.3 542
Temperature Dependence of Dielectric Loss of Shellac in Microwave Region—S. S. Srivastava, D. D. Puri, and P. C. Mehendru. (*Proc. nat. Inst. Sci. India*, pt. A, vol. 23, pp. 289–292; July 26, 1957.) A description of measurements on three grades of purified shellac in the 3-cm region at temperatures between 20°C and 50°C, using a standing wave technique. Tan δ increased rapidly with temperature in all cases.

621.318.2.002.2 543
Manufacture of Anisotropic Lead Ferrite Permanent Magnets Without the Use of a Magnetic Field during Moulding—F. Pawlek and K. Reichel. (*Naturwiss.*, vol. 44, p. 390; July, 1957.) Comparative demagnetization curves of oxide magnets are shown; the Pb-ferrite magnet moulded without magnetic field has higher resonance than one of Ba ferrite.

MEASUREMENTS AND TEST GEAR

529.7 544
Relation between the Proper Time of a Terrestrial Clock and Schwarzschild's Astronomical Time to an Approximation of 10^{-12}—O. Costa de Beauregard. (*J. Phys. Radium*, vol. 18, pp. 17–21; January, 1957.) An evaluation of the corrections necessary to a perfect terrestrial clock with reference to a perfect clock at rest at an infinite distance from the sun.

621.3.018.51(083.74)+529.786]:538.569.4 545
Frequency Shift in Ammonia Absorption due to Foreign Gases—K. Matsuura, Y. Sugiura, and G. M. Hatoyama. (*J. Phys. Soc. Japan*, vol. 12, p. 835; July, 1957.) See also 3949 of 1957.

621.317.3:621.314.6 546
Measurement of the Mean Rectified Value by Vacuum Diodes and Barrier-Layer Rectifiers—E. Czeija. (*Arch. tech. Messen*, no. 257, pp. 137–140; June, 1957.) The inherent errors in basic rectifier circuits and the suitability of the circuits for dc calibration are discussed.

621.317.3.029.64:621.375.9:538.569.4 547
Maser Noise Measurement—J. C. Helmer. (*Phys. Rev.*, vol. 107, pp. 902–903; August 1, 1957.) Measurements on an ammonia-beam maser amplifier at 24,000 mc confirm the theoretically low noise figure of such a device.

621.317.33:621.372.412 548
The Accuracy of a Method of Measuring the Series Capacitance and Inductance of Crystal Resonators—H. Rühl. (*Nachr. Tech. Z.*, vol. 10, pp. 297–302; June, 1957.) The accuracy obtainable with Herzog's method (*T.F.T.*, vol. 30, pp. 260–263; September, 1941. Abstract 2031 of 1942) has been investigated. A modification of this method is outlined leading to a simple formula for the series capacitance.

621.317.331:537.311.33 549
Semiconductor Specific-Resistance Meter by means of A.C. Method—A. Sato and S. Kanai. (*Rep. elect. Commun. Lab., Japan*, vol. 5, pp. 18–21; March, 1957.) Details of a transistor circuit are given.

621.317.332:621.314.7 550
Measurements of the Impedance Parameters of Junction Transistors—E. E. Ward. (*Brit. J. Appl. Phys.*, vol. 8, pp. 329–331; August, 1957.) Practical details are given of equipment for null measurements of earthed-base impedance parameters and α^1, defined as $\alpha/(1-\alpha)$. Measurements with collector open circuit are included and typical results at frequencies up to 30 kc are shown.

621.317.335.3:621.372.51.029.6 551
The Dielectric Disk used as Transformation Quadripole for the Magnification of the Node Displacement on Measuring Lines—Breitenhuber. (See 349.)

621.317.34:621.396.65 552
Measurement of Transmission Distortion in F.M. Radio Link Equipment—H. Hartbaum. (*Arch. elekt. Übertragung*, vol. 11, pp. 239–252; June, 1957.) In the method described static and dynamic nonlinearities are determined separately. Equipment operating on this principle can be used for testing the individual elements of radio links as well as the complete system.

621.317.361 553
Automatic Frequency Measurement Techniques—A. J. Green. (*Brit. Commun. Electronics*, vol. 4, pp. 342–345; June, 1957.) The equipment described has an upper counting limit of over 30 mc; a commercial-type 100-kc frequency standard and the 200-kc transmission from Droitwich are used as reference.

621.317.4.083:538.221 554
The Shape of Specimens for Measurements on Ferromagnetic Materials—P. K. Hermann. (*Arch. tech. Messen*, nos. 256–258, pp. 105–106, 129–132, and 155–158; May–July, 1957.)

621.317.7:621.397.62 555
Transistorized TV Pattern Generator—F. Rozner. (*Brit. Commun. Electronics*, vol. 4, pp. 346–349; June, 1957.) The generator described is battery operated and incorporates both p-n-p and n-p-n transistors. The mains are used only to provide frequency locking.

621.317.7:621.397.62:535.623 556
Choosing a Colour-Bar Generator—A. R. Stamiti. (*Radio & Telev. News*, vol. 58, pp. 62–63, 139; July, 1957.) Comparison of American signal generators for aligning color-television receivers. Performance features and prices of fourteen types are given in tabular form.

621.317.7.087.9:621.396.822 557
Probability Computer for Noise Measurement—A. W. Sullivan and J. D. Wells. (*Electronics*, vol. 30, pp. 208–209; October 1, 1957.) The equipment described is designed for the investigation of the amplitude distribution of atmospheric noise and its effect on radio communication systems. A constant-amplitude pulse is produced each time a reference level is exceeded by the noise input.

621.317.715 558
Sensitive D.C. Null Detector—F. Oakes and E. W. Lawson. (*Wireless World*, vol. 63, pp. 597–598; December, 1957.) Constructional

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pp. 110–120; June, 1957.) Details are given of the organization for the project and of the methods used, and some experiments and field-strength tests are described.

621.396.11.029.62/.63 586
Some Comparative Tests of Propagation Conditions in Bands II and IV—W. Knöpfel. (*Nachr. Tech. Z.*, vol. 10, pp. 233–235; May, 1957.) Comparative measurements show that attenuation in band IV due to local causes and diffraction is greater than that obtained for band II. To ensure equally good coverage a closer spacing of transmitters of higher power appears necessary.

621.396.11.029.62 587
Experimental Investigations of the Mechanism of Long-Distance Propagation in the Metre Wavelength Range—L. Klinker. (*Nachr. Tech.*, vol. 7, pp. 210–215; May, 1957.) The results of one year's observations at Kühlungsborn are summarized (see also 2586 of 1957). The fading rate does not change appreciably over distances ranging from 50 to 400 km and does not appear to depend on the mean turbulence velocity; however, a close correlation with the vertical drift component of tropospheric inhomogeneities was found. Partial reflections at layer inhomogeneities appear to be the predominant, if not exclusive, mechanism for propagation at m λ for distances up to about 500 km.

RECEPTION

621.376.332 588
F.M. Discriminator Bandwidth—G. J. Phillips. (*Wireless World*, vol. 63, pp. 571–574; December, 1957.) A significant reduction of co-channel interference by means of a wide-band discriminator is only obtained when the ratio of wanted-to-unwanted carriers is less than 6 db. For normal conditions of broadcast reception a narrow-band discriminator and an AM suppression ratio of 35 db are generally satisfactory. Other forms of interference are not improved by the use of a wide-band discriminator.

621.396.621:621.372.632 589
A Low-Noise Converter for Four Metres—G. M. C. Stone. (*R.S.G.B. Bull.*, vol. 32, pp. 541–543; June, 1957.) Full constructional details are given.

621.396.621:621.373.4 590
A Stable Oscillator for Two-Metre Receivers—W. H. Allen. (*R.S.G.B. Bull.*, vol. 32, pp. 544–545; June, 1957.)

621.396.8:621.396.712 591
The Fixing and Measurement of the Limits of Coverage of A. M. Broadcast Transmitters—P. Thiessen. (*Rundfunktech. Mitt.*, vol. 1, pp. 102–109; June, 1957.) It is proposed that the limit of the near fading zone should be determined on the basis of the statistical distribution of field strengths. The distribution of fluctuating field-strength amplitudes and phase displacements over two propagation paths is examined, and the relation between subjective assessments of quality, snr, and field-strength fluctuations is discussed. See also 592 below.

621.396.8:621.396.712.029.53 592
Statistical Investigation of the Quality of Reception of A.M. Sound Broadcasts—F. von Rautenfeld and P. Thiessen. (*Rundfunktech. Mitt.*, vol. 1, pp. 90–101; June, 1957.) Results of a series of subjective and objective tests are statistically evaluated to assess the quality of reception in West Germany of the 100-kw transmitter at Langenberg as affected by the 100-kw transmitter at Hamburg, distant 315 km and operating at the same frequency 971 kc, with or without reduced radiation towards Langenberg.

STATIONS AND COMMUNICATION SYSTEMS

621.396.1 593
Wavelengths as an International and Technical Problem—W. Nestel. (*Telefunken Ztg*, vol. 30, pp. 161–173; September, 1957. English summary, pp. 216–217.) The characteristics of the various modes of wave propagation in the entire rf spectrum are summarized and their significance in the past and future planning of international frequency allocations is discussed.

621.396.41:621.376.3:621.396.82 594
The Equalization of Base-Band Noise in Multichannel F.M. Radio Systems—C. A. Parry. (PROC. IRE, vol. 45, pp. 1527–1534; November, 1957.) Pre- and de-emphasis noise-equalizing networks with a maximum attenuation slope of 6 db per octave are discussed. Second-order distortion is assumed to be predominant. An improvement of snr of 3–6 db is obtained in the top channel when the mean power of the multichannel signal is adjusted to be unaffected by the networks.

621.396.43.029.62/.63:621.396.5 595
The Planning of Radio Link Networks for Metre and Decimetre Wavelengths—H. Paul. (*Nachr. Tech. Z.*, vol. 10, pp. 223–233.) Planning aids in the form of slide rules or similar devices are described; they are based on the relevant CCIF recommendations.

621.396.65:621.317.34 596
Measurement of Transmission Distortion in F.M. Radio-Link Equipment—Hartbaum. (See 552.)

621.396.65.029.63 597
A Decimetre-Wavelength Radio-Link Network providing High-Quality Program Channels using Pulse Phase Modulation—(*Telefunken Ztg*, vol. 30, pp. 85–118; June, 1957.) **Part 3—Modulation Equipment**—H. Oberbeck. (pp. 85–99, English summary, p. 150). **Part 4—The Radio Equipment**—E. Willwacher. (pp. 100–111, English summary, pp. 150–151). **Part 5—Equipment for Unattended Operation**—K. Hoffmann. (pp. 111–118, English summary, p. 151). Part 1: 3534 of 1956 (Brühl). Part 2: 3283 of 1956 (Schüttlöffel).

SUBSIDIARY APPARATUS

621.311.6 598
Choke or Capacitor Input?—(*Wireless World*, vol. 63, pp. 589–591; December, 1957.) Basic differences of operation and application of the two systems of power supplies are explained.

621.311.62:621.314.7 599
Design of Transistor Regulated Power Supplies—R. D. Middlebrook. (PROC. IRE, vol. 45, pp. 1502–1509; November, 1957.) Two applications of the circuit are described, one giving 0.5 a at 18–22 v and the other 1 a at 5–25 v. Output resistance is about 0.01 Ω and ripple on full load is less than 5 mv.

621.314.63:[546.28+546.289 600
Germanium and Silicon Power Rectifiers—T. H. Kinman, G. A. Carrick, R. G. Hibberd, and A. J. Blundell. (*Proc. IEE*, pt. A, vol. 104, pp. 327–330; August, 1957.) Further discussion on 2885 of 1956.

TELEVISION AND PHOTOTELEGRAPHY

621.397.26 601
Spectrum of Frequency-Shift Radio Phototransmissions—A. D. Watt. (IRE TRANS., vol. CS-4, pp. 27–40; October, 1956. Abstract, PROC. IRE, vol. 45, p. 112; January, 1957.)

621.397.61:621.396.664 602
Transistors boost Video for TV Studio Monitors—L. N. Merson. (*Electronics*, vol. 30, p. 205; October 1, 1957.) Brief description of a compact transistor preamplifier unit.

621.397.611.2 603
Typical Transmission Characteristics of the Image Orthicon Television Camera Tube—R. Theile and F. Pilz. (*Rundfunktech. Mitt.*, vol. 1, pp. 77–85; June, 1957.) The performance and limitations of image orthicons are discussed (see also 2926 of 1957). Instructions for optimum methods of operation are tabulated; some compromise is unavoidable and some faults can be reduced only by suitable lighting and arrangement of the scene.

621.397.62:525.623:621.317.7 604
Choosing a Colour-Bar Generator—Stamiti. (See 556.)

621.397.62:621.317.7 605
Transistorized TV Pattern Generator—Rozner. (See 555.)

621.397.62:621.373.52 606
A Transistorized Horizontal-Deflection System—H. C. Goodrich. (*RCA Rev.*, vol. 18, pp. 293–307; September, 1957.) A horizontal-deflection system using only transistors is described consisting of an oscillator, driver, output stage with eht supply, and phase detector. Audio-type power transistors operated beyond the manufacturer's ratings are used in the output stage, and the results indicate the desirable characteristics for an output transistor designed for this application.

621.397.62:621.375.4 607
Transistor Receiver Video Amplifiers—M. C. Kidd. (*RCA Rev.*, vol. 18, pp. 308–321; September, 1957.) General design problems are discussed and a practical two-stage amplifier is described which uses drift-type transistors.

621.397.621.2 608
Improvements in Television Receivers: Part 4—Line Synchronization with Automatic Phase Control—A. Boekhorst, B. G. Dammers, H. Heyligers, and A. G. W. Uitjens. (*Electronic Appl. Bull.*, vol. 17, no. 4, pp. 129–149; 1956/1957.) An analysis of principles and practical requirements. New types of line time-base oscillator and phase discriminator and a locking circuit are described. Parts 2 and 3: 3674 of 1957.

621.397.7 609
The B.B.C. Riverside Television Studios: some Aspects of Technical Planning and Equipment—H. C. Nickels and D. M. B. Grubb. (*B.B.C. Eng. Div. Monographs*, vol. 14, pp. 5–31; October, 1957.)

621.397.74:621.317.34 610
Waveform Testing Methods for Television Links—A. R. A. Rendall. (*Electronic Radio Eng.*, vol. 34, pp. 451–453; December, 1957.) A system of testing by means of three standard waveforms has been evolved. Acceptance limits are fixed in relation to subjective picture quality.

621.397.743 611
Television Coverage in the German Democratic Republic using a New Unified Frequency Plan—U. Kühn. (*Nachr. Tech.*, vol. 7, pp. 215–220; May, 1957.) The allocation of channels according to CCIR recommendations is discussed.

621.397.743 612
The Local Television Cable Network in West Berlin—R. Hoffmann. (*Nachr. Tech. Z.*, vol. 10, pp. 209–214; May, 1957.) The design

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and the various facilities available are described.

621.397.743:535.623 613
Transmission of Colour over Nation-Wide Television Networks—F. A. Cowan. (*J. Soc. Mot. Pict. Telev. Eng.*, vol. 66, pp. 278–283; May, 1957.) Coaxial and radio relay systems in the U.S.A. are described and their operation is discussed.

621.397.8 614
Minimizing the Effects of Ambient Light on Image Reproduction—G. L. Beers. (*J. Soc. Mot. Pict. Telev. Eng.*, vol. 66, pp. 347–354; June, 1957. Discussion, pp. 353–354.) In the method described an oscillating honeycomb screen placed in front of the picture tube prevents light from sources outside a predetermined viewing angle from degrading the image.

TRANSMISSION

621.396.61/.62:621.316.726 615
"Autosync" Frequency Control—R. J. Moser. (*QST*, vol. 41, pp. 11–16; June, 1957.) Outputs taken from the mixer and bfo of a commercial receiver are mixed and the resulting frequency is used to control an SSB transmitter, which is thus kept automatically in tune with the receiver.

621.396.61:621.396.664 616
Central Supervisory Equipment for Broadcast Transmitters—H. Müller. (*Telefunken Ztg*, vol. 30, pp. 196–200; September, 1957. English summary, p. 218.) A monitoring installation is described for supervising and modulating up to 10 medium- and short-wave transmitters from one control center.

621.396.61:621.396.82 617
Reduction of Adjacent-Channel Interference from On-Off Keyed Carriers—A. D. Watt, R. M. Coon, and V. J. Zurick. (IRE TRANS, vol. CS-4, pp. 41–58; October, 1956.)

621.396.61.005 618
High-Power Broadcast Transmitters for Medium Waves—A. Ruhrmann. (*Telefunken Ztg*, vol. 30, pp. 185–195; September, 1957. English summary, pp. 217–218.) Novel features of recently installed transmitters are described.

621.396.61.029.63:621.397.8:621.396.11.083 619
A 12-kW Pulse Transmitter for Propagation Tests in Band IV—P. Mallach. (*Rundfunktech. Mitt.*, vol. 1, pp. 86–89; June, 1957.) According to the type of directional antenna used, the transmitter described has an effective radiated power of up to 700 kw with 0.03–0.1-μsec pulse length at a repetition frequency of 15.625 kc. Various applications of the equipment and planned refinements are discussed with examples of results so far achieved.

621.396.61.07 620
Power Supply, Control and Protective Equipment (for high-power transmitters)—P. G. Zehnel. (*Telefunken Ztg*, vol. 30, pp. 201–206; September, 1957. English summary, p. 218.) Details of recent installations, including remote-control equipment, are given. See also 1588 of 1957 (Burkhardtsmaier).

621.396.712:621.398 621
The Remote and Automatic Control of Semi-attended Broadcasting Transmitters—R. T. B. Wynn and F. A. Peachey. (*Proc. IEE*, pt. B, vol. 104, pp. 529–539; November, 1957. Discussion, pp. 549–552.) Techniques developed by the B.B.C. for controlling semi-attended transmitting stations are described. The performance of the stations is analyzed and the economics of the system discussed. Considerable financial saving in operating costs is achieved without sacrifice of broadcasting continuity.

621.396.712:621.398 622
The Design of High- and Low-Power Medium-Frequency Broadcasting Transmitters for Automatic and Semi-attended Operation—W. J. Morcom and D. F. Bowers. (*Proc. IEE*, pt. B, vol. 104, pp. 540–549; November, 1957. Discussion, pp. 549–552.) Two mf transmitters are described having power outputs of 660 watts and 100 kw, respectively. Three of the low-power units are used in parallel to provide a 2-kw fully automatic transmitter. Two high-power units can be used as a semiattended 200-kw transmitter. See also 621 above.

TUBES AND THERMIONICS

537.533 623
Simplified Methods in the Statistical Mechanics of Stationary Electron Beams—B. Meltzer. (*J. Electronics Control*, vol. 3, pp. 355–366; October, 1957.) A mathematical analysis of the effects produced in an electron beam by statistical variations in the initial velocities either of thermionic or secondary emission electrons. Axial symmetry is assumed, but the statistical distribution of initial energies is not limited only to the Maxwellian type. Current density, electron temperature, and focusing are discussed, and comparisons are made with experimental results for a television camera tube.

621.314.63:537.311.33 624
On the Impact Ionization in the Space-Charge Region of *p-n* Junctions—F. W. G. Rose. (*J. Electronics Control*, vol. 3, pp. 396–400; October, 1957.) The Townsend equation for impact ionization in semiconductors is shown to be invalid at low voltages in the space charge region of *p-n* junctions. "Soft" breakdown must precede Townsend breakdown, although its effect on the characteristic will be masked by surface effects.

621.314.63:546.289 625
Thermal Turnover in Germanium *p-n* Junctions—A. W. Matz. (*Proc. IEE*, pt. B, vol. 104, pp. 555–564; November, 1957.) Static reverse characteristics are analyzed, taking into account an avalanche multiplication factor. Examination of the condition for thermal stability relates the onset of thermal runaway to a thermal turnover phenomenon. The existence of a thermal negative-resistance region is predicted and verified experimentally, a semiquantitative analysis of the results shows reasonable agreement with theory.

621.314.7 626
The Role of Transistors in Electronics—R. B. Hurley. (*J. Soc. Mot. Pict. Telev. Eng.*, vol. 66, pp. 330–332; June, 1957. Discussion, p. 332.) The advantages of transistors over tubes and other devices are summarized.

621.314.7 627
Transistor Circuits and Applications—A. G. Milnes. (*Proc. IEE*, pt. B, vol. 104, pp. 565–580; November, 1957. Discussion, pp. 581–585.) A review paper which deals with junction transistors, new concepts, and trends in design are discussed. The applications described include low-level and power amplifiers, square-wave and sinusoidal oscillators, power regulators, and computing and switching circuits, but special types such as tetrodes, *p-n-p-n*, and unipolar devices are not considered. One hundred and twelve references.

621.314.7 628
D. C. Stabilization of Junction Transistors—L. B. Johnson and P. Vermes. (*Mullard tech. Commun.*, vol. 3, pp. 67–96; April, 1957. *Electronic Appl. Bull.*, vol. 17, no. 4, pp. 151–177; 1956/1957.) The need for stabilization of dc operating conditions against temperature variations is pointed out. Detailed consideration is given to the most useful methods of achieving such stability. The effects of production spreads in transistor parameters on the design of stabilizing circuitry are discussed.

621.314.7 629
Theory of a Wide-Gap Emitter for Transistors—H. Kroemer. (PROC. IRE, vol. 45, pp. 1535–1537; November, 1957.) The injection deficit $(1-\gamma)$, where γ is the emitter efficiency, can be increased by several orders of magnitude if the emitter has a higher band gap than the base region.

621.314.7:621.317.332 630
Measurements of the Impedance Parameters of Junction Transistors—Ward. (See 550.)

621.314.7:621.317.799 631
Test Equipment for Transistor Production—A. B. Jacobsen and C. G. Tinsley. (*Electronics*, vol. 30, pp. 148–151; October 1, 1957.)

621.314.7:621.375.4:681.142 632
Transistors in Current-Analogue Computing—Kerfoot. (See 372.)

621.314.7:621.385.4 633
High-Frequency Semiconductor Spacistor Tetrodes—H. Statz, R. A. Pucel, and C. Lanza. (PROC. IRE, vol. 45, pp. 1475–1483; November, 1957.) Electrons are injected into the space-charge region of a reverse-biased *p-n* junction from an emitter contact on its surface. This input current is modulated by the voltage applied to an adjacent reverse-biased contact. Carrier transit in the space-charge region is rapid and practically unaffected by lifetime considerations. The input and output impedances of experimental units are as high as 30 mΩ. See also 1947 of 1957 (Statz and Pucel).

621.314.7:621.396.822 634
Shot Noise in Transistors—G. H Hanson and A. van der Ziel. (PROC. IRE, vol. 45, pp. 1538–1542; November, 1957.) Experimental results over a frequency range which greatly exceeds the transistor cutoff frequency in some cases are interpreted in terms of a recent theory by van der Ziel (600 of 1956). Reasonable agreement is obtained.

621.314.7.012.8 635
The Equivalent Circuit of the Drift Transistor—J. Almond and R. J. McIntyre. (*RCA Rev.*, vol. 18, pp. 361–384; September, 1957.) The quadripole admittances of a drift transistor, as given by Krömer (3389 of 1954), are examined as functions of frequency and used to derive π-type equivalent circuits for common-base and common-emitter configurations. The frequency variations of the admittance elements used in these circuits are compared with those of the admittance functions they represent and are shown to give good approximations up to frequencies of the order of the α-cutoff frequency of the intrinsic transistor. The differences in the behavior of the phase of α for diffusion- and drift-type transistors in grounded-base connection is discussed qualitatively.

621.383.2:546.34 636
Use of Lithium in Photoemissive Cathodes—A. H. Sommer. (*Rev. Sci. Instr.*, vol. 28, pp. 655–656; August, 1957.) A discussion of measurements of peak sensitivity and long-wavelength threshold and an interpretation of earlier results.

621.383.27 637
On the Mechanism of the Production of Dark-Current Pulses in Photomultipliers—Z.

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Náray and P. Varga. (*Brit. J. Appl. Phys.*, vol. 8, pp. 377–379; September, 1957.) The predominant component of the dark current is produced by different mechanisms in the negative and positive regions of the shield characteristic. In the former case electrons are emitted from the envelope, and in the latter case electrons from the inner electrodes produce scintillations on the envelope, giving rise to photon feedback.

621.383.5 638
Theoretical Interpretation of Hamaker-Beezhold Ballistic Effect in Barrier-Layer Photocells—G. Blet. (*J. Phys. Radium*, vol. 18, pp. 5–8; January, 1957.) The theory of statistical exchanges between traps and electrons is developed to explain the buildup and decay effects observed under different conditions of illumination. See 312 of 1957 (Blet and Ritti).

621.385.029.6 639
International Congress on Microwave Valves—(*Le Vide*, vol. 12, pp. 1–146; January/February, 1957.) A further selection of papers, as follows, presented at the Congress. See also 1267 of 1957. 1) **Some Technological Aspects of U.H.F. Triode Design**—E. G. Dorgelo (pp. 3–8, in French and English). 2) **Some Problems on Disk-Sealed Planar Triodes**—H. Nishio, T. Nemoto, and H. Murakami (pp. 9–22, in French and English). 3) **Construction and Circuit of a 4000-Mc/s Triode with L Cathode**—K. Rodenhuis (pp. 23–31, in French and English). 4) **The Space Charge Distribution in a Magnetron under Static Conditions**—J. Verweel (pp. 32–42, in French and English). 5) **Power and Gain Limitations of Helix-Type Travelling-Wave Tubes**—C. W. Barnes (pp. 43–48, in French and English). 6) **Helix-Type Travelling-Wave Valve for 24 Gc/s**—T. Miwa, J. Koyama, M. Mishima, and I. Yanaoka (pp. 49–52, in French and English). 7) **Low Noise Travelling-Wave Valve: ECL 1138**—M. Higuchi, K. Kawazura and J. Koyama (pp. 53–58, in French and English). 8) **Magnetless Magnetron**—A. Versnel (pp. 59–63, in French and English). 9) **A New Electron Lens System providing Convergent-Divergent Action**—T. Seki, Y. Nikaido, and K. Simada (pp. 64–73, in French and English). 10) **The Design of an Electron Gun for a Strip Beam Device**—A. B. Cutting and J. Fraser (pp. 74–82, in French and English). 11) **Transmit-Receive Switch Tubes**—M. L. W. Roberts (pp. 83–92, in French and English). 12) **Plug-In T. R. Tubes for use in S-Band Duplexers**—T. L. Dutt (pp. 93–108, in French and English). 13) **High-Power Duplexers**—L. Milosevic (pp. 109–116). 14) **Investigation of Pressures and their Evolution in Gas-Filled Valves**—J. Boissière and C. Romiguière (pp. 117–121). 15) **Power Measurements on Millimetre-Wave Valves**—H. Weill (pp. 122–127). 16) **Electron Deflection System of Travelling-Wave Cathode-Ray Tube**—H. Maeda and K. Miyaji (pp. 128–140, in French and English) 17) **High-Melting-Point Seals**—R. Benichou (pp. 141–146).

621.385.029.6 640
Biperiodic Electrostatic Focusing for High-Density Electron Beams—K. K. N. Chang. (Proc. IRE, vol. 45, pp. 1522–1527; November, 1957.) The potential valley formed by combining two counteracting periodic electrostatic fields is theoretically steeper than in previous focusing systems and should maintain stable flow in thick hollow beams. Proper cancellation of the space-charge field is also obtained. Preliminary experiments indicate that a travelling-wave tube can operate without using any magnets. See also 1954 of 1957.

621.385.029.6:621.3.072.6 641
Automatic Frequency Control for a Pulsed Klystron—(*Elec. Commun.*, vol. 34, pp. 136–140; June, 1957.) The klystron frequency is compared with that of a stable microwave resonator. The frequency-error information derived from one pulse is delayed and applied to correct the next.

621.385.029.6:621.372.8 642
Delay-Line Valves—F. W. Gundlach. (*Nachr. Tech. Z.*, vol. 10, pp. 265–276; June, 1957.) A theory is derived for tubes containing periodic waveguide structures. It is similar to recurrent-network theory and accounts for the properties of the tube in both forward and backward operation.

621.385.029.64 643
Slalom Focusing—J. S. Cook, R. Kompfner and W. H. Yocom. (Proc. IRE, vol. 45, pp. 1517–1522; November, 1957.) A linear array of line charges produces two equipotential surfaces which contain exact electron trajectories. Such a field is obtained by placing an array of positive wires between two negative plates; this structure can focus a ribbon-type electron beam. A backward-wave oscillator with this "slalom" focusing generated oscillations in the range 3.3–4.3 kmc.

621.385.029.65 644
A Versatile Source of Millimetre Waves—C. F. Hempstead and A. R. Strnad. (*Bell Lab. Rec.*, vol. 35, pp. 241–245; July, 1957.) A number of simplifications in design and improvements in fabrication of backward-wave oscillators are described, including a redesigned electron gun assembly. The latest versions of the backward-wave oscillator deliver 5 to 10 mw at between 45,000 and 57,000 mc; oscillation has been achieved at 200,000 mc.

621.385.032.213:537.533 645
On the Electron Gun with Hairpin Cathode: the Intensity of Emission and the Nature of the Electron Source—K. T. Dolder and O. Klemperer. (*J. Electronics Control*, vol. 3, pp. 439–454; November, 1957.) Emission per unit solid angle is maximum for zonal rays when grid bias is just sufficient to prevent multiple source formation and cathode-to-grid distance is of the order of one grid-aperture radius. Experiments show that emission can be direct from the cathode tip, or pass through a crossover, or be drawn from a virtual cathode. The last condition yields the highest emission over a relatively large solid angle.

621.3.032.213.13 646
Barium-Supplied Oxide-Coated Cathodes—T. Imai. (*J. Phys. Soc. Japan*, vol. 12, p. 831; July, 1957.) The beneficial effects on emission of excess Ba in the coating layer of oxide-coated cathodes are described and methods of introducing the Ba are discussed.

621.385.032.213.13 647
Activation of an Oxide Cathode by Deposition of Alkaline Earth Metal Ions via a Mass Spectrometer—R. M. Matheson, L. S. Nergaard, and R. H. Plumlee. (*RCA Rev.*, vol. 18, pp. 385–431; September, 1957.)

621.385.032.213.13:621.317.332.029.63 648
The Impedance of the Oxide-Coated Cathode at Ultra-high Frequencies—L. J. Herbst. (*Brit. J. Appl. Phys.*, vol. 8, pp. 277–282; July, 1957.) The impedance of the cathode was measured on a number of disk-seal triodes at frequencies between 500 and 2365 mc.

621.385.032.213.13:621.327.5 649
The Emission from Oxide Cathodes in Low-Pressure Discharges—M. A. Cayless. (*Brit. J. Appl. Phys.*, vol. 8, pp. 331–336; August, 1957.) Cathodes in fluorescent lamps are considered. The zero-field emission at a pressure of a few mm is found experimentally to be about 10 times that in a vacuum. Various processes of field and secondary emission are described.

621.385.032.213.2 650
The Effect of Oxygen and Sulphur on the Thermionic Emission from Matrix Cathodes—J. F. Richardson. (*Brit. J. Appl. Phys.*, vol. 8, pp. 361–362; September, 1957.) "The poisoning of the emission from matrix cathodes by oxygen and sulphur is reversible. Recovery of the emission is more rapid from oxygen poisoning. The shape of the recovery curves indicates that the poisoning agent is present on the surface and has also diffused into the cathode pores."

621.385.032.26:537.533 651
Exact Electrode Systems for the Formation of a Curved Space-Charge Beam—R. J. Lomax. (*J. Electronics Control*, vol. 3, pp. 367–374; October, 1957.) An analytical method of determining the shape of the Pierce-type electrodes required to maintain a two-dimensional space-charge beam is given, which is valid when the edge of the beam is curved. The method is applied to a solution of the space-charge equations given by Meltzer (957 of 1957); this leads to a completely enclosed electrode system.

621.385.1.01 652
Recommendations of the Special Committee on Valves of the N.T.G. (Nachrichtentechnische Gesellschaft) concerning Definitions in the Field of High-Vacuum Diodes and Grid-Controlled Vacuum Tubes—(*Nachr. Tech. Z.*, vol. 10, pp. 257–262; May, 1957.) Many of the German terms and definitions given are based on those recommended by the IRE Committee on Electron Tubes.

621.385.2:621.3.011.4 653
The Capacitance between Diode Electrodes in the Presence of Space Charges—I. A. Harris and C. S. Bull. (*Proc. IEE*, pt. B, vol. 104, p. 598; November, 1957.) Comment on 3357 of 1957 and author's reply.

621.385.3.029.63 654
The Design and Life of Planar Microwave Transmitting Tubes—H. D. Doolittle. (IRE Trans., vol. VC-8, pp. 31–35; May, 1957. Abstract, Proc. IRE, vol. 45, p. 1039; July, 1957.)

621.387:621.3.015.532 655
The Corona Triode—G. A. Ostroumov. (*Dokl. Ak. Nauk S.S.S.R.*, vol. 115, pp. 919–921; August 11, 1957.) The principle and circuit of an amplifying device is described which consists of a sharp point and a blunt electrode as anode with a high voltage between them so that a corona is formed. For operation at high voltages the anode is of a tubular shape. See also 2927 of 1956 (Hertz).

MISCELLANEOUS

621.3.002.2/.3 656
Miniaturization of Electronic Equipment—D. A. Findlay. (*Electronics*, vol. 30, pp. 177–204; October 1, 1957.) A report dealing with design and materials, various classes of components, and production techniques.

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