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PROCEEDINGS OF THE IRE®

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CONTENTS

ADVERTISING SECTION

THE COVER—The pattern on the cover is a design suggested by the Editor's page, "Poles and Zeros." It portrays in three dimensions the amplitude and phase responses of a network as viewed on the complex frequency plane. The valleys ("zeros") and peaks ("poles") represent points of resonance and antiresonance, respectively, of a two-terminal impedance. The contour lines encircling the peaks at constant heights mark off various values of amplitude response, whereas the lines running down the slopes correspond to different values of phase response. By manipulating the positions of the poles and zeros in such a display, a network designer can visualize directly the relative contributions of the *R*, *L* and *C* elements to the over-all responses of the network, a visualization virtually impossible by observation of the conventional network equations.

Scanning the Issue

Nikola Tesla, 1856–1943 (Pratt, p. 1106)—Nikola Tesla was one of the most brilliant inventors of his time. His contributions to electrical engineering and to the then infant field of radio were tremendous. He is best known for his invention of the rotating electric field in ac machinery. This one invention was responsible for the world-wide adoption and development of ac polyphase systems, as opposed to dc systems, for the transmission of electric power. Actually, most of his life was spent in experimenting with high-frequency phenomena and his achievements in this area were so far ahead of his time that the significance of his work was not fully appreciated until many years later. It is for this reason that even today his many accomplishments are not as widely known as those of his contemporaries. This year the one hundredth anniversary of his birth is being commemorated in the United States and in Yugoslavia, where he was born. The IRE is participating in this event by publishing in this issue an article prepared by the Chairman of the IRE History Committee which summarizes and appraises the remarkable career of Nikola Tesla.

A New Beam-Indexing Color Television Display System (Clapp, *et al.*, p. 1108)—This is the first of three companion papers in this issue describing a new and promising method of displaying color television pictures. The display system features a single-gun picture tube, the face of which is coated with vertical stripes of red, green and blue phosphors. Instead of forcing the picture beam to land on a particular phosphor as is done in other tubes, the beam is permitted to transverse the phosphor stripes unhindered. This eliminates the need for beam-directing shadow masks and beam-deflecting grilles common to other three- and one-gun tubes, greatly simplifying the mechanical structure of the tube. There remains the problem of modulating the beam in accordance with its position in order to produce the required color. How this is accomplished is the outstanding feature of the system. A stripe of material having a high secondary emission is placed behind every red phosphor stripe. The passage of a beam across these index stripes produces a pulsating secondary emission current, or index signal, which is indicative of the position of the beam with respect to the phosphor structure. For reasons explained in the paper, modulation of the scanning beam by the picture signal corrupts the resulting index signal and it is therefore convenient to introduce a second beam, called a pilot beam, whose function it is to track the picture beam and to produce a separate, usable index signal. The positional information thus derived is then combined with the color signal in such a fashion as to properly correlate the position and intensity of the picture beam at all times. The authors go on to compare this system in detail with other systems and show that it is considerably less complex, and give their opinion that it is potentially more economical than any other color receiver. If experience bears out this prediction, this may well be the color television tube of the future.

A Beam-Indexing Color Picture Tube—The Apple Tube (Barnett, *et al.*, p. 1115)—Having established the over-all principles of the beam-indexing system in the preceding paper, we now proceed to a description of the specific design, construction and operation of the color tube itself. The discussion covers, among other things, the number, width and positioning of the phosphor stripes, the design of an electron gun that can produce two beams which will track one another, spot size and intensity of the beams, and the index structure. The result is a 21-inch rectangular color picture tube which appears capable of producing high-quality pictures, both in monochrome and in color, and is believed to be potentially less costly to manufacture than other types.

Current Status of Apple Receiver Circuits and Components (Bloomsburgh, *et al.*, p. 1120)—To round out this trio of papers on the Apple system, a description is given of the construction, circuitry and operation of a developmental color television receiver utilizing the beam-index type of display. This concludes a report on one of the most interesting developments, both technically and commercially, in the young history of color television.

Directions of Improvement in NTSC Color Television Systems (Richman, p. 1125)—In this paper the author presents some ideas on how certain details of the NTSC color television standards might be modified so as to allow the use of simpler color receivers with better resolution and also to improve the quality of pictures displayed on monochrome receivers. The discussion centers about some minor changes in the transmitted signal which are designed primarily to increase the amount of visible information carried by the single-sideband portion of the chrominance channel, with more careful regard to luminance. In so doing, the author presents much food for thought at a timely moment in the development of color television.

A Precise New System of FM Radar (Ismail, p. 1140)—FM radar systems operate on the general principle of continuously transmitting a frequency-modulated signal and as the target echo is received, comparing its frequency with the frequency of the signal then being transmitted, which by that time has swept on to a different value. Thus by mixing the two signals a difference frequency is produced, the magnitude of which is related to the range of the target. Actually, the extraction of range information by this method has its complications and is perhaps less straightforward than the conventional pulse method. However, as the range becomes shorter and shorter the difference frequency remains more discernible than does the shortening time interval between two pulses. This short-range capability has led to the wide-spread and exclusive use of fm radar for low-altitude altimeters, providing a vital aid in the landing of aircraft. Various fm radar systems differ from one another primarily in the shape of the modulating wave that is employed, which in turn determines the types and magnitudes of errors that occur. The author of this paper proposes using a type of modulation which together with a clever method of processing the signal eliminates a fixed error that is characteristic of other types and results in a highly precise method of measuring not only the distance but also the speed of a target up to extremely short ranges. The system appears capable of indicating ranges as short as two feet and an experimental altimeter built by the author proved sensitive to altitude changes as small as three feet.

Maximum Angular Accuracy of a Pulsed Search Radar (Swerling, p. 1146)—The accuracy with which a pulsed search radar can give the direction of a target is often considered as roughly equal to the width of the beam. Actually, though, the rounded gain pattern of the beam causes a variation in the strength of the echo as the beam sweeps the target, which makes it possible to estimate the angular position with an error considerably less than the beamwidth. In this paper the author, using a well-known theorem of statistics, succeeds in deriving the theoretical maximum accuracy that can be achieved and examines the effect of receiver noise on this accuracy. The results of the analysis are presented in useful graphical form and then applied to a typical search radar. The method of analysis used here will probably find application in the study of several related problems in the future. In the meantime it has provided some very worthwhile information on an important performance characteristic of search radars.

An 8-mm Klystron Power Oscillator (Bell and Hillier, p.

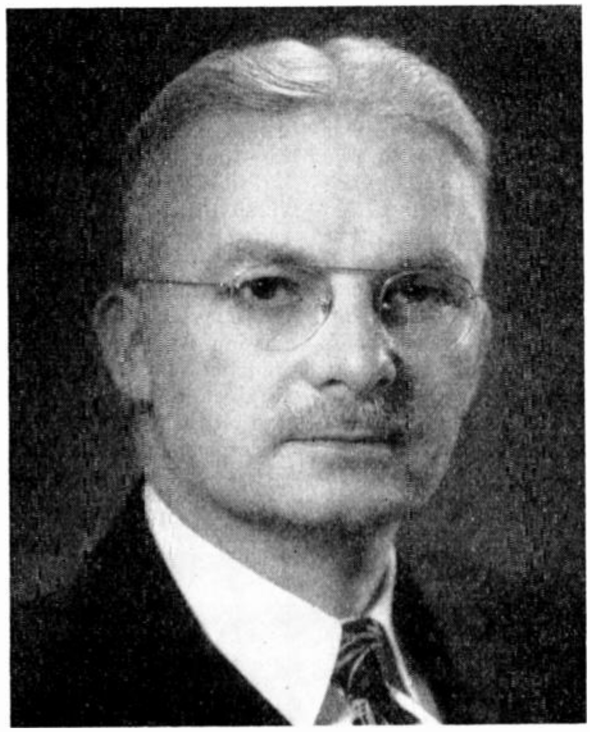
1155)—For the first time the power output of a millimeter-wave klystron has been raised above the milliwatt level. The design and performance of a tube is described which can produce a cw output of 12 watts at 34,000 megacycles, far surpassing anything previously reported. The paper includes a number of novel features and techniques of tube construction which are material to obtaining such unusual performance near the outer frontier of the radio-frequency spectrum.

Restrictions on the Shape Factors of the Step Response of Positive Real System Functions (Zemanian, p. 1160)—This paper is a continuation of a study, made by the author in two earlier PROCEEDINGS papers, of the transient responses of various types of networks and of the bounds within which certain important response characteristics must lie. Among other things, the earlier work defined the minimum rise time which a passive network with a shunting capacity across its terminals was capable of. Unfortunately, this quickest possible response carries with it the penalty of an infinite overshoot. In many situations it is desired that the output of the network be a replica of the shape of the input, as nearly as possible. This means placing a limitation on the amount of overshoot that can be tolerated, even though it is done at the expense of fast rise time. The present paper explores how much the minimum rise time must be extended in the case of a step input in order that the overshoot (or undershoot) does not exceed some specified value, yielding results of both theoretical and practical importance to circuit designers.

IRE Standards on Electronic Computers: Definitions of Terms (p. 1166)—Computer engineers will welcome this much-needed list of official definitions of some 175 terms which they use extensively in their everyday work. Reprints of this document may be purchased from IRE as noted on the first page.

***P-N-P-N* Transistor Switches** (Moll, *et al.*, p. 1174)—Negative resistance is in itself a very useful and widely used characteristic. Devices which exhibit negative resistance also have another useful property; namely, they have two stable dc steady states of operation. This bistable property suggests that a device of this sort might be used for such things as a switch or a waveform generator. Transistors can be made to give negative resistance characteristics by operating them in the alpha-greater-than-one region. This paper explores the design and operation of *p-n-p-n* transistors in this region and comes forth with a new type of semiconductor device which can be used, among other things, as a switch, photorelay or a saw-tooth generator.

Two-Terminal *P-N* Junction Devices for Frequency Conversion and Computation (Uhlir, p. 1183)—The process of shifting the frequency of a signal from one part of the radio spectrum to another is of the greatest practical importance in communications systems. A low-frequency signal is usually converted to a high frequency for convenience in transmission. At the receiver end, the signal is converted to a lower frequency for ease of amplification and handling. The analysis given in this paper of how this conversion operation is carried out in junction diodes leads to the formulation of useful principles for designing semiconductor converters. While it is true that semiconductor diodes are widely used for converting down to a low-frequency, such is not the case for "up-converting," and it is here that the author lays the emphasis of his work. He finds that amplification of signal power is possible with "up-converting" diodes, that they have relatively good power-handling capabilities, and that they can also be used as pulse amplifiers. These findings should stimulate the use of semiconductor diodes in an important and rather neglected area. Moreover, the analysis makes a long overdue contribution to the theory of diode converters by bringing up to date in terms of modern *p-n* junction theory a phenomenon which was last explained at a time when the concept of minority carriers was unknown.



Charles R. Burrows

DIRECTOR, 1956–1957

Charles R. Burrows was born in Detroit, Michigan on June 21, 1902. He received the bachelor's degree in 1924 and the electrical engineering degree in 1935 from the University of Michigan, the A.M. degree in 1927 and the Ph.D. degree in physics in 1938 from Columbia University.

From 1924 to 1945 Dr. Burrows was associated with the radio research department of the Bell Telephone Laboratories where he worked on radio wave, transoceanic short-wave, and ultra-short-wave propagation. He worked on the development of field-strength measuring sets, short-wave and ultra-short-wave transmitters for multiplex operation using negative feedback, the proximity fuse, radar and countermeasures equipment.

From 1945 to the middle of this year Dr. Burrows was Director of the School of Electrical Engineering at Cornell University. At the same time he was also Associate Chief Scientist at the Advanced Electronics Center of the Electronics Division, General Electric Company.

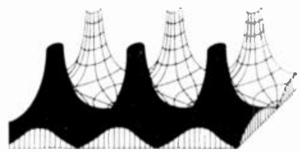
On July 1 of this year, Dr. Burrows joined the Ford Instrument Company, a division of the Sperry Rand Corporation, as Vice-President for Engineering. He is in charge of all their engineering, development and research activities.

From 1943 to 1945 he was Chairman of the Committee on Propagation of the National Defense Research Council. He was President of the Joint Commission on Radio Meteorology of the International Council of Scientific Unions from 1946 to 1954. From 1948 to 1954 he was International President of Commission II on Tropospheric Propagation of the International Scientific Radio Union. He was Vice-Chairman, and later, Chairman of the U. S. National Committee of U.R.S.I. He headed the American delegation to the U.R.S.I. General Assembly in 1952. At the present time, he is Vice-President of U.R.S.I.

Dr. Burrows is the holder of several patents and the author of many scientific and engineering papers. He received the Presidential Certificate of Merit for his wartime services. He is a Fellow of the A.I.E.E. and the American Physical Society. He is a member of the American Astronomical Society, American Geophysical Union, American Society for Engineering Education, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

Dr. Burrows joined the IRE as an Associate Member in 1924 and became a Member in 1938. In 1943 he became a Senior Member and, then, a Fellow. He is presently a member of the Appointments and Wave Propagation Committees.

Poles and Zeros



Backlog. A common misconception is that it takes forever and a day to get a paper published in the PROCEEDINGS OF THE IRE. To gather evidence to the contrary, we recently scanned the last twelve issues (excepting special issues), comparing the date of original submission of each paper with the date of publication. Excluding a few papers which authors held up a long time to revise, the results were: fastest publication, three months; slowest eight months; average, five months. Three years ago when the Editorial Board began its campaign against publication delays, the average was ten months

The campaign is still on, and we hope to clip another few weeks off the average, but we're close to the irreducible minimum right now. As further evidence, the July issue of the PROCEEDINGS had sixteen fewer editorial pages than normal because the backlog of accepted papers was close to zero—and you can't do much better than that. Summer is traditionally the low point in editorial logistics; we'll make up the sixteen pages before Christmas.

Keeping the backlog down takes coordination of many workers. Three reviewers must read each paper thoroughly and make their recommendations. This process can seldom be completed in less than three weeks and weeks of additional delay occur if only one copy of the manuscript is available. Hence the requirement for submission in triplicate. Staff action on the reviewer's reports, plus marking the manuscript for the printer, takes nearly another two weeks. Typesetting takes from two weeks to a month, depending on the printer's schedule and the length and technical difficulty of the paper. Our printer, Banta of Menasha, Wisconsin, is one of the most experienced technical compositors in the business, so any time lost in the mail between Menasha and New York is well worth it—and not serious in any event.

The author must then read galley proofs. This takes from a week to a month or more, depending on the author, his disposition and circumstances. Then come page make-up, reading page proofs, printing, binding and mailing. The press run for 57,000 copies (our print order last month) takes several days in itself, to say nothing of binding and mailing.

All in all, publishing a paper in three months means *no* lost time anywhere in the schedule. It also means no revision was required by the author, a process which can take anywhere from two weeks to two months. Maintaining a five-month average means that the staff can hold a "ready" paper in the backlog no more than two months at the most, and this is done only to permit assembling issues of balanced content.

With the continued devotion of authors, reviewers, referees and staff, it is our aim to keep the backlog down to no more than one issue's worth of unpublished material on hand. To do so we must have a reasonably steady flow of material, and we must be willing to adjust the number of pages in each issue to suit the circumstances. So doing, it is our hope that no author will decide against submitting a paper to the IRE because "it takes so long to get it in print."

History. The art and science of electronics being barely fifty years of age, it is no surprise that most of those who have contributed to its progress are still with us, hale and hearty. But time has marched on, and the second generation has come up tens of thousands of recruits in their twenties or early thirties who have already taken over the creative reins and who must soon assume the posts of the retiring elder statesmen. These younger men, unless they have been fortunate enough to study the recent history of science, have but a vague acquaintance with the great names of early radio and the allied arts. The brilliant concept of tuning a resonant circuit, for example, seems to them a simple fact of nature, clearly ordained in the order of things. That this idea had to be wrested from the unknown with the greatest of effort, that it was suggested by a visionary genius years before the experts achieved its reduction to practice, seems hardly credible. This story, and many like it, can be a source of real inspiration to the new lieutenants, and it is well that the stories be told by those who lived through them, while they still live to tell them.

So the Board of Directors has set up a Committee on History, headed by Haraden Pratt, whose task it is to refresh our memory and to remind us all that, whatever the difficulties of present day invention and development, the agony is not new. The latest output of this Committee appears on page 1106 of this issue. In it, Chairman Pratt tells the amazing story of Nikola Tesla, a genius the like of which we could well do with today, on the 100th anniversary of his birth.—D.G.F.

Nikola Tesla
1856-1943

HARADEN PRATT†, FELLOW, IRE

The following account was prepared by the Chairman of the IRE History Committee, at the request of the Executive Committee, as a tribute to one of the most brilliant experimenters in electrical and radio phenomena on the occasion of the 100th anniversary of his birth. —*The Editor*

THIS YEAR 1956 marks the Centennial of the birth of Nikola Tesla, acclaimed by some as the world's greatest electrician of his day. He was one of those particularly gifted with a generous number of talents. These included an encyclopaedic grasp of many fields of learning, an easy familiarity with several languages, thoroughness in the treatment of technical matters, the ability to lecture with tutorial facility, the writing of poems in more than a single tongue and the vision to create concepts in the applied sciences so far beyond the times that many years elapsed before instrumentalities became available so that they could be put to use.

Tesla's mind strode boldly ahead of current thinking. It was unfettered by the restrictions that blocked the efforts of others, and would blithely leap over these restrictions starting afresh with new ideas aimed to achieve a visionary end result on the road to which he would matter-of-factly indicate how important hurdles were to be conquered, whether or not the means for doing so were available.

Our existing industrial era would cease to function without Tesla's first and greatest contributions. His entry as a young man into the electrical era came when the future of electric power was in chaos and engineers were speculating as to whether the use of direct current should continue or alternating current should be made the standard. Direct current could be transmitted only a very short distance, but was suitable for running motors; whereas alternating current could be transmitted afar, but efficient motors for using it were nonexistent. His brilliant invention of the rotating electric field, making possible the very simple commutatorless, nonsynchronous, polyphase induction motor which required alternating current but eliminated the troublesome and costly brushes and commutators necessary for direct current use, cut the Gordian knot and established a need for the universal availability of polyphase power. The Tesla system was adopted for the first power plant at Niagara Falls, completed in 1895, which ushered in the era of polyphase power transmission. Today our vast power network is receiving energy from this original plant together with energy from the recently completed first atomic-energy generating facility.

An early important application of electricity was for street lighting, and Tesla devoted himself to improving arc lighting which, with normal alternating current power, produced an objectionable hum. He accordingly became interested in high frequency electric currents and, for the rest of his life, devoted himself primarily to experimenting with them. Hertz in 1887 first demonstrated the existence of electromagnetic waves, using for the source of high frequency power the relatively feeble oscillatory spark discharges of condensers. Tesla's lofty aspirations required high frequency power on a vast scale and at very high voltages. In 1891 he produced a rotating alternator having 384 poles and an output frequency of 10,000 cycles per second, followed by other machines developing up to 25,000 cycles per second. Tesla believed that undamped current generation was very important, but he was a quarter of a century ahead of the times, as this method did not come into practical use until after 1910. His early models used the inductor principle with stationary coils, which was the arrangement used for the huge commercial machines subsequently built for radio communication.

Rotating machines could not provide the high frequencies and voltages Tesla wanted, so he utilized principles that, he stated in his famous lectures, were well-known to electricians, such as tuned circuits, induction coils, and oscillatory spark circuits, which he combined with an oscillation transformer to create the spectacular luminous flaming arc discharge effects that brought his name to wide fame. He explained and demonstrated this apparatus and high frequency phenomena in a series of brilliant lectures, the first in 1891 before the American Institute of Electrical Engineers in New York, the second in 1892 before the Institution of Electrical Engineers and the Royal Institution in London, the Société Internationale Française des Electriciens and the Société Française de Physique in Paris, and the third in 1893 before the Franklin Institute in Phila-

† *Chairman*, IRE History Committee.

delphia and the National Electric Light Association in St. Louis. In these lectures Tesla's galloping mind traversed the gamut of scientific thinking including speculations on the electrical nature of the structure of matter. In addition he had an extensive exhibit at the Chicago World's Fair in 1893. It was only a matter of a couple of years when practically every technical school of note throughout the world had a Tesla Coil apparatus. The impact of these teachings on high frequency research by both scholars and students was enormous.

Tesla's revelations and accomplishments encompassed a wide variety of subjects. Early in 1890 he described how metal and dielectrics could be heated in the fields of specially designed high frequency coils, forecasting the commercial high frequency furnace and the present day dielectric heating industry. Experimenting on himself, he demonstrated the effect of high frequency currents on the human body, a technique now called diathermy. Prior to 1893, starting where Sir William Crookes left off, he further developed vacuum and gas-filled tubes adopting special types of glass and coating them with phosphors. He bent these luminous tubes to fit a room or to form words. Long after, neon signs appeared, and still longer after, fluorescent lamps were introduced. Tesla once commented that the way to efficiently conduct high frequency currents was by using a cable made up of many insulated small wires, but he observed such material was not available. In later years this appeared under the name of Litzendraht. One of the novelties he predicted was the use of cheap, synchronous electric clocks on a world-wide basis, and he demonstrated such clocks at the World's Fair in 1893. He said that the future of aviation, then nonexistent except for balloons, depended on the development and use of aluminum. It must be remembered that aluminum then was scarce and expensive to make and only became a practical material with the advent of the cheap and plentiful electric power needed for the reduction of its ore. Thus the aviation industry of today owes a substantial heritage to this unconventional mastermind. Tesla's disclosures in his early lectures, and patents on the generation of high frequency currents, described devices which became important many years later. For example, rotary spark gaps and series spark gaps with small spacings became basic elements in the wireless telegraph systems of the 1909 to 1920 period.

Tesla's versatility in creating uses for high frequency currents caused him, in his February, 1892 lecture, to forecast the possibility of transmitting power through space without the use of conducting wires. In that same month Sir William Crookes published a prediction that electromagnetic waves in space would be used for telegraphically communicating through space without the use of wires. Tesla's concept was to disturb the electrostatic condition of the earth, setting up standing waves on its whole surface by exciting it with high frequency power and then taking off power anywhere that wave amplitude was present. While Crookes did not discuss or pursue methods, Tesla, commencing with his 1893 lecture, described elevated antennas connected to the earth with wires for both transmission and reception, and pointed out the importance of applying the principle of electrical resonance to these arrangements, even making their tuning variable.

In 1896 he conducted experiments resulting in the transmission of signals some 25 miles to a Hudson River boat. To make continuous waves in a receiving system audible, he suggested the use of vibrating contacts, which years later became the accepted practice until the introduction of the heterodyne beat note method. In 1898 he demonstrated and patented a radio controlled vessel, the forerunner of the present-day guided missile. For this purpose he stated that "any waves, impulses, or radiations which are received through the earth, water, or atmosphere could be used" and that "vessels or vehicles of any suitable kind may be used as life, despatch or pilot boats, or the like, or for carrying letters, packages, provisions, instruments, objects, or materials of any description, for establishing communication with inaccessible regions and exploring the conditions existing in the same, for killing or capturing whales or other animals of the sea, and for many other scientific, engineering, or commercial purposes; but the greatest value of my invention will result from its effect upon warfare and armaments, for by reason of its certain and unlimited destructiveness it will tend to bring about and maintain permanent peace among nations." At the turn of the century Tesla talked about radio broadcasting saying: "I have no doubt that it will prove very efficient in enlightening the masses, particularly in still uncivilized countries and less accessible regions, and that it will add materially to general safety, comfort and convenience, and maintenance of peaceful relations. It involves the employment of a number of plants, all of which are capable of transmitting individualized signals to the uttermost confines of the earth. Each of them will be preferably located near some important center of civilization and the news it receives through any channel will be flashed to all points of the globe. A cheap and simple device, which might be carried in one's pocket, may then be set up somewhere on sea or land, and it will record the world's news or such special messages as may be intended for it." In 1917 he forecast radar by indicating the possibility of shooting out a pulsed concentrated ray of very high power vibrating at the tremendous frequency of millions of times per second and then intercepting it after being reflected from a hidden object and displaying this reflected ray on a fluorescent screen. The means for accomplishing these several concepts were not developed until twenty years or more later.

Tesla's idealized dream of causing the whole terrestrial globe to oscillate electrically engaged a great deal of his attention. He made extensive experiments at Colorado Springs in 1899 where he produced artificial lightning crashes 135 feet long. Later he constructed a 200-foot high tower on Long Island surmounted by a 70-foot metal sphere which was to be excited by millions of high frequency volts for broadcasting telegraphy, speech, vision, and power, but it was never completed. Tesla's indefatigable strivings to implement his apparently unclear and visionary concept did not succeed. He neglected to leave behind any clear record of conclusions from the Colorado experiments and other subsequent work. Unfortunately he was severely handicapped in later years because of lack of funds, not only for experimenting, but also for personal living. A friend once wrote of Tesla that the goddesses of Fame and Fortune are capricious, one of them having smiled on him but not the other.

Tesla characteristically seemed indifferent toward the commercial application of his ideas, preferring to follow the lure of new challenges. Since his basic objective after about 1893 was directed towards producing a world-wide series of grandiose electrical effects, the many ideas and items of apparatus which he produced were left for others to pick up and embody for less ambitious but more practical purposes. For this reason Tesla's influence on the development of radio was known to but a limited number of people. A few eminent persons who attended or read his lectures during the 1890 decade were inspired by his revelations and some others, who later delved into the backgrounds of the art, became aware of the pioneering import of his contributions.

Far ahead of his time, mistaken as a dreamer by his contemporaries, Tesla stands out as not only a great inventor but, particularly in the field of radio, as the great teacher. His early uncanny insight into alternating current phenomena enabled him, perhaps more than any other, to create by his widespread lectures and demonstrations an intelligent understanding of them, and inspired others not yet acquainted with this almost unknown field of learning, exciting their interest in making improvements and practical applications. Many developments generally attributed to others had their genesis in the trail-blazing teachings of this pioneer genius.

A New Beam-Indexing Color Television Display System*

R. G. CLAPP†, SENIOR MEMBER, IRE, E. M. CREAMER†, ASSOCIATE MEMBER, IRE, S. W. MOULTON†, M. E. PARTIN†, ASSOCIATE MEMBER, IRE, AND J. S. BRYAN†, ASSOCIATE MEMBER†, IRE

***Summary*—This paper describes a single-gun cathode-ray display system (the Apple System) for color television receivers based on the phenomenon of secondary emission. An index signal, derived from a secondary emissive structure built into the screen of the tube, continuously indicates the position of the scanning spot relative to the color phosphor structure. This positional information is combined with the color television signal, and the combined signal modulates the scanning spot in amplitude and phase in such a manner that the spot sequentially illuminates the primary colors in the appropriate amounts and proportions to reproduce the intended scene. This paper describes the general features of the system and the philosophy behind its development, and the derivation of the index signal and its utilization in the color-processing and grid-drive circuits.**

Introduction

From its inception many years ago, the aim of the Philco color television development program has been to produce a color television display in which the picture tube and its external beam-controlling parts are as simple as possible. Most other color display systems are based either on the premise that each of several color phosphors is excited by its own electron beam while being protected from the other beams by mechanical or electromechanical means, or, alternately, that a single electron beam is directed to several color phosphors by electromechanical means. These types of display require a mechanical structure within the tube and present problems in registration or focus, or both.

The beam-indexing system is based on the premise that a single electron beam can be used to excite the several color phosphors without auxiliary color deflection or beam shadowing. Instead of forcing the beam to land on a particular phosphor, the beam can be passed over all color phosphors in rapid succession and modulated in accordance with its position to produce the required color. Operation in accordance with this principle requires an indexing system to provide information concerning the whereabouts of the writing beam and a modulating system to provide the required beam modulation. The beam-indexing display system avoids the mechanical and registrational problems of other color

* Original manuscript received by the IRE, June 4, 1956; revised manuscript received June 28, 1956.
† Philco Corp., Philadelphia, Pa.

tubes, the implications of which are discussed at the conclusion of this paper.

Among the important advantages of the beam-indexing tube is its similarity to a black-and-white tube; in fact, in the absence of a chrominance signal, it cannot help making a good black and white picture. None of the writing beam in the Apple tube is intercepted or deflected in such a way as to waste any high-voltage power and there is no problem of matching the characteristics of three guns to obtain good color fidelity. As a result of these characteristics, a receiver using a beam-indexing tube can give performance superior to that of a three-gun-tube receiver. Moreover, in the opinion of the authors, the present system is potentially more economical than any other color receiver because it is simple in those portions where much of the cost of all television receivers is concentrated, and, in addition, has more possibilities of future improvement.

The two fundamental parts of the Apple system philosophy are *sequential writing* and *electrical indexing*. The expression *sequential writing* means that the beam passes successively over triplets of fine, vertical stripes of red, green, and blue phosphors, as shown in Fig. 1.

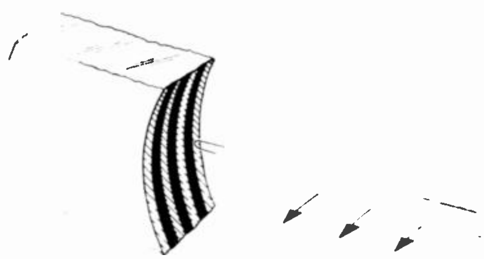


Fig. 1—Apple tube stripe structure.

A particular color is produced by modulating the beam during its passage over each triplet, according to the proportions of primaries of the desired color. The expression *electrical index* refers to a signal, derived from the luminescent screen of the Apple tube itself, that continuously gives information on the location of the beam. The beam current responds to two types of instructions: the color video signal from the transmitter and the index signal. The only circuitry unique to the Apple system is that required to perform these functions.

The signal required to produce the color picture on the beam-indexing tube resembles very closely that required by the FCC standard color television system. That is, the Apple system produces a high-quality black-and-white picture from the luminance signal and, by adding a high-frequency chrominance component to the luminance signal, produces colors. As will be shown, the similarity of these signals to the transmitted signals enables the beam-indexing tube to utilize the broadcast signal efficiently, with a minimum of color processing circuitry.

An important requirement of the system is good spot size. Obviously, when producing saturated primary colors, the spot size at peak beam current must be small enough to minimize the beam current that hits adjacent phosphor stripes. The means of obtaining small spot size are described in the companion paper.[1]

Several different forms of Apple display systems have been examined, which have their relative advantages and disadvantages. Rather than to describe these several forms, it seems preferable to discuss in detail one specific form of the system. The form chosen is that which was employed in the receiver shown at the Comité Consultatif International des Radiocommunications demonstration held in March, 1956. For the rest of this paper and for the two subsequent papers the philosophy and circuitry of this particular receiver will be described, although it will be recognized that there are many other ways by which color pictures can be made following the broad Apple concept.

Derivation of the Index Signal

The index signal is obtained from the tube by means of the structure shown in the insert in Fig. 2, where a line, called the *index stripe*, of a material having high secondary emission compared to the aluminized coating, is placed behind every red line. The secondary emission current produced as the beam crosses these index stripes is collected and amplified, resulting finally in a signal at the same frequency as that at which the beam must be varied to produce colors. This beam, which actually produces the picture, is called the *writing beam* to distinguish it from the *pilot beam* which derives the index information as described later in this paper.

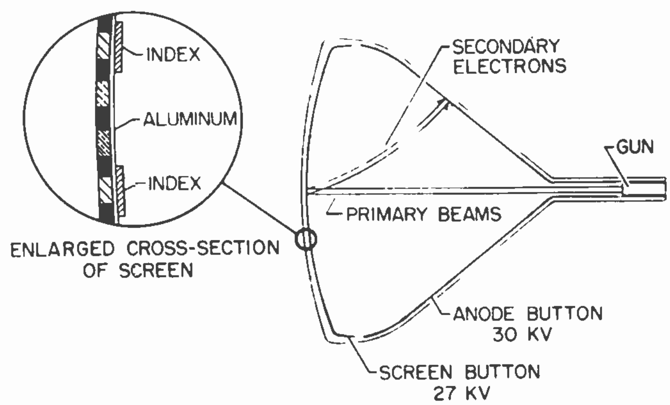


Fig. 2—Cross section of Apple tube.

The waveform produced by the index structure is shown in Fig. 3, which is an idealized curve of secondary emission ratio vs beam position. The equation representing such a structure scanned by a beam of constant amplitude is a Fourier series in cosines. The only term of interest is $A_1 \cos \theta$, the fundamental component. If the phase of this single component in the index current

[1] Companion paper, G. F. Barnett, F. J. Bingley, S. L. Parsons G. W. Pratt, and M. Sadowsky, "A beam-indexing color picture tube—the Apple tube," p. 1115, this issue.

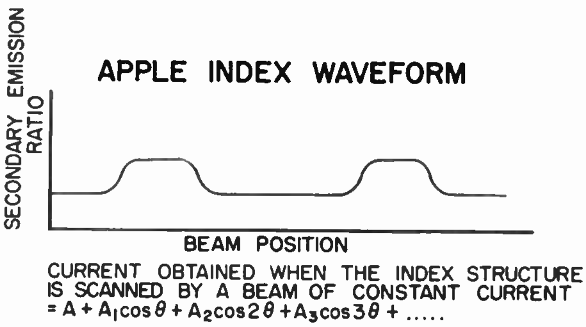


Fig. 3—Apple index waveform.

is preserved, sufficient information will be available. Amplitude variations of the coefficient A_1 are removed by means of limiters.

However, the writing beam is not of constant amplitude. The variations that produce color also produce an ac component of the secondary emission current at the same frequency as the desired index signal, and at any phase with respect to it. Since this ac component produces a perturbation of the desired index phase, the secondary emission current may not be used directly as the index signal.

The problem is overcome by the use of frequency separation. A second beam of low current, called the pilot beam, is introduced. Its beam current contains a constant amplitude component of frequency, F, called the pilot carrier-frequency; F is chosen to be above the video- and color-frequency range. An idealized diagram of the single gun which produces the two beams is shown in Fig. 4. Two sidebands are produced as the beams sweep over the index stripes, formed by the component F beating against the desired $A_1 \cos\theta$ component of the index function. Either sideband contains the desired phase information.

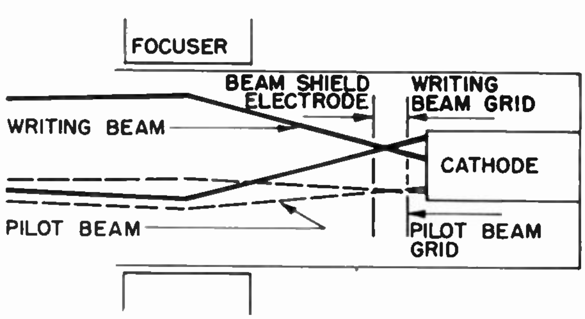


Fig. 4—Idealized cross section—Apple gun.

The pilot beam is aligned so that it always strikes the same color line as the picture writing beam. If the pilot carrier were made a part of the writing beam, gun nonlinearity might cause intermodulation between the pilot carrier and writing frequency signal. The sidebands produced by this intermodulation would have the same frequencies as the desired sidebands and might contaminate the index signal. One easy solution to this problem is the use of two beams. The pilot beam illuminates the luminescent screen at a low, constant level which affects the contrast ratio slightly. This background illumination is generally about ½ foot-lambert. A highlight brightness of 40 foot-lamberts allows a contrast ratio of 80 to 1.

The Apple Control Circuits

The index signal from the Apple tube must be amplified, combined with instructions from the transmitter and restored to the writing frequency, that is the frequency at which the beams cross the triplets, and then applied to the writing-beam grid of the tube to produce a color picture. Throughout these operations the phase of the index signal must be preserved.

A simple mixer is shown in the upper part of Fig. 5, in which the output voltage is the product of two input signals. Considering only the output terms as the sum or difference of the input frequencies, the output phase is the linear sum or difference of the two input phases, depending on which output sideband is considered.

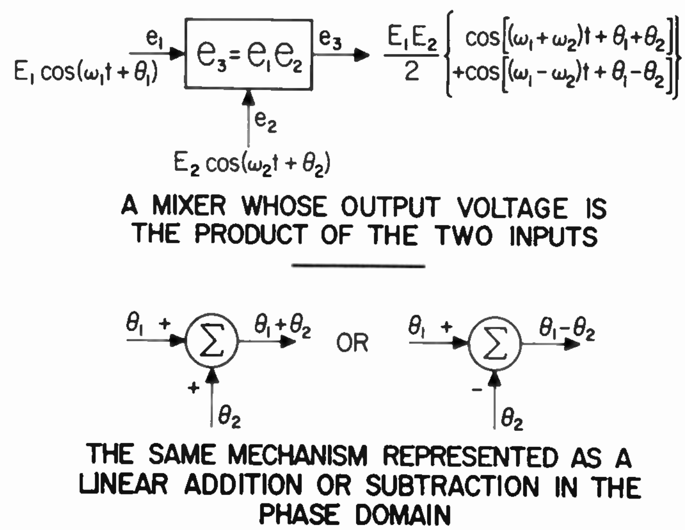


Fig. 5—An analysis of the operation of a simple mixer in the phase domain.

Thus, in the phase domain, a heterodyning process is an addition or subtraction of phases. This is shown diagramatically in the lower part of Fig. 5. The Apple index mechanism is just such a mixer, whose output is the product of the pilot beam current and the index function; and it is necessary to retain only one of the sidebands produced at the screen to obtain the essential phase information. A second heterodyning with the pilot carrier frequency is necessary to restore the sideband frequency to the original index frequency which is needed for writing colors. The block diagram of Fig. 6 shows the Apple indexing system. First, a pilot oscillator at 41.7 mc drives the pilot beam grid. The useful sideband output at 48.1 mc is amplified in the sideband amplifier and then goes to a mixer. Here it is heterodyned with the pilot oscillator output, producing the 6.4 mc signal for the writing grid.

If the horizontal scanning velocity is constant, the index signal flows to the writing grid without any change of phase, and produces variations of the writing beam current so that it illuminates successive lines of the same color; this produces a solid field of uniform color. If, however, the horizontal scan increases or decreases in speed, the ac component of the beam current is retarded or advanced in phase and so produces a slightly different color. This phase change with changing sweep speed or index frequency is proportional to the slope of the phase-frequency characteristic (the envelope time-delay) of the index sideband amplifier. Sweep nonlinearities produce color errors which are the product of the amplifier time delay and the incremental index frequency change.

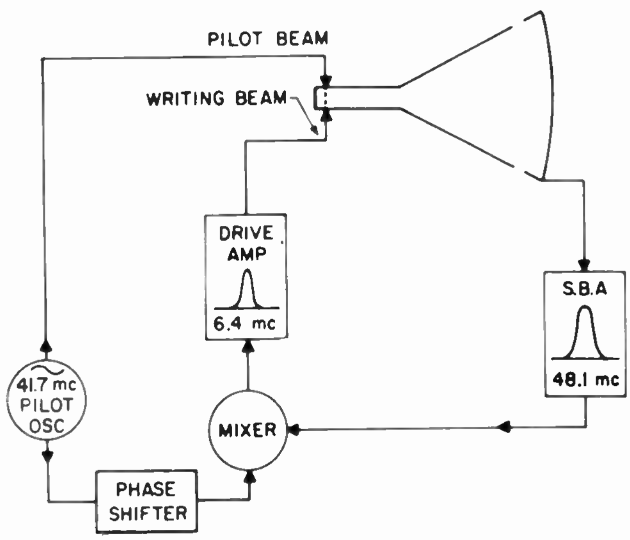


Fig. 6—Simple block diagram for generating flat fields.

In order to change deliberately the color of a flat field, it is necessary only to change the phase of the pilot carrier entering the mixer; this could be accomplished by the phase shifter shown in Fig. 6.

The optimum envelope time-delay of the sideband amplifier is affected by the possible contamination of the index signal by the writing beam. Fig. 7 shows the spectrum of the complete signal at the input to the sideband amplifier. The energy concentrations at one-half writing frequency above and below the useful sideband frequency are caused by harmonics of the writing frequency in the writing beam and video modulation of these harmonics.

For system stability the sideband-amplifier response must be well down at these points of energy concentration, for they represent interference to the index func-

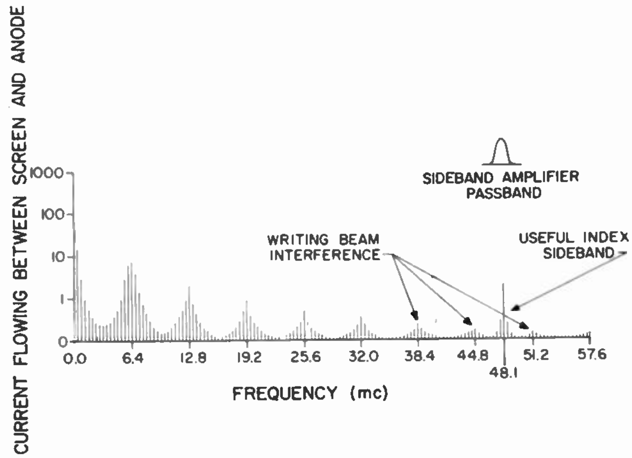


Fig. 7—The spectrum of the complete signal at the input to the sideband amplifier.

tion and can cause various forms of color interference. A design compromise must therefore be made in the selection of a sideband-amplifier response curve to have sufficient skirt selectivity to reject unwanted writing beam interference and yet have a short enough time delay to permit a realistic amount of horizontal sweep nonlinearity without too much color nonuniformity.

Typical design permits about a one-microsecond delay from the Apple tube screen to the writing grid. This delay produces about ten degrees of color error for a sweep nonlinearity of one-half per cent. Such linearity requirements, though severe, have been found to be practically realizable. Suitable circuits are described in one of the following papers.

Picture Writing Techniques

In order to make the system of Fig. 6 show complete color pictures instead of solid fields of color, it is only necessary to vary dynamically the phase and amplitude of the pilot-carrier signal entering the mixer. The voltage on the writing grid is simultaneously varied at video frequency to control the luminance. The total current illuminating the three phosphors depends on the video frequency portion of the signal applied to the writing grid, and the way in which this current is divided among the three phosphors is determined by the amplitude and phase of the writing component of the grid signal.

Consider the tube to have infinitesimal spot size and line width so that it is a sampler of very narrow aperture. Let the lines be equally spaced, as shown at the top of Fig. 8. If the video-frequency and the color-writing frequency portions of the signal are as shown in the Figure, accurate color fidelity exists. The video frequency portion of the signal is the linear sum of gamma-corrected red, green and blue, and the color-writing signal is the sum of three equally spaced vectors. This signal is similar to the combined color video available at the second detector.

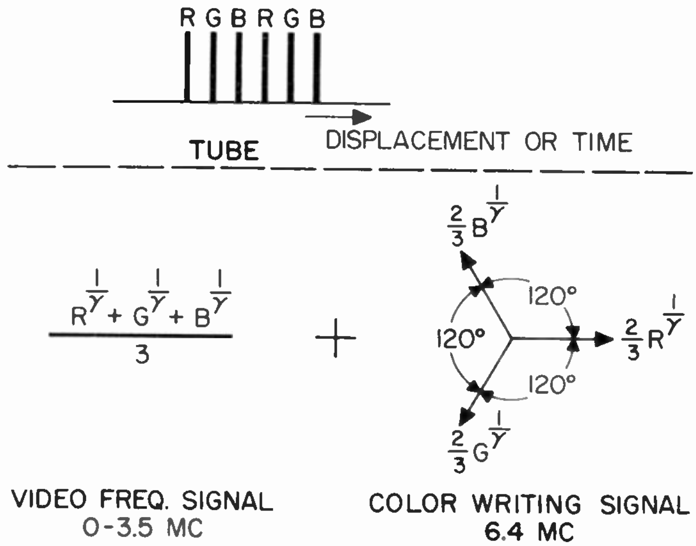


Fig. 8—Development of required signal for color writing.

The color subcarrier may be represented in terms of I and Q, or more conveniently for analysis of the present system, as shown on the left side of Fig. 9. The required drive signal is shown again on the right. Two differences exist. First, the received color subcarrier is not equi-angle; and second, the signal required to drive the tube must be locked to the stripe structure at 6.4 mc rather than to the 3.58 mc transmitted chrominance subcarrier. The color subcarrier can be simply corrected to the equi-angle form by an elliptic conversion or compensated for by unequal stripe placement on the tube, but the visible difference in the picture is less than 10 color degrees and it is customary to neglect this correction.

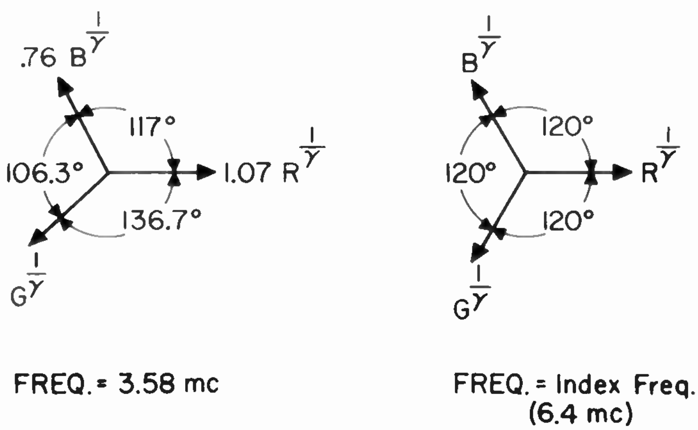


Fig. 9—Comparison of color subcarrier at second detector and required signal.

The conversion from 3.58 mc to 6.4 mc is achieved through heterodyning as shown in Fig. 10. The pilot-carrier signal required by the pilot-beam grid is generated by beating the 3.58 mc color reference signal against an oscillator 3.58 mc below the desired pilot-carrier frequency. The required pilot carrier entering the mixer is generated by beating the same oscillator against the 3.58 mc signal with color modulation. The mixer output is a signal at writing frequency locked to the color line structure but having the same amplitude and phase variations as did the original 3.58 mc color signal.

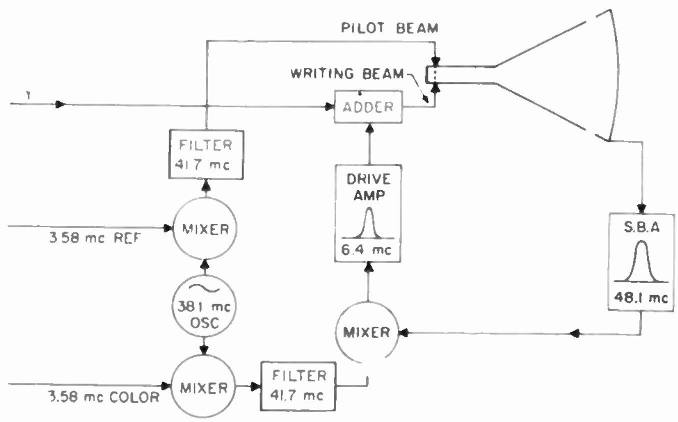


Fig. 10—Block diagram showing method of heterodyning to convert from 3.58 mc to 6.4 mc.

For the required video-frequency signal, the Y signal available at the second detector can be used as shown in Fig. 10. If slightly better colorimetric accuracy is desired, a *monochrome correction* signal of the type indicated algebraically in Fig. 11 can be added to the Y

$$\underbrace{.30R^{1/\gamma}+.59G^{1/\gamma}+.11B^{1/\gamma}}_{\text{"Y" SIGNAL AVAILABLE AT 2ND DETECTOR}} + \underbrace{.03R^{1/\gamma}-.26G^{1/\gamma}+.22B^{1/\gamma}}_{\text{"MONOCHROME CORRECTION"}}$$

$$= \underbrace{\frac{R^{1/\gamma}+G^{1/\gamma}+B^{1/\gamma}}{3}}_{} = \text{DESIRED SIGNAL}$$

Fig. 11—Addition of "Y" signal at second detector and "Monochrome Correction" signal to produce desired signal.

signal converting it to "M."[2] It is derived by synchronously detecting the 3.58 mc color signal with the 3.58 mc reference signal as shown in Fig. 12. These are all the steps that are needed if the spot size and line width are reasonably near the infinitesimal ideal. The effect of greater spot sizes and line widths is a slight desaturation of the colors. However, two simple steps can be used to correct the situation. The mathematical details are outside the scope of this paper; they involve the gamma of the tube, the line width, the width of the spaces between lines, and the exact details of the spot growth with current. Let it suffice to say that with the present values

[2] M is called the monochrome signal and is commonly defined as $\frac{1}{3}R+\frac{1}{3}G+\frac{1}{3}B$.

of these parameters, simply increasing the chroma gain about 33 per cent and adding a second low-frequency signal called *saturation correction* results in almost perfectly accurate color fidelity. The saturation correction signal biases the tube negatively an amount proportional to the amplitude of the color subcarrier and has the effect of reducing the conduction angle, particularly on primaries. The saturation correction signal is derived from the same detector as the monochrome correction, but does not represent a sacrifice in brightness since the chroma gain has been increased. Differences caused by monochrome and saturation corrections are small and recent practice has been to ignore them.

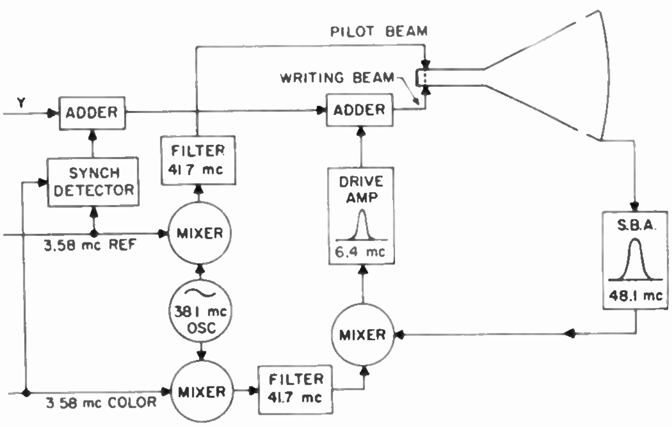


Fig. 12—Block diagram showing production of monochrom correction by synchronous detection of the 3.58 mc color signal with the 3.58 mc reference signal.

The block diagram shown in Fig. 10, then, represents the entire picture and index signal handling sections of the display. The only other parts which must be added, as shown in Fig. 13, are the sweeps, the high-voltage

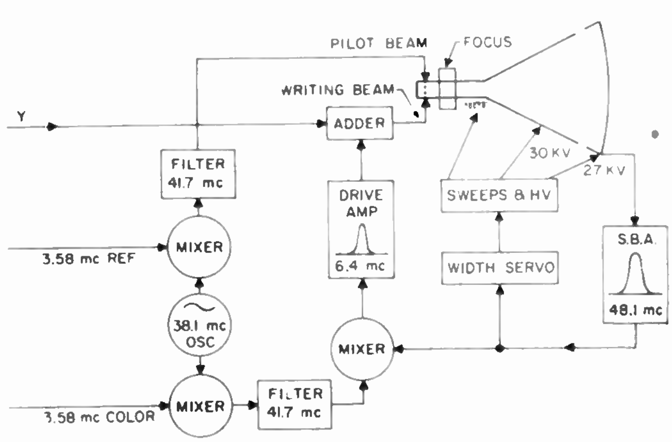


Fig. 13—Block diagram of receiver with sweep circuits and high voltage supply added.

supply with outputs at 27 kv and 30 kv, the magnetic focuser with vertical dynamic focus, and the width servo. Not shown are the reference generator, the audio, the IF and the tuner.

Comparison of the Inherent Complexities of Various Types of Color Television Display

There are several broad areas within color television receivers, and in their production methods, which should be studied to determine what degree of complexity is inherent in the display device. These areas are those of physical devices, circuitry, and adjustment procedures.

In the area of physical devices are the cathode-ray tube itself, the yoke, the focuser, any special auxiliary devices such as convergence coils, purity coils, and magnets, and any special magnetic or electric shielding which may be required. The penalty which one must pay for complexity in this area is primarily that of using more material in the receiver.

Examples of circuit complexities which are inherently associated with various display devices are the following: separate video channels for the three primary colors, abnormally wide bandwidths in any channel, regulated voltages, making two regulated voltages track each other, sweep circuits with remarkably good linearity, and the amplification of index signals. Complexity in this area leads to using more material, but it should be pointed out that apparent circuit complexity is liable to misinterpretation. The mere number of vacuum tubes in a receiver is often a very poor guide to evaluation of its true complexity. It is well known that in many cases the over-all complexity of a receiver can actually be reduced by adding tubes, provided the addition of tubes and circuits permits the removal of complicated tubes or circuits and simplification of adjustment of the receiver.

The third area, that of adjustment procedures, is the most difficult to evalutate quantitatively. The presence of the following factors is suggested as a true measure of complexity: large numbers of adjustments, either in the factory or those required of field service personnel, inter-related adjustments, adjustments whose effect is difficult to evaluate, adjustments requiring unusual test equipment, and adjustments which must be frequently repeated because of instability in the receiver or because of abnormal sensitivity to external effects. The penalties for complexity in this area are the use of more material, increased factory labor, more field service and, very often, inferior performance in the field.

There are three principal types of display to be considered: the 3-gun type represented by the shadow-mask and post acceleration tubes, the Lawrence tube, and the index type represented by the present form of the Apple tube.

The complexities of 3-gun tubes which are of most concern are those inherently associated with having three guns. The convergence of the three beams must be maintained all over the raster. This is difficult because the three beams have different centers of deflection and because of the effect of external magnetic fields, such as that of the earth. Maintenance of color purity all over the raster requires accurate control of the direction of arrival of the three beams, in the presence of the effects of external fields. Both convergence and purity vary from tube to tube. These effects have led to the use of relatively elaborate devices external to the tube to deflect the three beams, in addition to the normal sweeps. These complexities add to the material cost of the receivers, and may add as many as 25 adjustments in a typical shadow-mask receiver. These adjustments require judgment since they must be made by inspection of the effects of the adjustments on the picture; they are interrelated and they must be repeated if the physical position of the receiver is changed.

The use of three guns also imposes the requirement that the characteristics of the three guns be matched or compensated so as to produce good white balance at all brightnesses. Cathode-ray tube guns cannot easily be built with inherently matched characteristics, so circuit adjustments must be used. This adds four more critical adjustments which must be made while watching the tube face and which require judgment.

In the field of circuit complexity there is the requirement for color demodulators and separate channels for the three primary colors. The shadow mask type of three-gun tube has the further inherent disadvantage of inefficient use of high-voltage power, resulting in extra expense and complexity in the regulated high-voltage supply.

The various types of three-gun tube are relatively complex in themselves, both because of the three guns themselves and because of the other internal structures, shadow masks or grilles, required to direct the beams to the proper colored phosphors. None of these complexities exist in the single-gun index type displays.

The Lawrence tube has none of the complexities just discussed as applying to the three-gun displays, except the complexity of the tube itself and one aspect of the color purity problem. It has, however, two types of problem which are unique, and are inherent in its method of operation. One is the need for a high-power, synchronized switching signal. This causes extra circuit complexity and expense, and requires elaborate shielding to reduce radiation. The other is the need for special signal processing to obtain good colorimetry from a tube which deflects the beam sinusoidally over the color stripes. These complexities are largely in the field of circuitry. The Lawrence-tube adjustment procedures seem to be reasonably simple.

Index type displays, such as the present Apple system, have their own unique and inherent problems. One of the most important of these is the generation and amplification of the index signal. The generation of the signal is built into the tube at the expense of very little complexity. The amplification and handling of the index signal requires a moderate amount of conventional circuitry and about 30 adjustments. These adjustments represent a source of complexity only in their number. They are not interrelated; their effect is easy to evaluate since they are simple maximizing or minimizing operations while reading meters; they require no unusual test equipment; and once set up, they are stable for long periods of time.

The present Apple system requires accurate control of horizontal linearity and width. This represents a complexity only in the area of circuitry. There are no elaborate devices nor adjustment procedures involved.

Another requirement of index-type displays is small spot size. This requirement is met partly by care in tube design, which does not in itself result in tube expense nor complexity, partly by some extra complexity in the yoke and focuser, and partly by the use of a high, regulated anode voltage. The latter two requirements make the high-voltage supply almost as elaborate as that for shadow-mask tubes. The yoke and focuser alignment procedures required for good spot size are much simpler than those required in shadow-mask displays for convergence and purity, and are comparable to good black and white practice.

The only other unique and inherent requirement of the present type of Apple display is that of beam tracking. It is met partly by characteristics built into the tube at no extra expense and partly by observing the same precautions in yoke and focuser alignment as are required for good spot size.

In weighing these considerations, one reaches the conclusion that any color set is substantially more complex than a black-and-white set. By comparing the various types and severities of the complications of the different displays, one further finds reason to believe that the beam-indexing system is considerably less complex than other systems of comparable performance.

ACKNOWLEDGMENT

Obviously, this development was not the work of a small number of individuals. A great many engineers have made significant contributions. In particular, the authors wish to acknowledge the contributions of William E. Bradley, David Brunner, Monte I. Burgett, Charles Comeau, Richard K. Gardner, Richard Gudis, Lincoln Hershinger, Carl Mutschler, Charles Simmons, and David E. Sunstein.

A Beam-Indexing Color Picture Tube—The Apple Tube*

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Summary—This paper describes the Apple color picture tube, its dimensions, materials of construction, deflection and focus systems, and the geometry and deposition of the phosphor and secondary-emissive screen materials. The construction and operation of the electron gun, which produces two independent beams of very small cross section from a single cathode, are described in detail. Life test data and pilot production experience are discussed.

Introduction

The APPLE concept as described in the previous paper places some very special requirements on the cathode-ray tube. The essence of the Apple development has been to design a tube in which the task of maintaining tight tolerances is relegated to the manufacturing equipment rather than to the tube itself, where it would have to be faced every time a tube is made.

Corresponding to the many possible variations of Apple color systems, there are an equivalent number of variations of Apple tube designs. Rather than attempt to consider these in general terms, it is considered wiser to describe a specific representative example, the type of tube used in a system described in the previous paper and utilizing the circuits to be described in the following paper.

The Apple color picture tube (see Fig. 1) may be generally described as an all-glass, 21-inch rectangular picture tube providing 260 square inches of useful screen area, having a diagonal deflection angle of 74 degrees, and using magnetic focusing and deflection. More specifically, the color television display system described in the previous paper requires a picture tube that meets the following requirements:

1) The Apple tube must have a luminescent screen made up of vertical stripes of red, green, and blue phosphors that are sufficiently close together to be visually unresolvable at normal viewing distances and yet far enough apart to permit resolution of each line by the writing beam.

2) Enough triplets must be present to resolve all of the detail conveyed by the luminance component of the signal.

3) The phosphors must be so chosen that satisfactory primary colors are produced when individually excited and a satisfactory white occurs when they are excited equally.

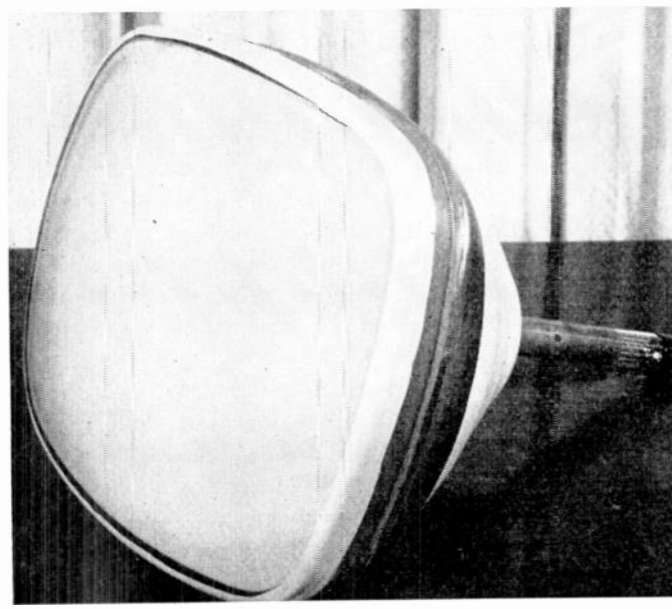

Fig. 1—Apple tube.

4) The spacing of triplets must be varied and the lines bent so that maximum circuit economy can be achieved by matching the triplet pitch at all parts of the raster to the normal sweep speed. This helps ensure constant index frequency all over the raster.

5) The lines must have sharp edges and constant width if accurate complimentary colors are to be produced.

6) There must be secondary emission index-producing stripes as an integral part of the screen. The tube should be aluminized to improve the efficiency at high operating voltages and to provide a low secondary emission base for the index material.

7) The tube must have two electron beams, a writing beam and a pilot beam. These beams should be made to track each other so that the variations in the horizontal component of separation at the fluorescent screen is small. The pilot beam can be a low-current, low-resolution beam.

8) The index lines must have the same period as the color triplets and the position of the index lines with respect to the triplets must be varied in a predetermined fashion to be explained later.

9) Since the Apple tube utilizes the entire 260 square inches of the tube face for visible picture area, it is necessary to extend the index lines a slight distance

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† Res. Div. and Lansdale Tube Co., Philco Corp., Philadelphia, Pa.

beyond the visible raster on at least two sides to be sure to get an index signal at all points.

10) The writing beam must be small enough to resolve a single color line at a peak current of 1500 microamperes, including the effect of the spot motion during the time the peak current flows.

In addition to these special requirements arising from the system itself, it is desirable that the tube be amenable to mass production and utilize as much as possible existing facilities and techniques in its manufacture.

Electron Gun

The color saturation obtainable at any particular brightness level in a beam-indexing tube is obviously limited by the spot size at the beam current associated with that brightness. If the spot is too large to land on one primary color stripe at a time, then desaturation of primary colors occurs. This consideration, plus that of reasonable structural resolution, made the development of an electron gun capable of producing a spot substantially smaller than usual in a monochrome tube a prime necessity for a beam-indexing tube. Small spot size is obtained in the Apple beam-indexing tube by ingenious utilization of electron optical principles, and by maximum simplification of the electron optics.

The electron gun is essentially of magnetic focus, triode design. Magnetic focusing was chosen over electric focusing for two reasons. First, for any particular tube-neck diameter, magnetic focusing permits the use of a larger lens diameter than does electric focusing. The beam diameter being the same in either, less aberration occurs in the larger lens. Second, the external magnetic lens can be accurately aligned to the electron beam after the tube is assembled, reducing tube scrap from gun misalignments.

The focused spot size has, as one limitation, the size of the first crossover of the electron beam. An extensive investigation was conducted to determine the effect of electrode configuration on the formation of the first crossover. Equipotential plots were made of many configurations of elements using a resistor network to simulate, on a greatly enlarged scale, the fields that would exist between the electrodes involved. Ray traces made, utilizing these field plots, *indicated the diameter and current density variation of the crossover vs electrode configuration and potential.* These studies confirmed that there were no limits on crossover diameter precluding the development of a practical beam-indexing tube, but that a cathode loading higher than usual in picture tubes would be necessary to achieve this small crossover.

The required crossover diameter is secured by close cathode-to-grid spacing, small grid aperture diameter, and a *thin* grid aperture as shown in Fig. 2. Techniques were developed which made extremely close cathode-to-grid spacings possible by using a spacer ceramic which is lapped top and bottom to a specified height. The cathode-support ceramic is also lapped flat on one side, and the dimension from this surface to the top of the uncoated cathode is closely controlled. Cathode-spray thickness is also closely controlled.

The writing-grid aperture is 0.020 inch diameter, and the beam has a bogie cutoff of 150 volts. The pilot-beam aperture is 0.014 inch diameter, and produces a bogie beam cutoff of about 50 volts. The grid aperture is made electrically *thin* by countersinking the hole so as to leave the cylindrical portion only 0.001 inch thick.

This combination, then, of small, countersunk grid aperture and close cathode-to-grid spacing, is primarily responsible for the small diameter first crossover, which is imaged on the screen by the simple electron optics described above and results in greatly reduced spot size.

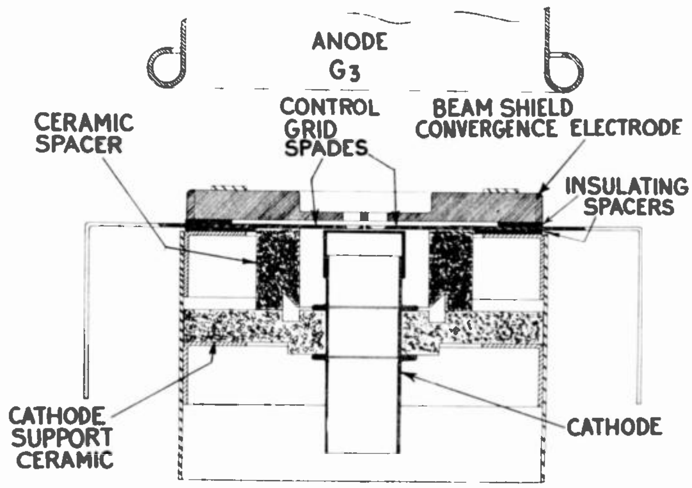

Fig. 2—Schematic line drawing showing gun details.

The second requirement is that the two beams *track* each other. Since one beam is used to tell where the other beam is, the relative position of the beams must be known at all times. When this positional relationship of the two beams follows a predictable law throughout scanning, the beams are said to *track*. In order that this tracking relationship be independent of manufacturing variations in deflection yokes, the two beams must traverse the same portion of the deflection field at the same time. For the yoke designed for use with this tube, the optimum situation is for the two beams to originate as closely together as possible and cross each other at the center of deflection.

The two beams are formed close together by using a single cathode and two separate, coplanar control grids, each with its aperture close to the end of the piece, the ends being separated by 0.002 inch. The center-to-center separation of the two beams at the grid plane is only 0.029 inch.

Convergence of the two beams so as to cause them to cross at the center of deflection is obtained by a *field lens* type of convergence electrode. This lens slightly bends the two beams toward each other without any appreciable focusing effect. The convergence electrode is actually part of the *beam shield* whose function will now be described.

The third special requirement of this beam-indexing tube arises from the need for preventing the control voltage of one beam from affecting the intensity or position of the other beam.

Without shielding, a signal applied to the control grid of one beam was found to produce both deflection and intensity modulation of the other beam. However, a simple shield *between* the two beams in the region just above the grid apertures effectively eliminates beam *crosstalk* as a limitation of the functioning of the system.

The beam shield takes the form of a thin, flat disc having two small holes with a bridge of metal between them. This beam shield is *not* a conventional accelerating electrode, and every attempt has been made *not* to have it perform any accelerating function. If the beam shield is operated at such a potential as to accelerate the electron beam, it obviously becomes an electron lens of very small diameter. The beam would fill a substantial portion of this lens with resulting aberration.

Reduction of lens action is accomplished by operating the beam shield at its average free-space potential and by keeping it thin. By field plots of the equipotentials in the region above the control grids, it was found that the equipotentials in this region are relatively flat and so are not appreciably distorted by a thin disc such as the beam shield. When operated at 600 volts the beam shield is, at worst, a very weak lens and produces only minor aberrations.

Luminescent Screen

The luminescent screen of the Apple tube consists of a repeating array of red, blue, and green vertical stripes. The stripes are not contiguous but have 50 per cent duty factor; that is, the spaces between the lines are as wide as the phosphor lines themselves. The spaces between the lines are filled in with a guard band made of a dark-colored, nonluminescent material. The presence of this band insures accurate line width, improves color saturation, and enhances contrast under normal ambient light by reducing the reflectivity of the screen.

Correct white balance is built into the screen of the Apple tube by adjusting the relative efficiencies of the blue and green phosphors by the addition of varying amounts of nonactivated material so that scanning of the screen with a constant, unmodulated beam produces white.

The phosphor array is not quite the simple structure of repeated lines described above. The triplet pitch, as mentioned above is varied to match the normal sweep speed as shown in Fig. 3. Another example of matching the screen geometry to the electron optics is the progressive curving of the phosphor lines from center to edge. This is much exaggerated in the drawing of the figure. The slight pincushioning corrects for the small amount of corner twist in the relative positions of the two beams caused by certain field parameters in the deflection yoke.

The guard bands and phosphor lines are placed on the inside of the tube face by a photoresist technique using dichromate sensitized polyvinyl alcohol.

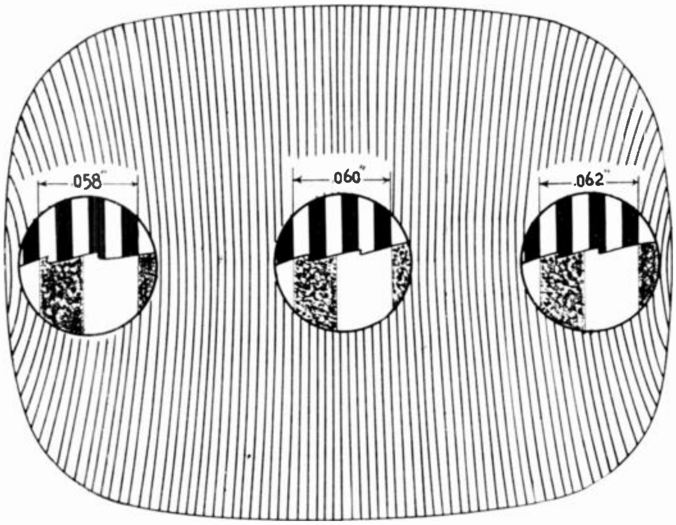

Fig. 3—Line drawing showing details of screen structure.

In order to cancel out any variations in the glassware of the tubes and thus increase possible glassware tolerances, the array of lines is placed on the face of the tube by a light projection system, schematically diagrammed in Fig. 4, in which the optical paths are made as nearly like the electron paths as possible. Thus, the projection lens has its optical center at the electron center-of-deflection. In order to make the exposure, the bulb is

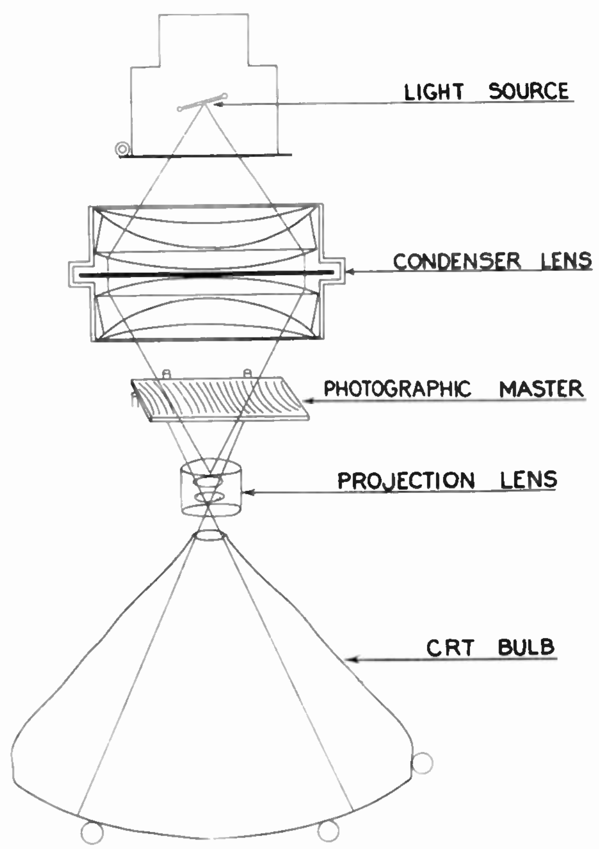

Fig. 4—Schematic line drawing of projection system.

open at a point where a flared neck with a flare diameter of about $2\frac{1}{2}$ inches may be sealed to the funnel.

The exposure equipment consists of:

1) a high pressure mercury-arc light source,
2) a wide aperture condensing lens,
3) a wide angle projection lens,
4) a kinematic mounting for positioning the bulb, and
5) a precision photographic line master.

The light source is conventional.

While of special design, both the condensing lenses and projection lenses were designed and produced using well-known techniques.

The kinematic mounting device, which permits simple, precise relocation of the bulb in the projector, uses six fixed, hardened steel balls. Three of them are in contact with the face of the bulb; two are in contact with one long side of the panel, and one with one short side. Relocation with 180° rotation is avoided by observing the location of the anode contact button.

The precision with which the phosphor lines can be placed on the tube with respect to each other depends upon the stability of the projecting equipment, accurate bulb-repositioning, and the precision of the line masters. Once the proper line masters have been prepared, however, precise reproductions of the tube luminescent screens are achieved without difficulty. The precise screens are achieved without difficulty. The precise relative position of the lines is built into the glass photographic masters and thus need not be built into each tube.

A complete discussion of the preparation of the photographic line masters would require too much time to be covered in detail here. It has entailed the design and construction of unique equipment and the development of a number of unconventional techniques. With this equipment precision linear rulings on glass or metal are converted into sets of properly distorted photographic masters, one each for red, blue, green, black, and index line deposition.

During exposure from the inside or gun side of the face plate, hardening of the resist occurs from the surface down toward the glass as exposure proceeds. If the phosphor and photoresist were mixed, an absolutely uniform layer would have to be deposited; otherwise, heavy sections would be under-exposed and not affixed to the glass, or if the exposure were long enough to ensure complete adherence of all desired areas to the glass, the phosphor particles acting as a dispersing medium, would reduce the precision and delineation of detail possible in the line pattern.

These difficulties are eliminated by first coating the bulb face plate with a film of clear photoresist which is then exposed. The exposed photoresist film is coated with a phosphor slurry, dried, and washed off. The *unexposed* areas of resist wash off readily, carrying phosphor from these sections with them. The exposed areas remain, holding a uniform layer of phosphor which adheres to the exposed resist lines.

The dark guard bands are applied first using the above described process but substituting a dark, non-cathodo-luminescent material for the phosphor. The red, blue, and green lines are then applied, one color at a time, using the appropriate photographic masters, and completely filling the spaces between the dark guard bands.

Index Structure

The final unique feature of the beam-indexing color tube is the index structure which provides the required continuous monitoring signal. This signal is generated by the difference in secondary emission between an array of magnesium oxide stripes applied to the gun side of the aluminized screen and the bare aluminum between these stripes as shown in Fig. 5. There are two contact buttons on one side of the tube envelope, and one of these is connected to the screen aluminum coating, making it possible to maintain the screen potential at approximately 27 kilovolts.

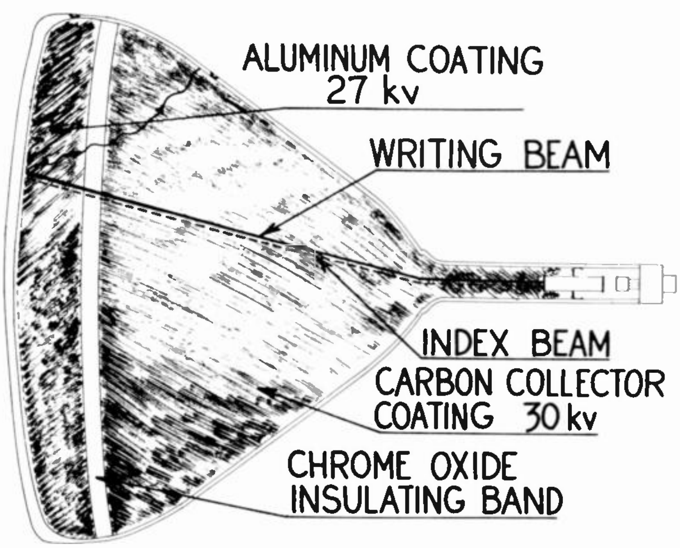

Fig. 5—Cross section of Apple tube.

The second contact button connects to the bulb coating which is maintained at 30 kilovolts. The 3 kilovolt differential between screen and bulb coating results in collection of the secondary electrons from the screen by the bulb coating.

The screen is aluminized in conventional fashion. An organic lacquer film is then applied by a simple flow-on technique to the gun side of the aluminum film. The lacquer strengthens and protects the aluminum during the application of the magnesium oxide stripes. The magnesium oxide stripes are applied to the lacquered aluminum in exactly the same way the phosphor lines were applied to the glass, except that a different photographic master is used.

There is one magnesium oxide stripe per triplet. The index stripes are on a 40 per cent duty factor, that is, 40 per cent of the triplet width is magnesium oxide, 60 per cent bare aluminum. This has been found to give the maximum fundamental component index yield. The distortions built into the index lines, while related to the distortions built into the phosphor lines, contain a corrective component, the controlled displacement of the index stripes, to compensate for index transit-time variations and tracking variations. Transit time varies enough to produce a phase shift of over 90° at side-band frequency between the center and edge of the screen. Making the transit time uniform is more difficult and expensive than moving the index structure laterally enough to compensate for it.

An interesting feature of the testing of the beam-indexing tube is the examination of the index structure. This may be done in detail by simply using the tube as though it were a monoscope and displaying the secondary emission pattern of the screen and index structure on a monitor tube.

Pilot Production and Life Test

Several years of development work and many months of pilot production activity on the Apple tube have demonstrated its reproducibility in manufacture. Equipment requirements, other than those required for the screening operation, are only those required to manufacture monochrome tubes.

Extensive, long-range life tests have failed to show any signs whatsoever of index-deterioration with either shelf life or operating lifetime up to 10,000 hours. In fact, *no* measurable changes in index yield for the whole screen or any part of it, have been noticed on any of several hundred life-test tubes.

Cathode emission problems at present loadings are not substantially different from monochrome tubes, and are believed to be less troublesome than might be encountered in tubes having more internal hardware, or multiple guns.

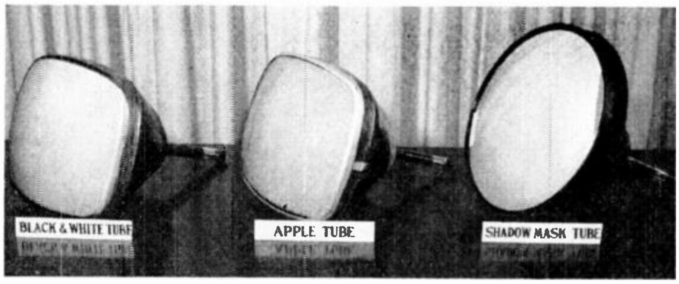

Fig. 6—Monochrome, Apple and Shadow Mask tubes.

A comparison of the finished Apple tube with a monochrome tube, as in Fig. 6, shows the same size envelope for the same size picture. Compared to the other color tubes, the Apple tube presents the largest picture size in proportion to the envelope size.

No metal sealing flanges are present, and no new techniques or equipment for making large panel funnel seals are required by the tube manufacturer.

The electron gun, shown in Fig. 7, is more like a monochrome gun that it is like any multiple beam gun used in other types of color reproducing tubes.

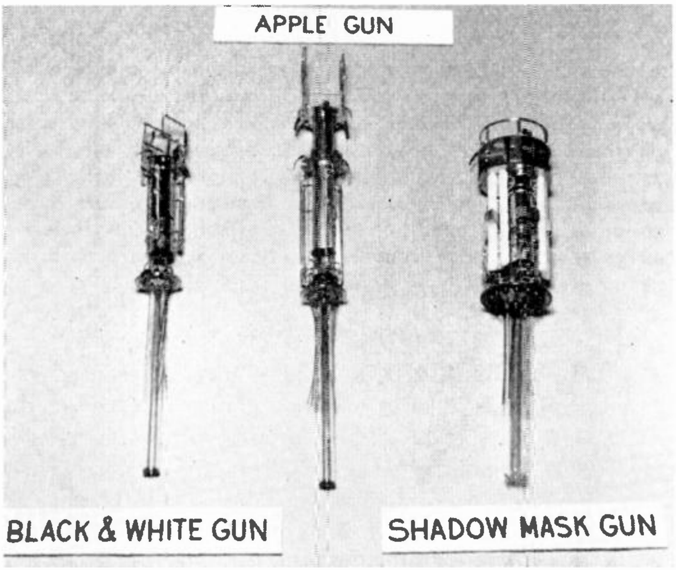

Fig. 7—Monochrome, Apple and Shadow Mask guns.

Conclusion

In conclusion, the beam-indexing color picture tube is believed capable of producing high-quality monochrome and full-color pictures. Resolution and brightness are outstanding.

In the opinion of the authors the tube permits potentially lower-cost manufacture than other types of color tubes.

Its manufacturability and life potentiality have been demonstrated to be satisfactory.

Acknowledgement

A major development work such as the Apple tube project requires the assistance and cooperation of a substantial number of persons over an extended period of time, and thanks and acknowledgement are due to many who cannot specifically be mentioned by name. Specific mention can be made only of a few whose contributions were unusually outstanding and who have been identified with this project over a considerable period of time. Early original suggestions came from C. Bocciarelli, A. Rittmann, and J. Tiley of the Philco Research Department. P. D. Payne and G. R. Spencer of the Philco Tube Development laboratory made many suggestions embodied in the present tube design or processing. H. R. Colgate, in charge of the pilot-plant engineering group of the Lansdale Tube Company, was instrumental in reducing to production practice many new techniques.

Current Status of Apple Receiver Circuits and Components*

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***Summary*—This paper reviews the components and circuits of a developmental color television receiver utilizing the Apple type of display. The block diagram of the complete receiver is presented, together with detailed diagrams of circuits peculiar to the display, *i.e.*, the index signal amplifier, color signal processing, high-voltage, horizontal sweep, and focus circuits. Photographs of the chassis layout and electron optical assembly are presented. The problem of integrating the circuits is outlined and typical receiver performance figures are stated.**

Introduction

THE COMPANION papers have presented the groundwork from which a complete receiver design can be developed. A program of successive receiver designs has been carried on and this paper will describe a version of receiver number seven. Historically, it should be pointed out that early receivers were built to provide technical information and as such they included many alternate versions of circuitry. Unfortunately, this led to many now unrealistic "guesstimates" of receiver circuit complexity. A full gamut of pulse-type writing drives by remodulation and heterodyne methods, carrier drives (including equiangle subcarrier correction), and many forms of sweep and high voltage circuits were explored. Only recently has an attempt been made to attack the receiver problem as a comprehensive design. The circuitry of this receiver was integrated early in 1955.

Construction

The construction of this receiver is shown in the accompanying illustrations. Fig. 1 is a front view of the receiver. The important features are the 260-square-inch rectangular screen, and the control locations. The customer-type controls in the control bar are horizontal sync, vertical sync, focus, contrast, hue, and chrominance. All required setup controls are accessible either through the slot below the control bar or at the rear of the chassis. The conventional adjustments are: 1) Horizontal oscillator frequency, 2) Width, 3) Parabola waveform for horizontal linearity, 4) Vertical linearity, and 5) Height. The special color receiver controls are: 1) Dynamic focus amplitude, 2) Sawtooth waveform for horizontal linearity, 3) Width modulation parabola waveform, 4) Width modulation sawtooth waveform, 5) Pilot carrier bias, and 6) Master hue.

Only six of these eleven setup controls are not found in a usual monochrome receiver. Of these six, only three appear to be required in future receiver designs.

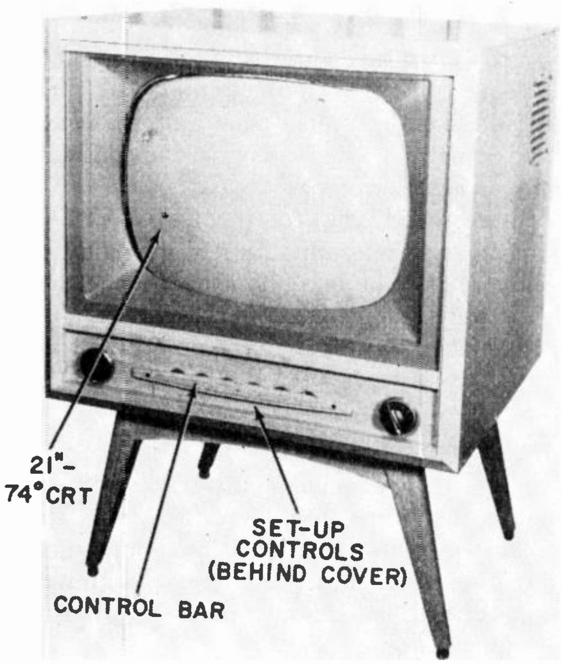

Fig. 1—Front view of Philco color receiver.

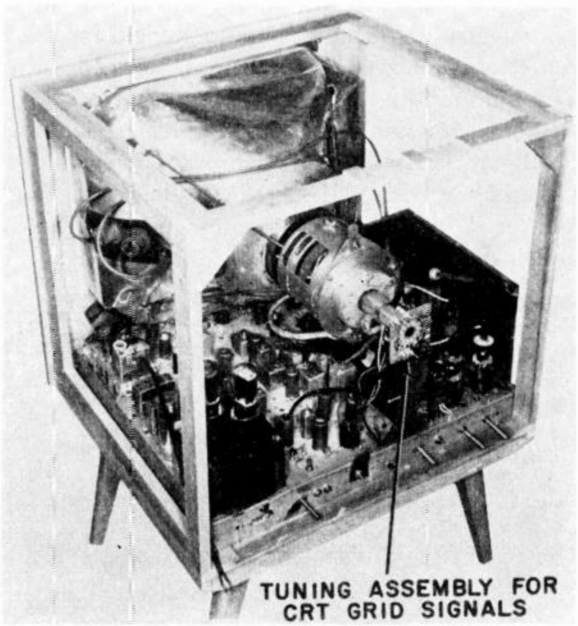

Fig. 2—Rear view of receiver assembly.

Fig. 2 shows the rear view of the receiver assembly. A small plate, associated with the crt socket, provides a location for pilot carrier and writing grid circuits.

* Original manuscript received by the IRE, June 4, 1956; revised manuscript received, June 28, 1956.
† Adv. Dev. Lab., TV Div., Philco Corp., Philadelphia, Pa.

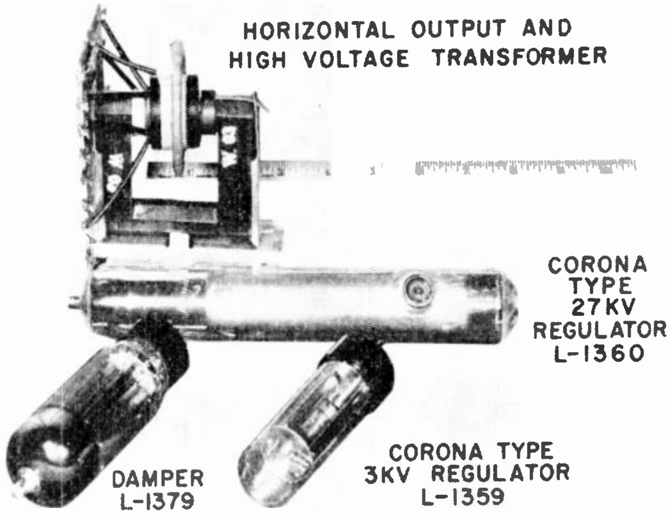

Fig. 3—View of the experimental damper tube, 3 kv regulator tube and 27 kv regulator tube.

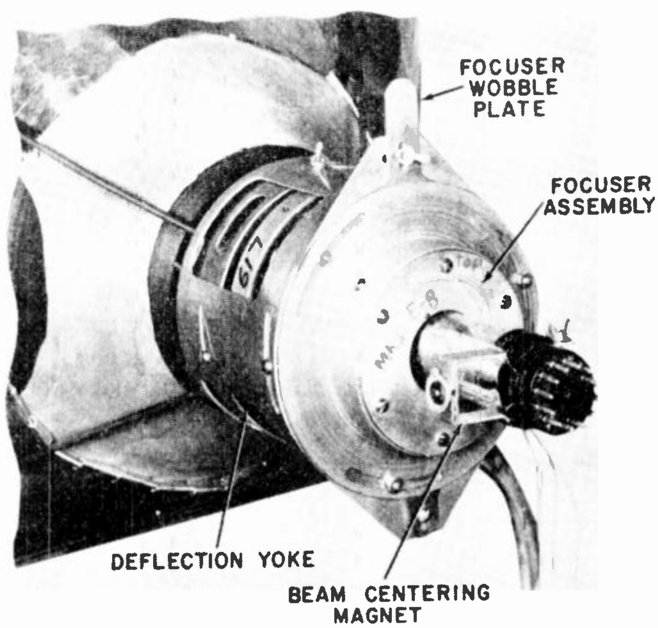

Fig. 4—View of experimental yoke-focuser mount.

The pilot carrier signal is coupled from the chassis on the coaxial lead shown. Writing-frequency signals are carried by open wire leads at present.

Obviously, good bonding of the chassis to the crt assembly is required. This is achieved by foil straps at front and rear of the chassis and crt. The chassis construction is conventional and the chassis size is 21 by 24 inches. The only nonstandard components associated with the chassis are the tubes in the sweep high-voltage section.

Fig. 3 shows the experimental damper tube, the 3-kv regulator tube, and the 27-kv regulator tube used in the receiver. The horizontal output transformer is included for comparison purposes.

Fig. 4 is a view of the experimental yoke-focuser

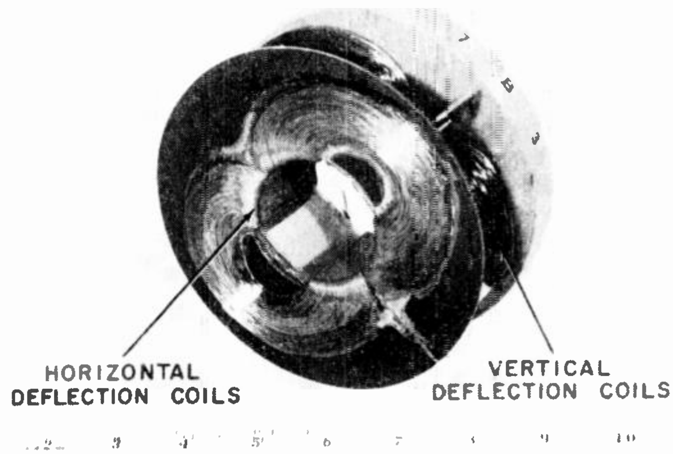

Fig. 5—Details of the yoke.

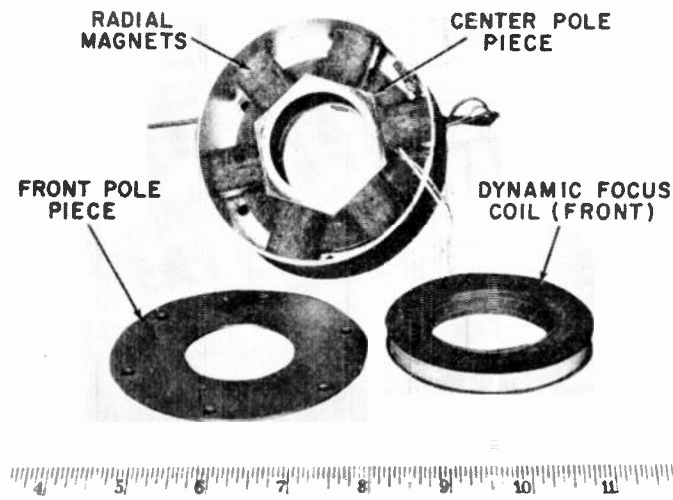

Fig. 6—Construction details of the dual gap focuser.

mount. It is conventional, with the yoke centered from the tube neck and the focuser carried by a "wobble plate" in a plane normal to the tube neck. The focuser is mechanically aligned to the tube neck by adjusting the "wobble plate." A beam centering magnet is used to obtain writing beam alignment.

Briefly, the complete electron-optical alignment is as follows. First, with vertical deflection only, the yoke is rotated so that a vertical beam trace is aligned with the color stripes. Second, the centering magnet is adjusted to place the beam along the focuser axis. A modulation pattern obtained by 60-cycle connections to the dynamic focus coils determines the unique position of proper beam alignment. These two steps complete the alignment of the crt assembly.

Fig. 5 shows details of the yoke. The cylindrical windings have an inside diameter of two inches and a core length of $1\frac{1}{2}$ inches.

Fig. 6 shows the dual gap focuser construction. The permanent magnet field is supplied by six radial magnets and provides 90 per cent of the focus strength. The electromagnetic field for dynamic focus is supplied by a pair of coils within the housing. Dual gap focusing has been used to prevent a rotation of the axis of the dual beam focused spots, relative to the screen structure, with variations of focus strength required by the dynamic focus action.

The remaining important part of the crt assembly is shown in Fig. 7. To obtain an indexing signal from the screen an external band of a conductive coating encircles the screen viewing area to form a coupling to the screen aluminum film. A metal mounting band is strapped over this coating and is used to support the tube and yoke-focuser cup in a manner similar to that found in monochrome receiver practice. The mounting feet, which support the assembly in the cabinet, are insulated from the metal mounting band, but are grounded to the foil shield. The circuit elements formed by the band, the aluminized screen to which it is coupled, and ground, are tuned to resonance at the index sideband frequency. Index signal take-off is accomplished by a coaxial lead connected to the mounting band. To shield the index circuit from external interference an aluminum foil shield is folded to cover the rim of the crt. In the figure one corner of this shield has been opened for purposes of illustration. The tube and circuitry do not appear sufficiently sensitive to magnetic fields to require any magnetic shielding or compensation for earth field effects.

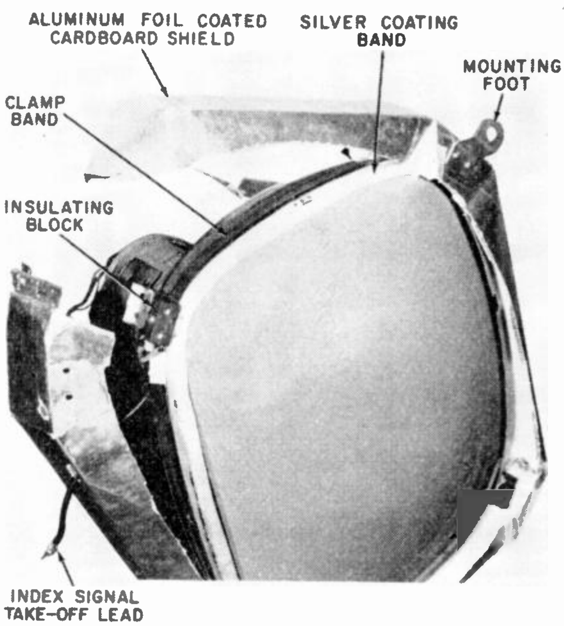

Fig. 7—View showing details of crt mounting assembly.

The foregoing explanation of the details of the crt assembly may give an understanding of recent practice in this area. Further simplification of the physical structures external to the crt are expected with operating requirements better known now than at the start of this receiver design.

Receiver Circuitry—Block Diagram

Details of the circuitry of this receiver are shown in Fig. 8. This is one version of a complete Apple receiver.

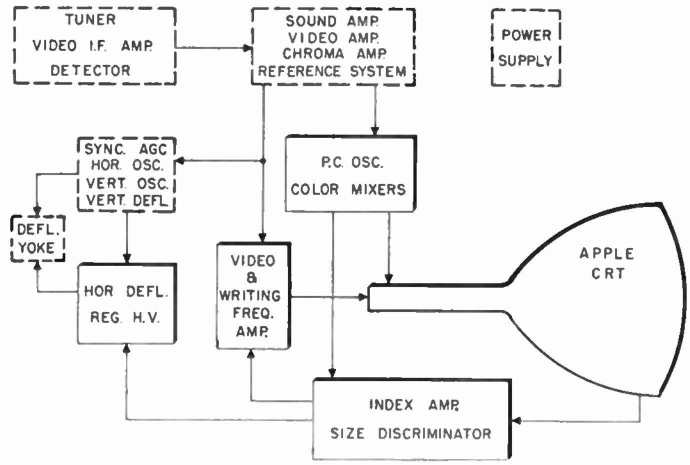

Fig. 8—Block diagram of the Apple receiver.

The circuits shown in the light dashed lines follow conventional color receiver practice and the Apple receiver imposes no special requirements here. In fact since the reference system and chrominance amplifier drive single channel signals to relatively low-level mixers they require only nominal attention. The remainder of the receiver outlined in heavy solid lines is shown in four separate sections: the index amplifier, the pilot carrier oscillator and color mixers, the writing frequency amplifier, and the horizontal sweep and high voltage.

Fig. 9 shows the circuit arrangement in more detail.

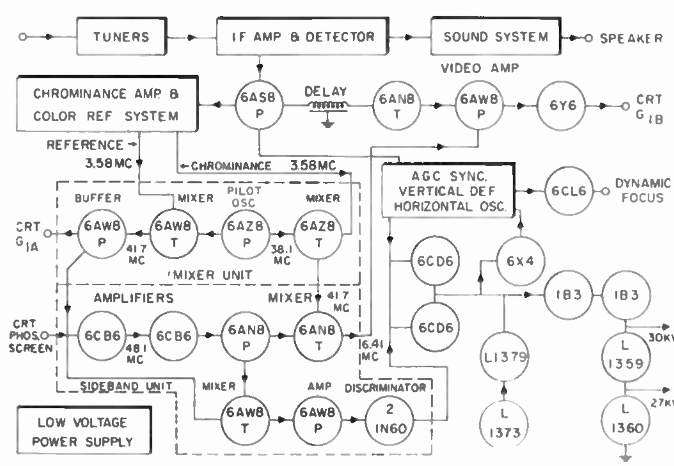

Fig. 9—Functional diagram of the Apple receiver.

The horizontal sweep-high voltage section is very similar to monochrome practice. To date, a pair of 6CD6 tubes have been used for the drive, and a special high-perveance diode, the L-1379, is the damper. The 30-kv supply is obtained by a voltage doubler of 1B3's. Six special points about this circuit should be mentioned.

1) A "width" control, for the close but long time-constant control, of the average color writing rate, is obtained by controlling the average bias on the drive tube grids with the output of a writing frequency discriminator. This precludes grid leveling on the grid drive signal but is otherwise a satisfactory control which has been found not to disturb the sweep linearity.

2) Sweep width modulation at the vertical scanning rate to match the raster pincushion to the color line pinchushion is provided by a small amount of drive tube bias variation with vertical parabola and sawtooth components derived from the vertical output stage.

3) To aid in maintaining horizontal sweep linearity with changes in line voltage, and to maintain a nearly constant picture height, it appears advantageous to derive the plate supply voltage for the horizontal and vertical oscillators from the regulated energy in the horizontal system. The 6X4 shown provides a 400-volt supply for this purpose.

4) An antiringing damper, the L-1373, is used to suppress transients of the output transformer which would otherwise appear as uncontrolled variations in the horizontal sweep linearity. This is a small tube requiring a rating of about 50 milliampere average current and 1 kilovolt peak inverse voltage.

5) Vertical dynamic focus only is used in this receiver and for this a vertical frequency parabola is applied to a focus control tube.

6) The high-voltage supply must have two regulated outputs for reasons of maintaining optimum focus, horizontal sweep operation, and index. This regulation requirement has been accomplished by use of L-1359 and L-1360 all glass gas regulators developed by the Philco Lansdale laboratories. Although this type of tube has a somewhat clouded past history the newly developed types have been tested thoroughly and have been highly successful, giving no trouble in receiver operation. The three requirements for stability of width, linearity, and high-voltage regulation can be met by these circuits.

The mixer unit consists of two tubes whose triode sections accomplish nearly all the color signal processing required by the receiver. The functions of this section are to generate an unmodulated pilot frequency carrier, and to transfer the chrominance modulation to a second pilot frequency carrier. To supply the unmodulated pilot carrier signal, a pentode is used as a 38.1 mc oscillator. Oscillator output is mixed with 3.58 reference signal and the sum frequency of 41.7 mc is selected, amplified, and applied to the crt pilot signal grid where about 40 volts peak-to-peak are required. The pilot carrier bias control previously mentioned is a dc bias control on this crt control grid. Pilot oscillator output is also mixed with the receiver chrominance signal and again the sum derived to form the chrominance modulated signal of 41.7 mc. The essential requirements of this mixer section are nominal and may be summed up by the admonition, "Eliminate stray couplings."

The mixing of index signal and pilot carrier beam current occurs at the crt screen and permits frequency separation of index information as a sideband of the pilot carrier from color writing information. The sideband unit is, at this moment, the largest special circuit group in the receiver but it is also the most straightforward. It comprises a 3-stage amplifier with a center frequency of 48.1 mc and 2-mc total bandwidth. Required selectivity is achieved in passive circuits associated with the index take-off circuit and ahead of the first amplifier. This amplifier has two outputs. One is mixed with the unmodulated 41.7 mc pilot carrier to form a color writing frequency difference signal that is applied to a discriminator to derive the width control signal previously noted. The second is mixed with the chrominance modulated 41.7-mc pilot carrier to form a chrominance modulated writing frequency signal which also includes the positional information of the index signal.

The functions of the video amplifier are normal. The luminance signal from the detector is amplified and applied to the crt writing grid. The chrominance signal from the sideband unit is amplified by the last two stages of the video amplifier, and with the luminance makes a composite video signal for the crt writing grid. About 150 volts of peak-to-peak signal, including the sync pulse, are desirable to achieve 40 foot-lambert highlight brightness pictures. The master hue control is located in the reference system for the purpose of aligning the phase of the final writing signal with crt screen structure.

The circuits shown in Fig. 9 comprise a fully operable receiver that is capable of making excellent pictures. However, many embellishments of the circuits of Fig. 9 are possible and indeed all of the following have been tested in prior receivers.

1) Monochrome correction of the "*Y*" signal to an "*M*" signal.[1]

2) Chrominance signal correction from the transmitted vector relationship to an equal angle signal.

3) DC restoration or dc coupling of luminance or chrominance signals.

4) Sundry writing frequency circuit processing for enhancement of saturation by control of writing frequency signal conduction angle.

[1] The "*M*" signal is defined in the companion paper by R. G. Clapp, E. M. Creamer, S. W. Moulton, M. E. Partin, and J. S. Bryan, "A new beam-indexing color television display system." p. 1108, this issue.

The action of these circuits has been found to be as one would predict, yet final usage in a receiver is not easy to establish since each added circuit has its drawbacks. The situation is not unlike dc restoration in monochrome receivers. In the present receiver design good colorimetry appears to depend more on the amplitude linearity of the circuits than on additional circuit functions.

Over-All Receiver Integration

There are several points of over-all receiver operation which should be singled out for comment.

Amplifier Delay

There are two related requirements on the index and writing circuits of the Apple receiver.

1) Selectivity must be adequate to minimize cross-talk of writing beam information into the indexing signal.
2) Delay must be short enough to index the chroma writing information accurately to index beam position.

These somewhat opposite objectives are satisfactorily realized by localizing the major selectivity at the sideband amplifier input. Subsequent amplifiers, including the sideband mixer and the video stages handling chroma, are broad-band, typically 5 mc, to give an over-all circuit bandwidth of approximately 2 mc. In addition, all color processing is kept outside the index amplifier chain to minimize amplifier delay. The circuit delay of the Receiver 7 circuits is approximately 0.9 microseconds.

Sweep Velocity

With amplifier delay of this order the constancy of hue with scanning is affected by the constancy of the index frequency. Since the index frequency is produced by the horizontal scanning of the pilot beam over the vertical index line structure, realization of an essentially constant index frequency depends on the proper correspondence of sweep velocity and the screen index line geometry. Receiver 7 relies upon circuits similar to those found in monochrome practice and on component stability for its proper operation. The yoke current waveform is of exponential type realized with a developmental low-impedance damper tube, and the crt index line pitch along any horizontal scanning line is a matching exponential. With the average index frequency held accurately by discriminator control of horizontal scanning width, a match between crt color line geometry and raster geometry is achieved with resistive sawtooth and parabola waveform controls.

External Field Influences

It has already been noted that no magnetic shielding has been used in this receiver as the earth's field effects are negligible. In addition, experience has shown that unusual care to avoid hum fields in the vicinity of the crt is unnecessary. This results from the close proximity of writing and pilot beams which prevents an error-producing differential action from taking place.

Typical Receiver Performance

Several characteristics of the over-all receiver performance are of interest. In a manner similar to the makeup of the transmitted color signal, the color processing circuits are in no way required to make a monochrome picture. The colorimetric white point of the picture is determined by the crt screen and there is no static or dynamic white balance problem. Resolution is certainly not limited by spot size because the spot can be no more than $\frac{1}{3}$ of a brightness picture element. The contrast ratio is excellent, since there is no secondary emission problem at the screen and the pilot beam current can be less than 1 per cent of the peak writing beam current. Screen face reflectivity is low due to the nonreflecting guard lines. Colorimetry can be as good as the circuitry one may wish to include in the receiver, and elementary circuitry has been found adequate.

To achieve a primary color, a color line must be resolved by a moving modulated spot and hence there is the requirement to make the screen structure (that is the phosphor stripe width) as coarse as permissible and to drive the beam to a current limited by the focused spot size. This determines the available highlight brightness on color pictures. The preceding paper and the preceding description of this receiver indicate that excellent results have been achieved in this area. The present receiver performance is 40 foot lamberts highlight brightness with good primary color saturation. Figs. 4 to 7 show the simplicity of the crt assembly and the list of receiver setup controls is complete. Circuit progress has led to chassis simplification such that the chassis of Fig. 2 is 21 by 24 inches and contains the complete receiver including the power supply. This receiver, as a developmental type, does not use an excess of dual section tubes, yet its complement is only eight tubes more than a shadow mask receiver containing the same nondisplay circuitry. In the foreseeable future this differential may be not more than five tubes. This seems a small price to pay, in so-called circuit complexity, to gain an electron-optical system requiring only two alignment adjustments and a cathode-ray tube completely free of static and dynamic white balance and magnetic field problems.

Acknowledgment

The contributions of many engineers are included in this receiver program. Among others, the long term contributions of the following are gratefully acknowledged: E. G. Clark, H. B. Collins, E. J. Quinlan, P. W. Scholtes, W. F. Simon, H. H. Wilson, Jr., and P. G. Wolfe.

Directions of Improvement in NTSC Color Television Systems*

DONALD RICHMAN†

Summary—This paper discusses possible directions of improvement in the NTSC color television standards. These are aimed at minor modifications of the transmitted signal which allow the use of simpler color receivers with more accurate resolution in colored areas. Compatibility also appears likely to be improved by one of the features presented, that is, monochrome receivers would display improved picture quality.

The modifications are specifically selected to permit color receivers designed for the unmodified standards to display satisfactory pictures, particularly including correct large-area color.

Since the system is essentially defined by the nominal form of receiver which reproduces the transmitted picture in accordance with the signal specification (the reference monitor), emphasis is on changes affecting the form of receiver.

Introduction

THE NTSC color television standards[1-4] provide for transmission of brightness information on a power law or gamma-corrected basis, consistent with the fact that the human eye tends to respond to fractional rather than incremental changes in brightness.[5,6]

In color television, there arises the problem of the form of encoding the added coloring information. This gamma-correction problem was not fully solved at the time the NTSC standards were formulated, and the FCC specifications of the standards were carefully worded to leave the door open for later developments.[7]

The NTSC color television standards include a form of gamma correction which falls short of ideal in that a major part of the luminance comes by the monochrome channel, while a minor part comes by the chrominance or color-difference channel. With a power exponent of $\gamma = 2$, these luminances are directly additive; the fraction which comes by the chrominance channel depends on both saturation and hue, increasing with saturation.

The total luminance, which is assumed to be the "resolution carrying" color coordinate,[8] comes partly by a wide-band monochrome channel and partly by a narrower band chrominance channel,[9-11] hence the present form of gamma correction inherently contains a "partial conflict of requirements between color and monochrome receivers" in the sense that resolution may not be simultaneously correct on both. The present standards are written to render monochrome detail as accurately as possible; this partial conflict of requirements cannot be removed without changing the encoding of large area colors.

It is clearly desirable to use the chrominance channel as efficiently as possible, particularly if it is to help carry luminance information. The chrominance channel consists of a double-sideband subcarrier with a pair of color-difference coordinates encoded on quadrature axes of modulation. A supplementary single sideband region of about twice the bandwidth of the double sideband region carries additional resolution in one of the coordinates. This represents an attempt to use the fact that for picture elements intermediate in size to those which require one or three *colorimetric coordinates*, there appears to be a range where *two colorimetric* coordinates may suffice.[12]

In addition to the restriction on luminance bandwidth,[13,14] certain basic properties of bandwidth-limited nonlinear transmission systems result in the occurrence of dark regions of low saturation, called notches, accompanying many color transients.[15,16]

The receivers which fully use the present signal have in essence a preferred set of demodulation axes. All three channels (wideband color-difference, narrowband color-difference, and monochrome) contain delay elements determined by the narrowest channel bandwidth; the penalty for fully widening the narrowest channel

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† Hazeltine Corp., Little Neck, N. Y.

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bandwidth is "quadrature crosstalk" including spurious luminance components; further, the nominal passband shape of the wide-band color-difference axis, providing single sideband boost, is not entirely practical for commercial receivers. It is not surprising that the bulk of receivers manufactured thus far are of a narrow-equal-band chrominance type which makes little use of the extra single sideband information.

The first and perhaps major direction of improvement involves encoding the single-sideband information more efficiently, so that it more accurately transmits *visible* information, with more careful regard to luminance; as auxiliary features we will camouflage the notches and shift the single sideband boost to the transmitter. The effect on the receiver is one of simplification; the nominal receiver has a "full-band equal-band" chrominance channel, having no preferred axes of demodulation, and with the extra feature that the burdensome chrominance pass-band-shape is removed from the receiver. By thus substantially increasing the performance difference between receivers having wide-band and those having narrow-band chrominance channels, while largely removing the difference between them in complexity, we might produce a substantial improvement in the *average* service rendered by the system.

The method for increasing the utility of the single-sideband chrominance components is called *Signal-Controlled Encoding*. The transient response of a bandwidth-limited, nonlinear color system depends on both the direction of the transient in (three-dimensional) signal space and the color through which the transient occurs. Thus, we may hope to improve the efficiency of the single-sideband chrominance channel if we let the composition of these transient components be varied, in accordance with the instantaneous color, so as to minimize the subjective error of the reproduced picture. For example, we might "transmit" a selected chrominance component near white, but transmit other components, such as the subcarrier-luminance in more highly colored areas.

In order to utilize a selected coordinate of resolution in any colored area, we *need* a receiver which, like the "full-band-equal-band" receiver, is capable of reproducing any preselected coordinate accurately; if we encode our signal efficiently, the inevitable spurious or missing components will be in the local natural visual directions of minimum visibility for all points in the chrominance plane; we might say these components are invisible, within the limitations imposed by the basic method of gamma correction.

The consequences of band-sharing further determine the properties of the nominal receiver. When two frequency-interleaved signals are decoded and displayed on receivers containing the conventional nonlinearities of gamma-restoration at the picture tubes and quadrature distortion at the second detectors, spurious *stationary patterns* are produced. These are visible on both color and monochrome receivers.

In principle, the quadrature distortion could be removed by advanced techniques, but for existing or anticipated monochrome or color receivers, it exists as a problem.

It appears possible to precorrect at the transmitter for the major spurious stationary byproducts of band-sharing as they affect the monochrome signal. This *luminance precorrection* would improve both monochrome and color receivers. But, since the rf/IF response is a factor in determining the amount of precorrection, the correction would be most effective for a selected form of receiver. There seems to be no way to dodge this problem; if we standardize a precorrection, we are defining an additional property of our nominal receiver: the effective rf link. But if we do not precorrect, we are also making a choice, and apparently not the most desirable one.

The directions of improvement discussed in this paper are intended to increase the resolution and compatibility of the present color television system by minor revisions of the transmitted signal which need not interfere with the continuity of the service. Best performance would be obtained on the simplest *form* of color receiver. Reception of color transmissions on monochrome receivers would also be improved.

The Problem

Having described what we hope to accomplish in this paper, we have the task of demonstrating that what has been stated is indeed so. To do this we will need to take a careful look at a good many of the concepts and ideas embodied in the NTSC standards and to pin down some facts by some simple mathematics. First of all, let us spell out the problem.

Luminance in the NTSC System

To begin with, let us find out how much of the picture luminance is carried by the monochrome and subcarrier (color-difference) signals.

The monochrome signal is represented as Y'. The color difference signals are $[R'-Y']$, $[G'-Y']$, and $[B'-Y']$; or, in general $[C'-Y']$. The *square* brackets indicate that these are bandwidth-limited signals. Then we can write the "transmitted" luminance at nominal gain as Y, where

$$\begin{aligned} Y &= 0.59(Y' + [G' - Y'])^\gamma + 0.30(Y' + [R' - Y'])^\gamma \\ &\quad + 0.11(Y' + [B' - Y'])^\gamma \\ &= \sum_{c=G,R,B} a_c(Y' + [C' - Y'])^\gamma. \end{aligned} \tag{1}$$

Within the scope of the NTSC standards we can set $\gamma = 2$ for simplicity.

Eq. (1) expresses the total reproduced luminance as the sum of the luminances produced by each of three colored light sources; each luminance component is proportional to the square of the sum of the monochrome and color difference signals.

The "constant luminance" constraint used in choosing the NTSC parameters is $\sum_c a_c[C'-Y']=0$. Using this, and $\sum_c a_c=1$, we get the following by expanding (1):

$$\begin{aligned} Y = {} & (0.59 + 0.30 + 0.11)Y'^2 \\ & + 2Y'(0.59[G' - Y'] + 0.30[R' - Y'] \\ & + 0.11[B' - Y']) \\ & + 0.59[G' - Y']^2 + 0.30[R' - Y']^2 \\ & + 0.11[B' - Y']^2 \end{aligned} \tag{2}$$

or

$$Y = Y'^2 + \Delta^2.$$

In (2)

$$\begin{aligned} \Delta^2 & = 0.59[G' - Y']^2 + 0.30[R' - Y']^2 \\ & \quad + 0.11[B' - Y']^2 \\ & = 0.456[I']^2 + 0.152[I'][Q'] \\ & \quad + 0.672[Q']^2. \end{aligned} \tag{3}$$

Thus the luminances of the subcarrier and monochrome signals are additive in a square law system.

For the present, we are postponing introducing effects resulting from band-sharing (frequency interleaving) and system nonlinearities.

On monochrome receivers, the luminance is Y_M where

$$Y_M = (Y')^2.$$

In order to be able to apply a "frequency response" approach to the resolution problem, and to make the physical effects easier to see, we can use a "small signal" representation; that is, let us differentiate (2):

$$dY = 2Y'dY' + 2[\Delta][d\Delta]. \tag{4}$$

(Note that the "gain" for the "mixed" highs carried by dY' is $2Y'$.) Thus, considering just the luminance, which is determined by the two parameters, Y' and the *luminance index*, Δ, we see that the transient response is determined by both the color through which the transient occurs, as defined by Y' and Δ, and by the direction of the transient in color space, as defined by dY' and $d\Delta$.

This principle can also be shown to apply to arbitrary transients in three-dimensional color space; in fact, it is a property of all nonlinear color systems. To see what this means, let us consider the transmission ratio for high-frequency components of a luminance transient for which no chrominance highs are transmitted; for the present we will study only the effect of restricting the subcarrier bandwidth relative to monochrome; additional effects due to difference between the bandwidths of I' and Q' will be treated later.

With this intent, Fig. 1 represents a luminance transient which has been separated into two parts; a low-

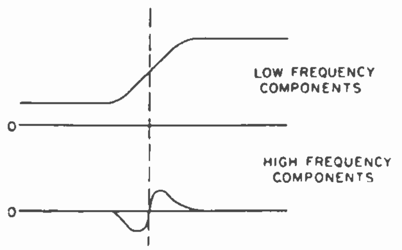


Fig. 1—Transient components.

frequency part which comes by both the Y' and Δ channels, and a high-frequency portion for which only the Y' portion will be transmitted. The relative transmission ratio for the highs is

$$R = \frac{dY_{\text{highs, transmitted}}}{dY_{\text{highs, in picture}}} = \frac{Y'dY'}{Y'dY' + \Delta d\Delta}. \tag{5}$$

Because Y', Δ, dY' and $d\Delta$ are all independent of each other and since the differential quantities can have either polarity, the highs can be too large, too small, present when they should be absent, absent when they should be present, or even reversed in polarity. In fact all of these things are possible in all parts of color space, so long as dY' and $d\Delta$ are independent of each other.

Consider for example a saturated-red to pastel (say gray) transition as it appears on color or monochrome receivers.

In the red region if $R=1$, $Y=0.30$ and $Y'=0.30$, whence $Y'^2=0.09$.

In the gray region, say $Y'^2=0.2$ whence $Y=0.2$ also, and $Y'=0.45$; thus going from red to gray, Y decreases from 0.3 to 0.2 while $(Y')^2$ *increases* from 0.09 to 0.2 and Y' increases from 0.30 to 0.45. Since the *highs* come entirely from the Y' signal, *the highs are reversed* in polarity because Y' increases while Y decreases. The resolution comes out all right on monochrome receivers because *both* large area brightness and resolution are derived from the Y' signal in monochrome receivers.

Thus, with the present method of gamma correction, since the monochrome brightness and colored brightness are derived from *different transmitted parameters*, (almost invariably) the highs cannot simultaneously be correct for both color and monochrome receivers. We may call this a "partial conflict of requirements" of the resolution components.

The luminance contributed by the chrominance subcarrier is Δ^2. It is useful to plot contours of constant Δ^2 in the chrominance plane. The contours are ellipses, with major and minor axes designated H' and P'; the axes are slightly in advance of I' and Q' as shown in Fig. 2. If H' and P' are also unit vectors, then

$$\Delta^2 = 0.432H'^2 + 0.696P'^2. \tag{3a}$$

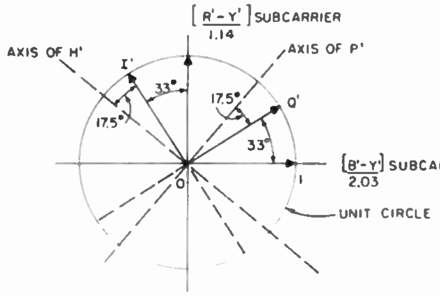
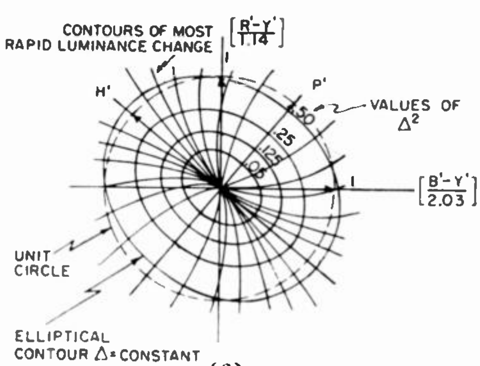
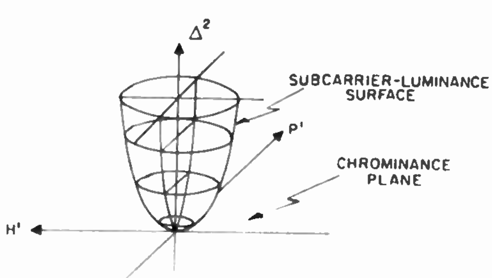


Fig. 2—Chrominance plane showing three sets of orthogonal coordinates.

Fig. 3(a) shows the elliptical contours in the chrominance plane, while Fig. 3(b) is a sketch of the subcarrier-luminance *surface* associated with the chrominance plane. The values of Δ^2 shown on the contours are for nominal operation of the receiver. If the saturation control is turned up 3 db, the luminance contributed by the subcarrier is of course doubled.



(a)



(b)

Fig. 3—(a) Luminance contours in chrominance plane. (b) Subcarrier-luminance surface of the chrominance plane.

Note that the variation in subcarrier luminance is *large* for *transients in the Q direction* near the green or magenta regions, and that luminance always changes fastest along contours which are defined by $P' \propto H'^{0.432/0.696}$; these contours are shown in Fig. 3(a).

We can use the elliptical contours of constant luminance in the chrominance plane to demonstrate the existence and nature of *notches*; that is, a darkening and desaturation accompanying transients in the subcarrier. From Fig. 3(b) we can see that to get directly *across* from one point to another on the subcarrier-luminance surface, we must "step into a hole."

Fig. 4 shows examples of chrominance transients which produce luminance notches. The straight line paths *ACE* represent the case of transients for which the orthogonal (I and Q) components have equal bandwidths and equal transition times, while paths *ABCDE* represent the case where the Q' bandwidth is narrow compared to the I' bandwidth, causing the Q' transient to start sooner and end later than the I' transient. Points A, C, and E are common. In each case the subcarrier luminance at the center point C, is much less than at A and E, although the transients are between points of equal subcarrier luminance. This will be shown later to be a consequence of the inherent loss of information due to restricting the bandwidth in a gamma-corrected signal. Note that the shortest path along the ellipse from A to E, which would remove the luminance notch completely, has higher saturation; also, such a transient would require a larger pass band.

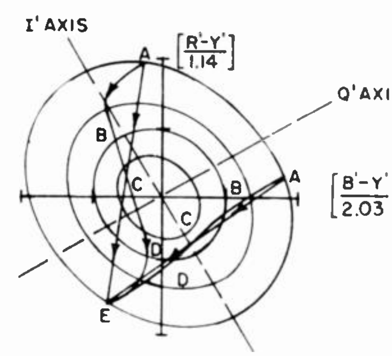


Fig. 4—Correctly-delayed transient paths in the chrominance plane.

One method for improving the resolution of color receivers, at the expense of monochrome receivers, would be to compose the "mixed" highs as follows:

$$dY'_{\text{modified}} = \frac{\text{required luminance transient} \equiv dY}{\text{gain for the mixed highs} \equiv 2Y'} \tag{6}$$

$$= dY' + \frac{\Delta}{Y'} d\Delta \cdot$$

By (5) this would correct the high frequency gain to unity. Of course, this would *invert* resolution *errors* from color receivers into monochrome receivers instead, because the signal which is added to correct color receivers then appears as a spurious component in monochrome receivers.

As a further step, the monochrome signal might be

$$Y'_{\text{modified}} = \sqrt{Y - [\Delta]^2}. \tag{7}$$

This would also invert the luminance notches into monochrome receivers, and would desaturate edges in color receivers because it results in adding white.

Obviously before attempting to choose a compromise which might improve the few existing color receivers at the expense of the many existing monochrome receivers, it is clearly worth investigating how we might increase the effectiveness and quantity of the information carried by the chrominance channel. We will see that there is another approach available which appears to be a desirable one.

Why I/Q?

There is considerable evidence that for intermediate components of resolution, color vision is largely a two-color process; this has been referred to as the "effective tritanopia" of the central fovea.

The narrow-band Q', wideband I' encoding of the subcarrier presently used was determined by finding experimentally that constant direction in the chrominance plane for a pair of perpendicular axes which, for a specific encoding of the subcarrier, appeared to make the resultant pictures most pleasing. The basic assumption was that one (rectangular) coordinate of the subcarrier is essentially invisible and hence not needed for the range of frequencies which are transmitted by single-sideband chrominance.

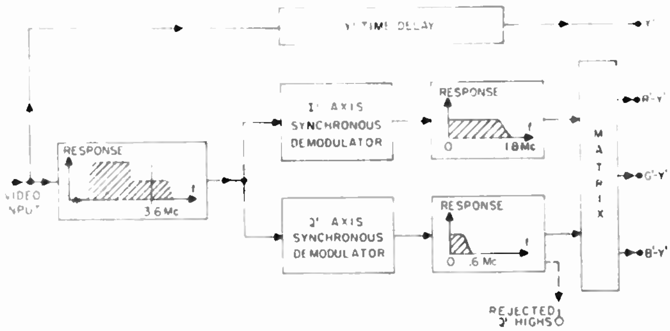


Fig. 5—A conventional NTSC receiver decoder, with narrow-band Q' and wideband I'.

There is direct evidence as to the validity of this assumption. Fig. 5 shows a block diagram for one conventional form of I/Q (receiver) decoder. The chrominance signal passes through a chrominance filter having a nominal gain of unity in the double sideband region extending 600 kc to each side of the subcarrier, and a nominal gain of two in the lower single sideband region extending approximately another 1200 kc. The chrominance signal is demodulated along the I' and Q' axes by synchronous detectors, and the outputs are filtered; the Q' channel filter is conventionally intended to reject the detection components above 600 kc, and the resultant signals are matrixed and combined with the appropriately delayed Y' signal.

This system is basically self-inconsistent. Here is why: If the spurious high-frequency Q' components such as appear in the rejected highs due to quadrature crosstalk were actually not visible, it would not be necessary to reject them. The narrow-band Q' filters used are for the purpose of rejecting these components. Since spurious high frequency Q' components *are* visible, comparable components of the actual picture would also be, thus there is a loss in information in rejecting these at the transmitter. The key to the problem is that the rejected highs, which come from quadrature crosstalk of the I' signal into the Q' channel in this single sideband system may produce *significant luminance.*

SIGNAL-CONTROLLED ENCODING

What we ought to do with these spurious potentially visible components is find a way to put them usefully to work. Let us examine how we might do this, with a view toward determining the system and receiver changes which are indicated, as well as the method of encoding at the transmitter. We can proceed on the assumption that the subcarrier luminance is the major factor contributing to the limitations just discussed of the present system. If at some future date we have available the results of a complete study of best choices for local wideband axes at all points in the chrominance plane, and as a function of the monochrome component, it is possible that additional minor changes of the *transmitter*, along the lines of what we develop here, might produce a small additional improvement.

Wideband Contours in the Chrominance Plane

For frequency components in the region of two-color vision, which are presumably those corresponding to the single-sideband frequencies of the chrominance subcarrier, we can divide the chrominance plane roughly into the three regions sketched in Fig. 6. In region I (saturated colors), the luminance contributed by the subcarrier is large and may be the dominant colorimetric parameter of the subcarrier. In region III, near white, the subcarrier luminance is negligible and some other colorimetric parameter, perhaps I', is dominant. In between is region II where some advantageous compromise must be found.

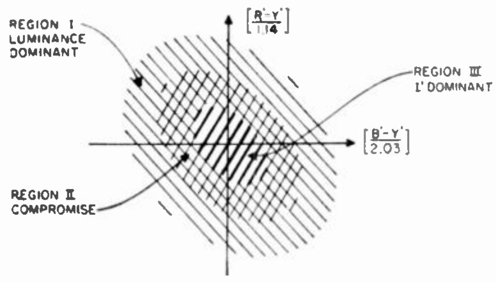


Fig. 6—Chrominance plane sketch showing regions where different parameters dominate.

Corresponding to these regions we can sketch the set of *wideband contours* in the chrominance plane shown in Fig. 7. At any point the contours have the direction of local wideband axes. Near white these are lines parallel to the I' axis; for saturated colors the contours are in the direction of most rapid luminance variation.

Part of our task is to find how to encode information of the required type within the available channel; but we must also make sure that when we transmit this information, the receiver will be able to display it!

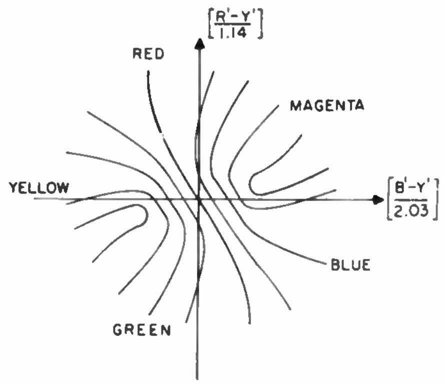


Fig. 7—Chrominance plane sketch indicating possible contours of the wide-band axis.

Luminance Sensitivity in the Receiver

Let us analyze the luminance sensitivity of the NTSC receiver to incremental luminance components. This can be applied then to the luminance produced by the single sideband components. In general, since Y is a function of Y', I', and Q', then

$$dY = \frac{\partial Y}{\partial Y'} dY' + \frac{\partial Y}{\partial I'} dI' + \frac{dY}{\partial Q'} dQ'. \tag{8}$$

For the luminance increments produced in a receiver by any possible I' and Q' components in the frequency range of the single sideband components of chrominance, therefore,

$$Y_{SSB} = \left[\frac{\partial Y}{\partial I'}\right] I_{SSB}' + \left[\frac{\partial Y}{\partial Q'}\right] Q_{SSB}'. \tag{9}$$

The transient response in a bandwidth-limited nonlinear color system is dependent on both the transient and the color through which it occurs. Here $[\partial Y/\partial I']$ and $[\partial Y/\partial Q']$ are determined by the color through which the transient occurs. By using the expressions for $[\Delta]^2$ given in (3) and differentiating as indicated by (8) we can find expressions for $[\partial Y/\partial I']$ and $[\partial Y/\partial Q']$ in terms of either I' and Q', or $[R'-Y'][G'-Y']$ and $[B'-Y']$.

In terms of I' and Q', we get

$$\begin{aligned} \left[\frac{\partial Y}{\partial I'}\right] &= \frac{\partial[\Delta]^2}{\partial I'} = 0.912\ [I'] + 0.152\ [Q'] \\ \left[\frac{\partial Y}{\partial Q'}\right] &= \frac{\partial[\Delta]^2}{\partial Q'} = 1.344\ [Q'] + 0.152\ [I']. \end{aligned} \tag{10}$$

The numerical values appearing in (10) permit us to plot, in the chrominance plane, the lines along which $[\partial Y/\partial I']$ and $[\partial Y/\partial Q']$ are zero. These are shown in Fig. 8. The line $[\partial Y/\partial Q']=0$ leads the I' axis by a small angle. The line $[\partial Y/\partial I']=0$ lags the Q' axis by a slightly larger angle.

Consider now what this implies about the narrow-band-wide-band receiver shown in Fig. 5, in which the *only* output produced by the single sideband components is in the I' channel. The line $[\partial Y/\partial I']=0$ corresponds to a blind axis for luminance for *this* receiver, in the sense that nothing which is transmitted in the single sideband can produce luminance when the color is on the $[\partial Y/\partial I']=0$ line; near this axis the luminance sensitivity is necessarily small.

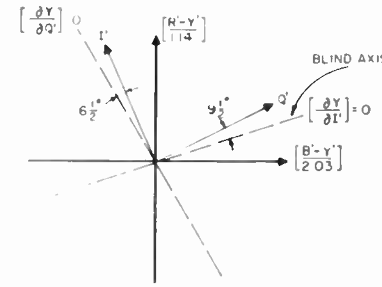


Fig. 8—Chrominance plane showing luminance sensitivity axes and the blind axis of the narrow-band/wide-band receiver.

The Simplified Receiver

One attractive form of receiver which has no blind axis for luminance is shown in Fig. 9. The pass bands are equal, wide and flat. This permits demodulation along any *axes;* matrixing may be simplified; Y' delay may be reduced. As an added feature, the single sideband chrominance gain at the receiver has been reduced to unity. Besides the obvious simplification, there are other advantages to this gain reduction which will appear in a later section of this paper.

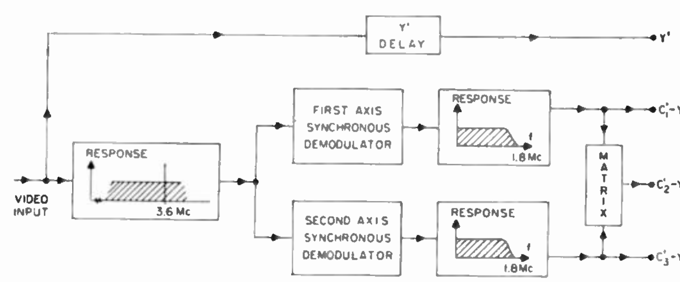


Fig. 9—Simplified receiver decoder.

The corresponding gain factor will be included in the computation of the required signal encoding.

Encoding the Single Sideband Signal

Luminance Correction: As the first step in finding how to encode the single sideband information we can develop the form of signal which would produce the correct luminance on the simplified receiver for transients through any points in the chrominance plane; later we can introduce the desired transition from region I to regions II and III of the plane.

Using the subscript s for those frequency components of the original picture for which the chrominance signal is transmitted by a single sideband, the luminance we wish to produce may be written as

$$Y_S = \left[\frac{\partial Y}{\partial I'}\right] I_S' + \left[\frac{\partial Y}{\partial Q'}\right] Q_S'. \tag{11}$$

The problem of course is that the single sideband channel can transmit only one coordinate, rather than the two-coordinates I_s' and Q_s'. We can conceive of our problem as one of precorrecting the I' signal for the quadrature luminance distortion of the single sideband signal in the simplified receiver, and of adding to the I' signal some (comparably precorrected) measure of the contribution to luminance of the Q' signal.

Before going through the necessary trigonometry, it may be helpful to get a physical picture of this "quadrature luminance distortion." The color-difference signals derived by the equal band receiver are those which represent demodulation along the three color-difference axes. These axes differ in phase from the arbitrarily selected axis of the encoded signal, say I', and hence the three outputs differ in phase from each other. The resulting luminance transient produced by the three square law color-sources then depends, in amplitude and phase, on how much of each color is present. Fig. 10 shows time-phase diagrams illustrating

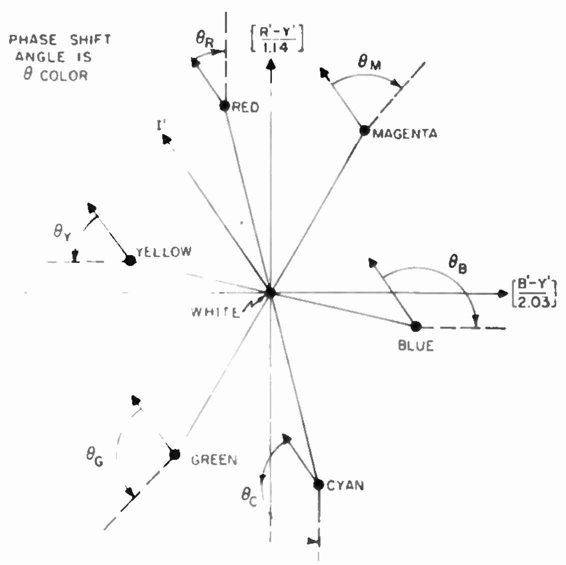


Fig. 10—Time-phase diagrams for transients through several colors in the chrominance plane indicating effective phase shifts for luminance transients transmitted by single sideband and displayed on simplified receiver.

this effect at a number of colors in the chrominance plane. At each color the I' direction is indicated as a zero of time phase, and the phase of the resultant luminance transient is shown as a broken line. The phase difference, which is clearly *predictable*, has an associated amplitude error which is equally predictable. The resultant is zero only for the white point. Fig. 11 indicates how this affects the reproduction of a transient. The upper curve shows the variation of a luminance transient with time, except for the luminance of the single sideband components, which are shown separately. For a *fixed single sideband component*, changing the *color* through which the transient occurs (from green to blue to red, etc.), will cause the single sideband *luminance* to

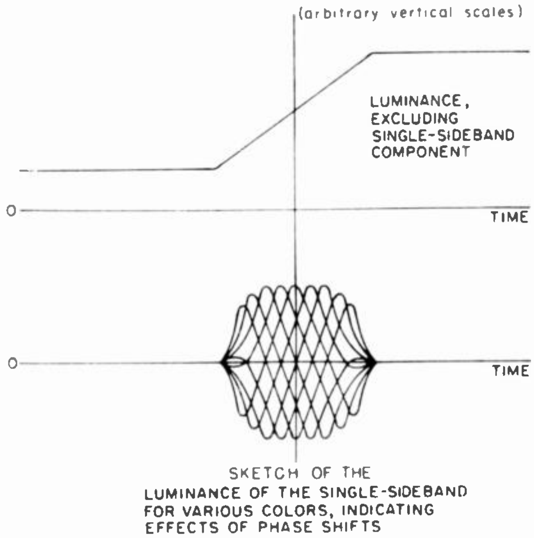


Fig. 11—Single-sideband transient transmission of one colorimetric coordinate.

occupy one or another of the positions in time, as roughly indicated in the figure.

In order to determine how to encode the signal we can compute the luminance sensitivity of this simplified receiver to the single sideband signal and compare this with the required luminance as expressed by (11). The desired expression can be written directly with the help of the following notation, which is useful for single sideband analysis because it permits operating on complete signals in compact form.

Since all of the signals are representable as Fourier series, just write

$$Z \quad \text{for} \quad \sum_K Z_K \cos\{\omega_K t + \phi_K\};$$

then, if all of the components of a signal are phase-shifted equally, by angle λ, write

$$Z(\lambda) \quad \text{for} \quad \sum_K Z_K \cos\{\omega_K t + \phi_K + \lambda\}.$$

Define the *quadrature function*, obtained by delaying all phases of a signal by 90°, with a special symbol; let $Z(-90°) \equiv Z^X$ and hence $(\cos\theta)^X = \sin\theta$. The usual trigonometric rules for products appear as follows in this notation:

$$Z\cos\theta = \tfrac{1}{2}\{Z(\theta) + Z(-\theta)\} \tag{12a}$$

$$Z\sin\theta = \tfrac{1}{2}\{Z^X(\theta) - Z^X(-\theta)\} \tag{12b}$$

$$Z^X\cos\theta = \tfrac{1}{2}\{Z^X(\theta) + Z^X(-\theta)\} \tag{12c}$$

$$Z^X\sin\theta = \tfrac{1}{2}(-Z(\theta) + Z(-\theta)\} \tag{12d}$$

$$Z(\theta) = \{\cos\theta Z - \sin\theta Z^X\} \tag{12e}$$

$$Z^X(\theta) = \{\cos\theta Z^X + \sin\theta Z\} \tag{12f}$$

$$Z(-\theta) = \{\cos\theta Z + \sin\theta Z^X\} \tag{12g}$$

$$Z^X(-\theta) = \{\cos\theta Z^X - \sin\theta Z\}. \tag{12h}$$

(Note that when $\theta = wt$, terms phase shifted by $(-\theta)$ are lower sidebands.)

No two of these are independent; the others are derivable from any one by taking quadrature functions, but all are listed for convenience.

Now, for simplicity, choose a phase scale, which differs from the FCC notation by 33°, and express the subcarrier signal in the general form,

$$I' \cos \omega_{sc}t + Q' \sin \omega_{sc}t. \tag{13}$$

Calling the precorrected I' and Q' signals I_{ps}' and Q_{ps}', we can apply (12a) and (12b) to (13) and write the *lower* single sideband signal as

$$E_{LSB} = \tfrac{1}{2}[I_{ps}'(-\omega_{sc}t) - Q_{ps}'^{X}(-\omega_{sc}t)]. \tag{14}$$

We see here the consequence of transmitting *only* a single coordinate; (the lower single sideband), the negative of the quadrature function of the Q' signal appears irretrievably added into the I' channel. But only the *sum* has real significance; therefore, the signal appearing at the output of the I' demodulators is

$$\tfrac{1}{2}[I_{ps}' - Q_{ps}'^{X}]. \tag{15}$$

The signal at the output of the Q' demodulators lags this by 90°, consistent with (13), hence it is its quadrature function, which is:

$$\tfrac{1}{2}[I_{ps}'(-90°) - Q_{ps}'^{X}(-90°)] = \tfrac{1}{2}[Q_{ps}' + I_{ps}'^{X}]. \tag{16}$$

The reproduced luminance is then

$$Y_{ps} = \frac{1}{2}\left[\frac{\partial Y}{\partial I'}\right][I_{ps}' - Q_{ps}'^{X}] + \frac{1}{2}\left[\frac{\partial Y}{\partial Q'}\right][Q_{ps}' + I_{ps}'^{X}] \tag{17a}$$

or, rearranging,

$$Y_{ps} = \frac{1}{2}\left\{\left[\frac{\partial Y}{\partial I'}\right]I_{ps}' + \left[\frac{\partial Y}{\partial Q'}\right]I_{ps}'^{X}\right\} + \frac{1}{2}\left\{\left[\frac{\partial Y}{\partial Q'}\right]Q_{ps}' - \left[\frac{\partial Y}{\partial I'}\right]Q_{ps}'^{X}\right\}. \tag{17b}$$

Now we have only to equate the desired luminance as expressed by (11) to the luminance obtained, as expressed by (17b). Remembering that only the single signal (15) is transmitted, we can arbitrarily equate corresponding terms of (11) and (17b) to get

$$\frac{1}{2}\left[\frac{\partial Y}{\partial I'}\right]I_{ps}' + \frac{1}{2}\left[\frac{\partial Y}{\partial Q'}\right]I_{ps}'^{X} = \left[\frac{\partial Y}{\partial I'}\right]I_s' \tag{18a}$$

$$\frac{1}{2}\left[\frac{\partial Y}{\partial Q'}\right]Q_{ps}' - \frac{1}{2}\left[\frac{\partial Y}{\partial I'}\right]Q_{ps}'^{X} = \left[\frac{\partial Y}{\partial Q'}\right]Q_s'. \tag{18b}$$

Now to solve, first multiply by

$$\frac{2}{\sqrt{\left[\frac{\partial Y}{\partial I'}\right]^2 + \left[\frac{\partial Y}{\partial Q'}\right]^2}}$$

and let

$$\left\{\begin{aligned} \frac{\left[\frac{\partial Y}{\partial I'}\right]}{\sqrt{\left[\frac{\partial Y}{\partial I'}\right]^2 + \left[\frac{\partial Y}{\partial Q'}\right]^2}} &= \cos\lambda \\ \frac{\left[\frac{\partial Y}{\partial Q'}\right]}{\sqrt{\left[\frac{\partial Y}{\partial I'}\right]^2 + \left[\frac{\partial Y}{\partial Q'}\right]^2}} &= \sin\lambda. \end{aligned}\right. \tag{19}$$

Then (18) becomes

$$I_{ps}' \cos\lambda + I_{ps}'^{X} \sin\lambda = 2\cos\lambda\, I_s' \tag{20a}$$

$$Q_{ps}' \sin\lambda - Q_{ps}'^{X} \cos\lambda = 2\sin\lambda\, Q_s'. \tag{20b}$$

We can apply (12g) to (20a) and (12h) to (20b) to get

$$I_{ps}'(-\lambda) = 2\cos\lambda I_s' \tag{21a}$$

$$-Q_{ps}'^{X}(-\lambda) = 2\sin\lambda Q_s'. \tag{21b}$$

Then, shifting phases by λ and $(\lambda - 90°)$, respectively, and then applying (12a) and (12d), respectively, we get these forms for the results:

$$I_{ps}' = 2\cos\lambda I_s'(\lambda) = I_s' + I_s'(2\lambda) \tag{22a}$$

$$Q_{ps}' = 2\sin\lambda Q_{ps}'^{X}(\lambda) = Q_s' - Q_s'(2\lambda). \tag{22b}$$

These are the effective I' and Q' signals which we have been looking for.

The single sideband can now be expressed in the following illuminating forms; since

$$E_{LSB} = \tfrac{1}{2}[I_{ps}'(-\omega_{sc}t) - Q_{ps}'^{X}(-\omega_{sc}t)] \tag{14}$$

we get

$$E_{LSB} = [\cos\lambda I_s'(\lambda - \omega_{sc}t) + \sin\lambda Q_s'(\lambda - \omega_{sc}t)] \tag{23a}$$

or

$$\begin{aligned} E_{LSB} &= \tfrac{1}{2}[I_s'(-\omega_{sc}t) - Q_s'^{X}(-\omega_{sc}t) \\ &\quad + I_s'(2\lambda - \omega_{sc}t) + Q_s'^{X}(2\lambda - \omega_{sc}t)] \\ &= I_s'(-\omega_{sc}t) - \tfrac{1}{2}[I_s'(-\omega_{sc}t) + Q_s'^{X}(-\omega_{sc}t)] \\ &\quad + \tfrac{1}{2}[I_s'(2\lambda - \omega_{sc}t) + Q_s'^{X}(2\lambda - \omega_{sc}t)]. \end{aligned} \tag{23b}$$

From (23a) we can see that the signal in the channel is automatically switched to come from I_s' or Q_s' as required, depending on the local color, since $\cos\lambda$ and $\sin\lambda$ are each zero when the other is maximum. Further, note from (23a) that the constant $\tfrac{1}{2}$ has disappeared; the gain of two has been transferred to the transmitter.

The particular form (23b) is useful when we wish to accomplish the transition from region I to regions II and II; here E_{LSB} is written as the original I' signal, including the gain factor of two, plus the added single sideband components necessary to correct for luminance.

We can accomplish the desired transition, for example, by making the added terms vary as a function, $m(\Delta)$, of the subcarrier luminance index, Δ. The opti-

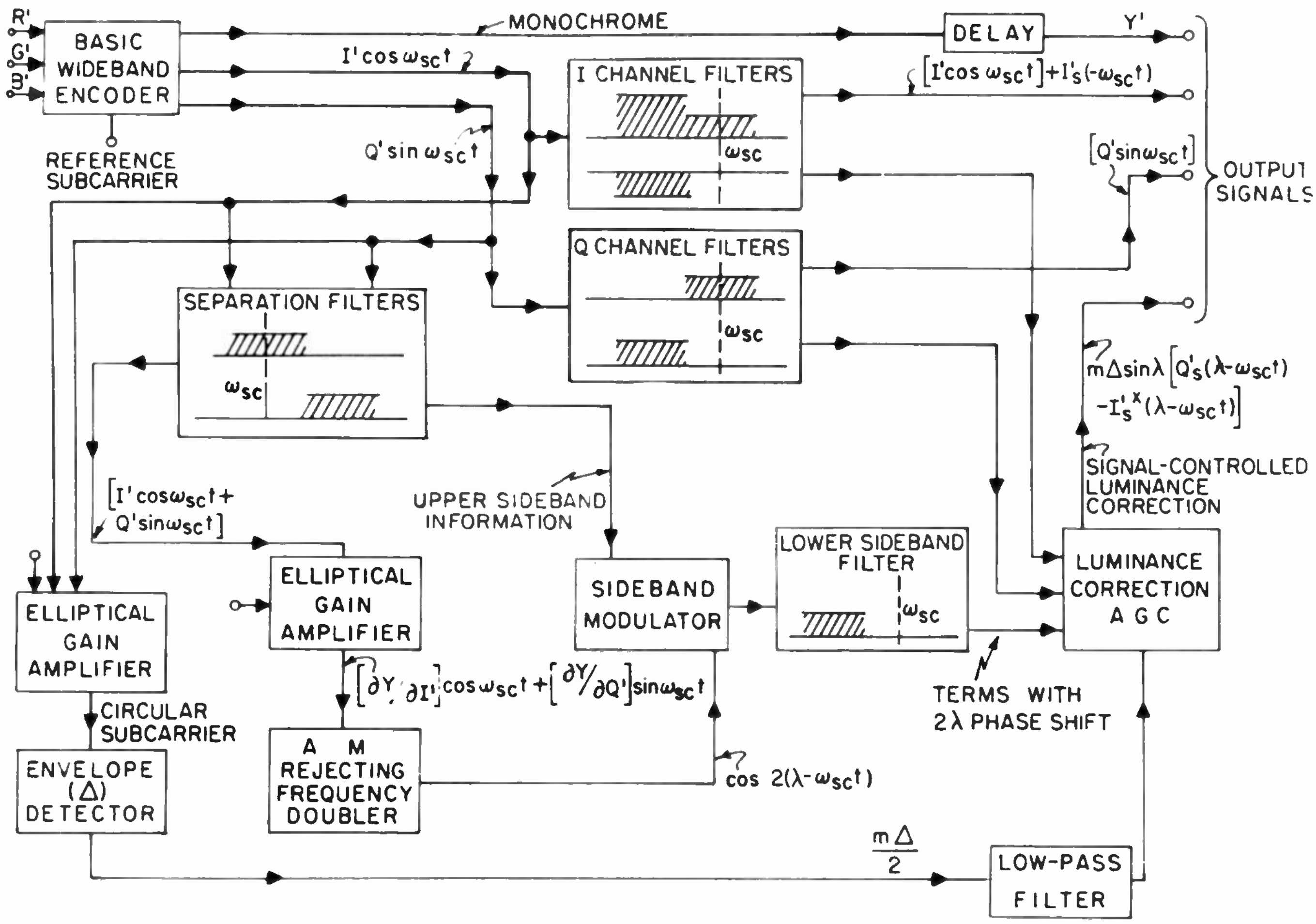


Fig. 12—Signal-controlled encoder.

mum form of $m(\Delta)$ can be determined experimentally; let us assume that a constant m times Δ is close enough to optimum so that the final form of the single sideband is

$$E_{LSB} = I_s'(-\omega_{sc}t) + \frac{m}{2}\Delta[-I_s'(-\omega_{sc}t) - Q_s'^X(-\omega_{sc}t) + I_s'(2\lambda - \omega_{sc}t) + Q_s'^X(2\lambda - \omega_{sc}t)] \quad (24a)$$

or

$$E_{LSB} = I_s'(-\omega_{sc}t) + m\Delta\sin\lambda[Q_s'(\lambda - \omega_{sc}t) - I_s'^X(\lambda - \omega_{sc}t)]. \quad (24b)$$

A Transmitter Encoder Block Diagram: There appear to be a rather large number of arrangements of the transmitter which can produce the desired signal. The block diagram shown in Fig. 12 is perhaps a good one; it develops the signal as the sum of the terms of (24a).

Fig. 12 shows a signal-controlled encoder which is intended to produce the single-sideband components described by either (24a) or (24b). Here is how it works.

1) The basic wideband encoder supplies Y', and wideband double sideband encoded I' and Q' signals.
2) Y' is delayed and supplied to the output.
3) I' is passed through an I channel filter and supplied to the output; an inverted lower sideband of I' is fed to the luminance correction agc for modulation by Δ.
4) Q' is filtered and supplied to the output; the lower sideband is also supplied for modulation by Δ.
5) I' and Q' are separated into narrow-band color and *upper* sideband signals in the separation filters.
6) The wideband double sideband color difference signal is converted into a circular subcarrier (radius proportional to Δ) in an elliptical gain amplifier which equalizes the luminance sensitivity on the H' and P' axes. The signal is then detected in an envelope detector and passed through a narrow-band filter. The filtered output is supplied as an agc signal to the luminance correction agc amplifier. (This filtering arrangement may be a desirable one because of the notch problem discussed in the next section).
7) Narrow-band color is supplied to an elliptical gain amplifier which converts it to a form having $[\partial Y/\partial I']$ and $[\partial Y/\partial Q']$, as expressed by (10) on quadrature axes. This is then operated on by

an a.m. rejecting frequency doubler which produces a constant amplitude second harmonic subcarrier with a phase shift of 2λ.

8) The upper sideband information is heterodyned against the phase-controlled second harmonic to produce a new *lower* single sideband containing the needed information from the I' and Q' channels.
9) The lower sideband is separated in a lower sideband filter and supplied to the luminance correction agc amplifier.
10) The luminance correction agc amplifier supplies to the output the extra single sideband components necessary to correct luminance in region III and to affect a compromise in region II of the chrominance plane.

Resume of Signal-Controlled Encoding

We can sum up the ideas and equipment changes involved in signal-controlled encoding as follows:

1) For each color as defined by a point in the chrominance plane, there is some local optimum direction of the wideband axis.
2) For saturated colors, luminance probably dominates in defining the optimum direction, with the presently practiced method of gamma correction.
3) The narrow-band/wide-band receiver decoder is not capable of displaying all of the information we would like, because it has a blind axis for subcarrier-luminance transients.
4) A simplified receiver decoder has acceptable properties, and no blind axis.
5) It is possible to encode the single sideband portion of the color-difference subcarrier signal so that it carries more information about visible components of the picture, by means of a signal-controlled coordinate transformation.
6) What we are trying to do is minimize the subjective error associated with any chrominance transient.
7) What we have done here is demonstrate the basic principles with an arrangement that probably gets the bulk of the available benefit. Modification of the encoding, and use of the monochrome components for added signal control are also possible.

Camouflaging the Notch

We want now to take a close look at the physical nature of systems using bandwidth-limited gamma-corrected signals, with the purpose of finding a way to reduce the visibility of some distortions inherent to such processes.

The distortions are chiefly visible as luminance notch components; we wish to precorrect these as far as possible by camouflaging what we cannot remove, that is, coloring it so it is less visible. As an aid, we have available the widened luminance bandwidth of the signal-controlled single sideband.

Bandwidth-Limited Gamma Correction for a Single Coordinate

Fig. 13 shows a bandwidth-limited gamma correction system for a single coordinate. A signal E is developed in the camera and is effectively band-limited by a transfer characteristic $N_C(f)$. It is then rooted to $E^{1/\gamma}$ and passed through the narrow-band filter $N_E(f)$. At the utilizer it is inversely modified to $[E^{1/\gamma}]^\gamma$ and filtered by $N_D(f)$. The eye introduces additional filtering. We wish to examine the effects of introducing the narrow-band filter $N_E(f)$ between the inverse nonlinear operations.

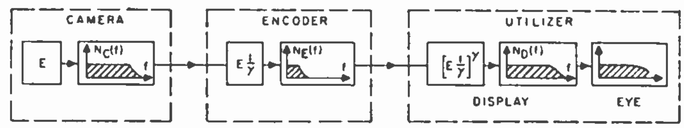


Fig. 13—Bandwidth-limited gamma-corrected channel illustrating information loss in nonlinear system.

The generation of notch components is rather simply demonstrated by considering the response to step waves which are rapid compared to the rise time corresponding to $N_E(f)$. If $e_s(t)$ has the shape of the symmetrical step wave response and $\sqrt{\bar{E}}$ represents a local average (the low frequency component) of $\sqrt{E}$, we can express the output of the rooter as follows, when $\gamma = 2$

$$\sqrt{E} = \sqrt{\bar{E}}\ (1 + e_s(t)); \tag{25}$$

the output of the squarer in the utilizer is then:

$$[\sqrt{E}\,]^2 = \sqrt{\bar{E}}^{\,2}\,(1 + 2e_s(t) + e_s^2(t)). \tag{26}$$

Fig. 14 shows the relevant waveforms.

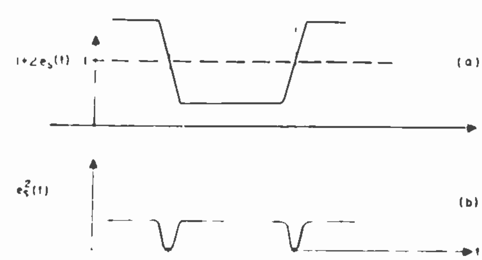


Fig. 14—Generation of notches in nonlinear process

Fig. 14(a) shows the linear components, while Fig. 14(b) shows the quadratic component, which includes a notch at each edge. In this kind of "single coordinate" notch, which occurs on transients between different levels, the effect is to narrow bright areas and broaden dark areas.

This effect occurs in a normal monochrome channel, but is usually visually not noticed because the several filter bandwidths (of Fig. 13) are comparable in that case. However, it also occurs and is significant, for radial transients in the chrominance plane; here the luminance notches are determined by a fairly narrow bandwidth. Widening the bandwidth for the subcarrier-luminance narrows the notches, and makes them less visible. In the case of somewhat slower transients, the notch components are also made shallower by widening the bandwidth.

Subcarrier Notches in Two-Dimensional Case

The subcarrier-luminance can also have notches between points of equal terminal value. This is the "stepping into a hole" effect mentioned earlier.

It appears to be possible to reduce the visibility of luminance notches in general by modifying the transient path, at the transmitter, for transients in the chrominance plane. One approach is presented below.

Fig. 15 shows a transverse transient path AA in the chrominance plane; transient AA goes between two points of equal terminal value of subcarrier luminance, but traverses a region of lower luminance, thereby producing a notch. Now let us represent the actual chrominance signal, E_{sc}, as an elliptical subcarrier modulated by the luminance index, Δ.

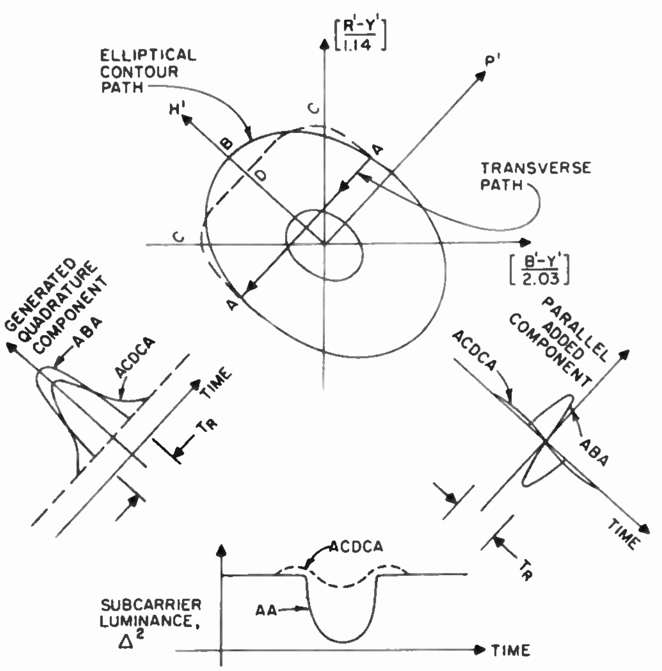


Fig. 15—Notch correction for a transverse chrominance transient.

Then

$$E_{sc} = \Delta \cdot \{\text{elliptical subcarrier}\}$$
$$= \Delta\{1 + k\cos(2\theta - 2\theta_H)\cos(\omega_{sc}t + \theta)\} \quad (27)$$

where θ is the phase angle of the subcarrier; θ_H is the angle of the axis of H'; and k is a constant determined by the difference between the major and minor axes of the ellipses.

The time variation of the luminance index, Δ, is modified by the selective circuits through which the signal passes; since superposition theory permits us to apply the selectivity to the component parts of the signal and add the results, let H' and P' be the components which are effectively band-limited to perhaps 1.8 mc so that

$$\Delta = \sqrt{0.432[H']^2 + 0.696[P']^2}. \quad (28)$$

Since the (rectangular) components H' and P' of any transition proceed at the same rates, the straight line paths result. This method of applying the filtering results in transient paths such as AA which have luminance notches.

The correct value for Δ may be determined from the relatively wideband luminance and monochrome signals; call this Δ_Y. Then

$$\Delta_Y = \sqrt{Y - Y'^2} \quad (29)$$

where of course

$$Y = 0.59G + 0.30R + 0.11B$$

and

$$Y' = 0.59\sqrt{G} + 0.30\sqrt{R} + 0.11\sqrt{B}. \quad (30)$$

If we could modulate the gain of the subcarrier by

$$\rho = \frac{\Delta_Y}{\Delta} = \frac{\sqrt{Y - Y'^2}}{\sqrt{0.432[H']^2 + 0.696[P']^2}} \quad (31)$$

the subcarrier signal would become

$$\rho E_{sc} = \frac{\Delta_Y}{\Delta} \cdot \Delta \cdot \{\text{elliptical subcarrier}\}$$
$$= \Delta_Y \cdot \{\text{elliptical subcarrier}\}. \quad (32)$$

The signal described by (32) has the correct luminance; it does not have the long-duration notch of path AA; ideally it may follow the elliptical contour ABA, which requires a wider bandwidth to transmit it in the same *time* as transient AA. But if we pass the modified subcarrier through a subsequent phase equalized filter of comparable bandwidth (± 1.8 mc) the transient $ACDCA$ results; this transient is spread slightly in time and is essentially free of a luminance notch.

The shape of transient $ACDCA$ is found as follows: the components added to transient AA by converting to transient ABA may be resolved into a convenient pair of quadrature components; one parallel to AA and one in quadrature with AA; these are indicated in Fig. 15, along with the waveforms, before filtering (ABA) and after filtering ($ACDCA$). The filter rise time is T_R. The parallel added component is rapid and has odd symmetry; it is removed by the selectivity. The *quadrature* component generated is broadened by the selectivity; its average value is not zero. The transition CC takes about as long as the original transition AA.

The resulting luminance for path $ACDCA$ as compared to AA is shown in an auxiliary sketch in Fig. 15. The shapes are typical for the ***notch components*** for almost all transients in the chrominance plane; transverse, skew, or radial. The unique exception is symmetrical transients going exactly through the origin.

Subsequent single sideband filtering in a *signal-controlled encoder* will transmit the luminance component *essentially correctly*.

Saturation of Subcarrier Notches

Saturation effects associated with subcarrier notches may be worthy of note. It can be seen from Fig. 15 that the straight line AA has less *saturation* in the middle than the other paths. The corrected subcarrier path restores this lost saturation along with the luminance.

If the missing luminance were supplied by making the monochrome signal momentarily stronger during this transient as for example by the method of (7), there would instead be a further loss of saturation in the region.

Instrumentation

A block diagram for the notch corrector is shown in Fig. 16. It uses signals available from a basic encoder, and is connected between the basic encoder and the signal-controlled encoder. Here is how it works:

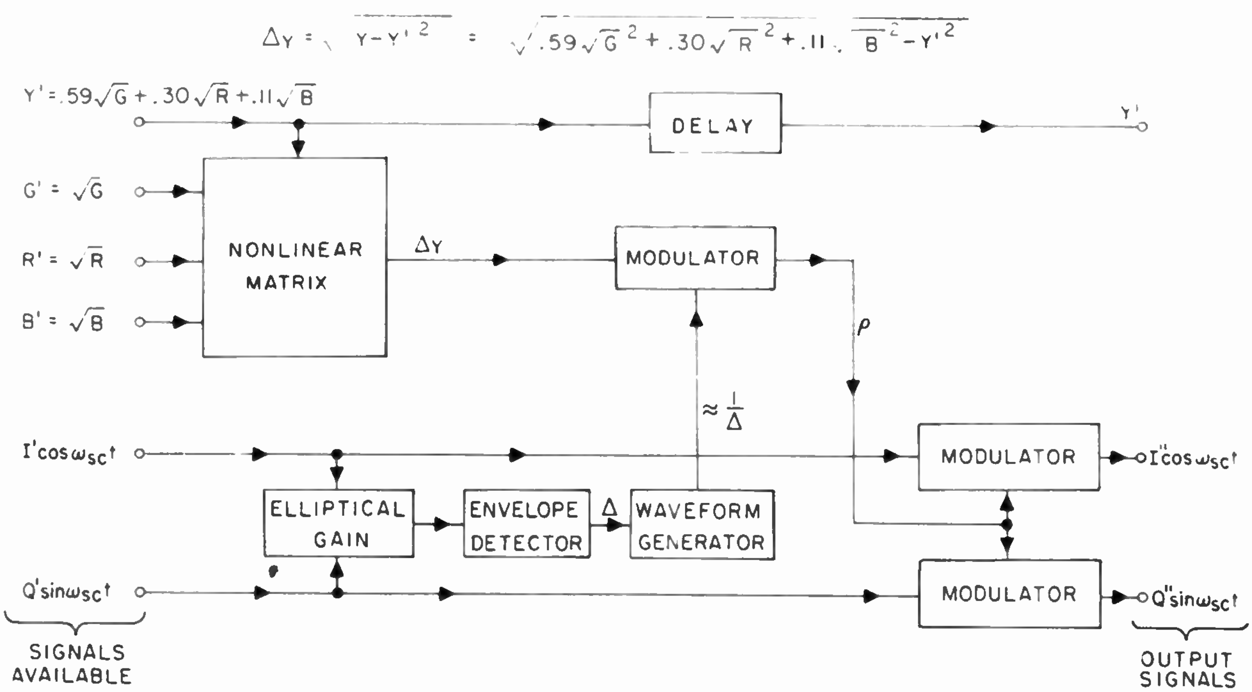


Fig. 16—Notch corrector.

1) A nonlinear matrix generates the signal Δ_Y from R', G', B', and Y', by generating the signals of (29) and (30).
2) The signal representing Δ for the subcarrier is generated by an elliptical gain amplifier, and an envelope detector.
3) The waveform generator develops from Δ a waveform approximating $1/\Delta$.
4) It may be desirable that the developed signals Δ and Δ_Y be constrained (by limiting) not to approach closer to zero than some selected level.
5) The signal $\rho=\Delta_Y/\Delta$ is generated in a modulator.
6) The subcarrier components may then be modulated by ρ.

As an additional feature, a signal derived from Δ_Y by band-limiting may be used as the agc control signal in the signal-controlled encoder.

Nonlinear Frequency Interleaving

The separate signals carrying monochrome and color-difference information must be interleaved in frequency in order to fit them into the available channel space. This frequency interleaving with a separation of odd harmonics of half line frequency causes the linear cross-talk components to alternate in time and have low visibility.

However, there are some inherent nonlinear processes associated with using the signals which introduce stationary, or nonalternating patterns. We want now to examine these; first, to get a physical picture of what happens, and also to find how to reduce some of the resultant errors by precorrection at the transmitter; of course, we will keep an eye open for possibilities which make the simplest form of receivers (particularly IF/rf response) more practical. And we also want to take special note of the effects on monochrome receivers of the presence of subcarrier components.

Effects Due to Use of a Square Law Monochrome Display

It may be informative to consider first only the effects due to use of square law displays, and postpone evaluating the complication of these effects by quadrature distortion at the receiver second detector.

The frequency bands for the monochrome and color signals are of course modified by the receiver selectivity. The specific form of selectivity will be included in the analysis of quadrature distortion; for the present, the following general nomenclature is adequate:

M = monochrome signal in monochrome channel
$\tilde{C}_M$ = color signal in monochrome channel
C = color component in a color channel (after synchronous detection)
$\tilde{M}_c$ = monochrome component in a color channel (after synchronous detection).

The superscript ~ is used to denote odd harmonic components. The luminance in a monochrome receiver is then

$$Y_M = (M + \tilde{C}_M)^2 = M^2 + 2M\tilde{C}_M + \tilde{C}_M^2. \quad (33)$$

The term M^2 represents normal monochrome reproduction. The term $2M\tilde{C}_M$ is presumably a low visibility term. The term $\tilde{C}_M^2$ consists of ***even*** harmonics and hence visible terms; half of these are low-frequency terms. They represent spurious crosstalk of color into monochrome. In an average receiver which might have (roughly) full response to 3 mc and be 6 db down at 3.6 mc a significant portion of $\tilde{C}_M^2$ may be contributed by the single sideband color.

We might hope to improve monochrome resolution by including a correction derived from the *stationary* crosstalk components. For example, in this ideal case resolution could be improved by transmitting as much of

$$M = \sqrt{Y'^2 - \tilde{C}_M^2} \quad (34)$$

as bandwidth and depth of modulation will allow. When we include quadrature distortion at the second detector, the mathematics is considerably more cumbersome but this principle states the objective; the sum total of the *stationary* terms should produce Y'^2, by suitable precorrection of the transmitted monochrome signal.

Effects Due to Use of a Square Law Color Display

The reproduced luminance in a color receiver may be found by incorporating the crosstalk terms into (1) of this paper. Then we get:

$$\begin{aligned} Y_{\text{color}} &= \sum_c a_c(M + \tilde{C}_M + [C' - Y' + \tilde{M}_c])^2 \\ &= M^2 + 2M\tilde{C}_M + \tilde{C}_M^2 \text{ (monochrome)} \\ &\quad + 2\left\{\sum_c a_c[C' - Y' + \tilde{M}_c]\right\}(M + \tilde{C}_M) \\ &\qquad \text{(zero luminance color)} \\ &\quad + \sum_c a_c[C' - Y' + \tilde{M}_c]^2 \text{ (square law color)}. \quad (35) \end{aligned}$$

The terms contributing may be divided into monochrome components, zero luminance color, and square law color terms. These have been further separated into desired terms, low visibility interleaved terms, and spurious visible terms, and tabulated in Table I.

TABLE I
BASIC FORMS OF TERMS OBTAINED FROM FREQUENCY-INTERLEAVED SIGNALS AND SQUARE LAW COLOR DISPLAY

	Desired [Same as (1)]	Low Visibility Interleaved Terms	Spurious Visible
Monochrome	M^2	$2M\tilde{C}_M$	$\tilde{C}_M^2$
Zero Luminance Color	$2M\sum_c a_c[C'-Y']$	$\begin{cases} 2M\sum_c a_c\tilde{M}_c \\ 2\tilde{C}_M\sum_c a_c[C'-Y'] \end{cases}$	$2\tilde{C}_M\sum_c a_c\tilde{M}_c$
Square Law Color	Δ^2	$2\sum_c a_c[C'-Y']\tilde{M}_c$	$\sum_c a_c\tilde{M}_c^2$

The spurious terms are:

1) A term $\tilde{C}_M^2$; this appears on monochrome receivers and should be included in the monochrome luminance precorrection.
2) A term $2\tilde{C}_M\sum_c a_c\tilde{M}_c$; this has no *luminance* but can cause visible low frequency coloring near sharp edges.
3) A term $\sum_c a_c\tilde{M}_c^2$; this is the luminance component of what is normally called *cross color*.

It does not appear reasonable to correct this last term by modification of the monochrome signal; furthermore, it is not possible to remove it by modifying the chrominance signal as this term often occurs in gray areas where there is no normal subcarrier luminance.

We can see here the added benefits obtained by reducing the receiver gain to the single-sideband components by 2 to 1 and raising the transmitter gain instead, as previously mentioned:

1) Most of the components of $\sum_c a_c\tilde{M}_c^2$ come from the single-sideband region; the effect of these is reduced 4 to 1.
2) The average effect of $2\tilde{C}_M\sum_c a_c\tilde{M}_c$ appears likely to be made less significant.
3) The small increase of $\tilde{C}_M^2$ is obliterated by removing this term with monochrome luminance precorrection.

Therefore it appears that the combination of adjusted single-sideband gain and luminance precorrection may increase the net benefit to be obtained from use of the single-sideband in the color system; both monochrome and color receivers would demonstrate improved picture quality.

Effect Due to Quadrature Distortion at the Second Detector

As the next step we can investigate the physical nature of quadrature distortion at the video detector. Its existence is well-known; the expressions "loss of resolution in bright areas," and "subcarrier rectification" both refer to it; it is a significant source of the 920-kilocycle beatnote between color and sound carriers which so drastically affects receiver design. Now let us analyze it.

The signal at the second detector of a television receiver consists of two video portions: the double-sideband video components and the single-sideband video components, including the frequency-interleaved color-difference information.

For our present purposes, it is convenient to represent the double-sideband components in a simplified symmetrical form which is most accurate for the lowest frequency components. This is $(1-u_0)$ cos $_\omega pt$; here u_0 is proportional to the low-frequency monochrome component, with the picture carrier amplitude at blanking normalized to unity.

The boosted video single sideband appearing at the second detector is then

$$-\sum_K u_K Z_K \cos(\omega_p t - \omega_K t - \phi_K); \tag{36}$$

the component amplitude is u_K; the selectivity applies the gain constant Z_K. The signal to be detected is

$$E = (1 - u_0)\cos\omega_p t - \sum_K u_K Z_K \cos(\omega_p t - \omega_K t - \phi_K). \tag{37}$$

The output of the second detector is

$$\left| (1 - u_0)\cos\omega_p t - \left\{ \sum_K u_K Z_K \cos(\omega_K t + \phi_K)\cos\omega_p t + \sum_K u_K Z_K \sin(\omega_K t + \phi_K)\sin\omega_p t \right\} \right|_{\text{average peak}}. \tag{38}$$

The terms multiplying cos $w_p t$ represent the desired modulation while the term multiplying sin $w_p t$ is a spurious quadrature term. Now write

$$u = \sum_K u_K Z_K \cos(\omega_K t + \phi_K)$$
$$u_X = \sum_K u_K Z_K \sin(\omega_K t + \phi_K). \tag{39}$$

The detected voltage may then be written as:

$$[E_D] = [\sqrt{(\text{coefficient of } \cos\omega_p t)^2 + (\text{coefficient of } \sin\omega_p t)^2}]$$
$$= [\sqrt{(1-u_0-u)^2+(u^X)^2}]$$
$$= [\sqrt{1+\{(u_0+u)^2+(u^X)^2-2u_0+u)\}}]$$
$$= [\sqrt{1+\{x\}}]. \tag{40}$$

Here the square bracket represents limitation of bandwidth due to selectivity in the *video* circuits.

By using the usual series for $\sqrt{1+x}$, which is

$$\sqrt{1+x} = 1 + \frac{x}{2} - \frac{x^2}{8} + \frac{x^3}{16} - \frac{5}{108}x^4 + \frac{35}{1080}x^5 - \cdots \tag{41}$$

we can see that the *linear* (lowest order term) is in the expected form since

$$[E_D] = 1 - u_0 - u + \text{higher powers}. \tag{42}$$

The distortions which concern us here are the nonlinear effects which appear in the terms higher than the first power in a series expansion for $[E_D]$. There are two effects which are certainly worthy of mention: 1) By writing out the odd power terms and collecting those terms containing components which would pass through the chrominance pass band, we could demonstrate a reduction in saturation and chrominance detail for highly colored areas. 2) We can determine the displayed brightness, to more accurately determine a desirable form of luminance precorrection. This second case will be carried through in detail here.

Correction of the Monochrome Component

The monochrome brightness should be Y'^2 with the present form of gamma correction. Thus, if the monochrome signal u_0+u were encoded properly, we would have

$$Y'^2 \propto (1 - [E_D])^2. \tag{43}$$

The form of (43) results from our normalization of black level to unity since black is reinserted by the display bias.

If we expand (43) and obtain a power series up to the second power in u, u_0, or u^X, we can get, after some manipulation, this form, in which the square brackets now represent the bandwidth limitation of the video circuits:

$$Y'^2 \propto [u_0 + u]^2 - [(u^X)^2(u_0+u)][u_0+u] - [(u^X)^2][u_0+u] + \text{higher terms}. \tag{44}$$

The first term represents ideal square law reproduction; the second term is a quadrature distortion of the ideal term; the third term is a first-power term coupled in by quadrature distortion.

To put these in terms of M and $\tilde{C}_M$, we note that from the FCC color television standards,

$$\frac{Y' \text{ maximum}}{(u_0+u) \text{ maximum}} = \frac{\text{Blanking to White Range}}{\text{Blanking Level}} = \frac{62.5}{75} = \frac{5}{6}. \tag{45}$$

Now, if $u = \underline{u} + \tilde{u}$ where $\underline{u}$ represents video resolution and $\tilde{u}$ represents the odd harmonic components in the color subcarrier, then

$$u_0 + \underline{u} + \tilde{u} = \frac{5}{6}[M + \tilde{C}_M] = \frac{5}{6}[M_0 + \underline{M} + \tilde{C}_M]$$

where

$$M_0 = \frac{6}{5} u_0 \quad \text{and} \quad \underline{M} = \frac{6}{5} \underline{u}. \tag{46}$$

approximation, and one which may give some insight into the nature of the correction is obtained this way: use Y' for M in the correction terms. Ignore the difference between terms like $[M]^2$ and $[M^2]$, then

$$M \approx \sqrt{\left\{[Y']^2\left[1 + \frac{25}{36} q^2\right] - [\tilde{C}_M]^2\left[1 - \frac{25}{36} q^2\right]\right\} + \left\{\left[Y'\right]\left[\frac{5}{6} q^2\right]\right\}}. \tag{49}$$

We can therefore rewrite (44) to get:

$$\begin{aligned} Y'^2 \approx\ & [M + \tilde{C}_M]^2 \\ & - \frac{25}{36}[(\underline{M}^X + \tilde{C}_M{}^X)^2(M + C_M)][M + \tilde{C}_M] \\ & - \frac{5}{6}[(\underline{M}^X + C_M{}^X)^2][M + C_M]. \end{aligned} \tag{47}$$

Now expanding and retaining only the stationary terms, noting that the product of a function with its quadrature function produces no dc term, and letting $M^{X2} + \tilde{C}_M{}^{X2} \equiv q^2$ we get

$$\begin{aligned} Y'^2 \approx\ & [M^2 + C_M{}^2] - \frac{25}{36}[M][Mq^2] \\ & - \frac{25}{36}[C_M][\tilde{C}_M q^2] - \frac{5}{6}[M][q^2]. \end{aligned} \tag{48}$$

Eq. (48) is the final form of equation; the transmitter design problem may now be stated this way:

> The signal M to be transmitted in the monochrome channel should be designed to produce the correct value of $(Y')^2$ on a reference monitor having a specified rf/IF response, second detector, and luminance channel bandwidth.

Satisfying (48) gives the first order correction of Y'. Let us now solve this in one simple approximation to find the nature of the correction. Perhaps the simplest

The first term represents a correction of the simple form expressed earlier by (34). The second term accounts for the added first-power term which results when quadrature distortion precedes application of the signal to a square law display. When the monochrome component satisfies the conditions imposed here, the *net* monochrome component of luminance produced on receivers approximating the specified selectivity will be (nearly) Y'^2.

The usual solution of the quadratic expression (48) may also be a useful form.

It has therefore been demonstrated that a simple form of signal exists by means of which we can precorrect (at least to first order) for effects of conventional receiver nonlinearities on displayed luminance.

Conclusion

The analysis and suggestions embodied in this paper have shown that it appears to be possible to improve the resolution and compatibility of color television signals of the NTSC type, and to produce these improved color pictures on simplified color receivers. It is hoped that the concepts presented here may be of use in development of improved color television standards.

Acknowledgment

The encouragement and support of A. V. Loughren and C. J. Hirsch is gratefully acknowledged.

A Precise New System of FM Radar*

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Summary—**The paper describes a new system of fm radar by which both the range and the speed of the target can be accurately measured. The "fixed error," which is a characteristic of fm radar apparatus now in use, has been eliminated, affording correct measurement of ranges, even if they are very short. The functioning of the new system has been investigated experimentally and proven to be very successful. Consequently, the author in cooperation with the Hasler Laboratories in Bern, Switzerland, developed a low-range airplane altimeter patterned after the new system.**

INTRODUCTION

IN THE SYSTEM of fm radar now in use,[1] the way of measuring the range and speed of a target, relative to the radar transmitter, may be briefly described as follows.[2] The radar transmitter radiates in the direction of the target an fm wave such as that shown in Fig. 1(a). The transmitted radio frequency

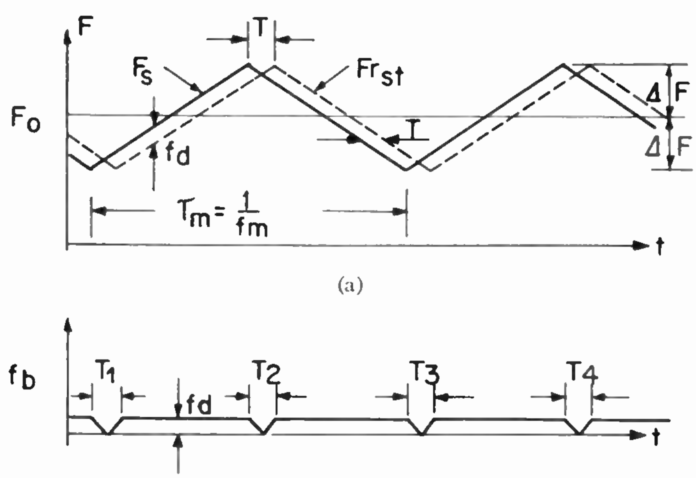

Fig. 1—Frequencies present in a stationary target.

F_s has a mean frequency F_0 and changes periodically at a constant lf f_m in a symmetrical sawtooth manner between maximum and minimum frequencies of $F_0+\Delta F$ and $F_0-\Delta F$, respectively. Some transmitted energy will be reflected at the target's surface and will return to the transmission point after a delay time T so that

$$T = \frac{2d}{c} \tag{1}$$

where d = distance between target and radar apparatus and c = light velocity in free space. This returning signal represents the received signal in the case of a stationary target, and its frequency $F_{r_{st}}$ may be drawn to the same time base of F_s simply by shifting F_s to the right by the time delay T, as shown in Fig. 1(a).

In the radar receiver, a local signal at frequency F_s is mixed with the received signal F_r, giving an output signal at the beat frequency $f_d = F_s - F_r$. As shown in Figs. 1(a) and 1(b), this beat frequency is constant except in small time intervals $T_1, T_2, T_3, \cdots = T$. Generally, $T \ll \tau_m = 1/f_m$, so that the mean value of the beat frequency can be considered equal to the constant value f_d. From (1) and Fig. 1 it follows that

$$f_d = \frac{8d}{c} \cdot \Delta F \cdot f_m. \tag{2}$$

Hence, according to (2), if f_d is measured, d will be directly indicated. This is done by amplifying the beat note signal to a usable level, and applying it to an averaging frequency counter, indicating its average frequency which is a direct measure of the range.

A complete study of the waveform of the beat note signal shows that its average frequency f_d must always be an exact multiple of f_m. Therefore it is apparent that d can accurately be measured only when $8d\Delta F/c$ is a whole number. At other values of d, however, the indicated f_d may correspond either to a higher or to a lower value of d which makes $8d\Delta F/c$ a whole number.

The difference between any two successive values of ranges d_n and d_{n+1} which makes $8d\Delta F/c = n$ and $n+1$ respectively produces a confusing error at all ranges lying between them and is always constant. This error is called the "fixed error" and its value can easily be obtained by substituting f_m for f_d in (2), giving

$$d_{\text{error}} = \frac{C}{8\Delta F}. \tag{3}$$

From (3) it is seen that in order to decrease the fixed error, ΔF should be increased. In the working apparatus, several mc ΔF were used, but the fixed error remained of considerable value.[3]

For a constant ΔF, as d_{error} remains constant at all ranges it produces a relative error which increases with the decrease of the range and puts an end to the use of the usual fm radar to determine accurately ranges which are relatively short.

According to the Doppler effect, if the target is approaching at a speed v relative to the radar apparatus, the received frequency at any instant will increase above that received if the target is stationary by the speed frequency

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[1] For simplicity, this will be called "the usual system."
[2] For its complete study, refer to the literature.

[3] $d_{\text{error}} = 6$ feet for $\Delta F = 20$ mc.

$$f_v = F_0\left[\frac{c+v}{c-v} - 1\right]$$

and because always $c \gg v$, this is usually approximated to the form

$$f_v = F_0 \cdot \frac{2v}{c}. \qquad (4)$$

The received frequency from a moving target $f_{r_{mov}}$ can therefore be drawn on the same time base of F_s simply by shifting $F_{r_{st}}$ upwards by the speed frequency f_v [Fig. 2(a)]. The beat frequency f_b resulting from the mixing of F_s and $F_{r_{mov}}$ is shown in Fig. 2(b). This does

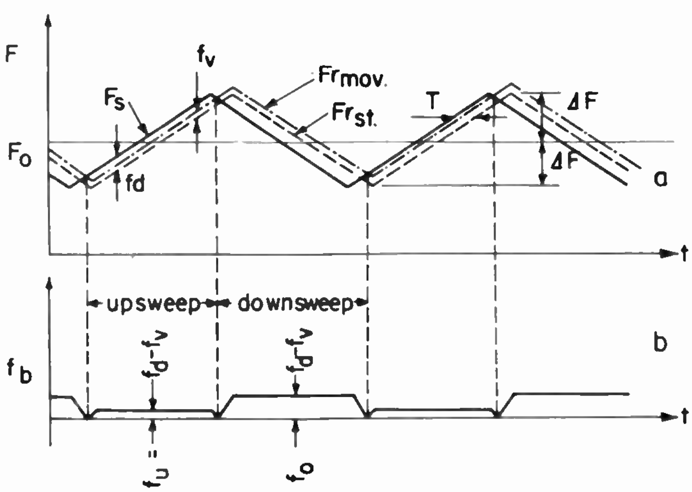

Fig. 2—Frequencies present in a moving target having $f_d > f_v$.

not remain constant during the whole modulation period as in the case of a stationary target, but changes every half a modulation period. If $f_d > f_v$, the resulting upsweep frequency f_u and the downsweep frequency f_0 will be

$$\left.\begin{aligned} f_u &= f_d - f_v \\ f_0 &= f_d + f_v \end{aligned}\right\} \qquad (5)$$

giving

$$f_d = \left(\frac{f_u + f_0}{2}\right)$$

and (6)

$$f_v = \left(\frac{f_0 - f_u}{2}\right)$$

and when

$$f_u = 0,$$

$$f_d = f_v = \frac{f_0}{2}.$$

If it happens that $f_d < f_v$, the result is

$$\begin{aligned} f_u &= f_v - f_d \\ f_0 &= f_v + f_d \end{aligned} \qquad (7)$$

giving

$$f_d = \left(\frac{f_u + f_0}{2}\right)$$

and (8)

$$f_v = \left(\frac{f_0 - f_u}{2}\right).$$

From (5)–(8) it is seen that for the separate indications of f_d and f_v, an averaging and a switched frequency counter are essential. The first measures the average beat note frequency, f_d, when $f_d > f_v$, or f_v, when $f_v > f_d$, while the other measures half the difference between f_0 and f_u, f_v, when $fd > f_v$, or f_d, when $f_d < f_v$.

In practice, fm radar apparatus is designed so that f_d remains greater than f_v in almost all working circumstances, and hence the averaging and switched frequency counters can be directly calibrated to indicate the target's range and speed respectively. However, for relatively near and speedy targets, f_v may exceed f_d, causing false range and speed readings to be indicated.

The main weak points of the usual fm radar system may be summarized as follows.

1) The existence of the fixed error causes the accuracy in measuring the range to deteriorate very quickly as the range decreases.

2) False range and speed data will probably be indicated for near and speedy targets.

3) Large frequency deviations must be produced in order to decrease the fixed error.

4) A symmetrical saw-tooth form of fm must be used when both the range and the speed of the target are to be measured.

In the new system, however, only simple sine wave fm and relatively small frequency deviations are essential. Furthermore, no fixed error exists, and the target's range and speed can always be very sensitively indicated at all their values.

The New FM Radar System

The block diagram of this system and the frequencies present in it in the case of a stationary target are illustrated in Fig. 3 and Fig. 4 respectively. Here, the transmitter is sinusoidally fm at the lf f_m so that

$$F_s = F_0 + \Delta F \cos 2\pi f_m t \qquad (9)$$

with $\omega = 2\pi f$

$$\Omega_s = \Omega_0 + \Delta\Omega \cos \omega_m t. \qquad (10)$$

Hence, the transmitted signal u_s[4] will be

[4] Throughout this paper, the amplitude of any signal u_x will be designated by U_x.

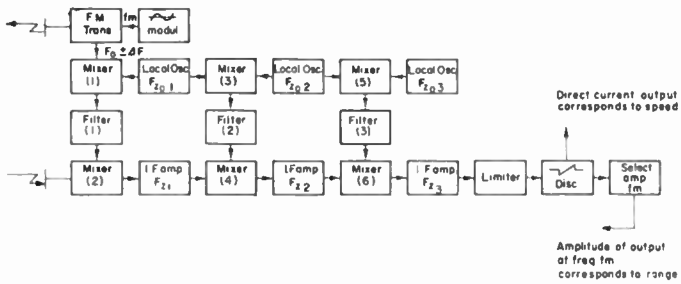

Fig. 3—Block diagram of the new system.

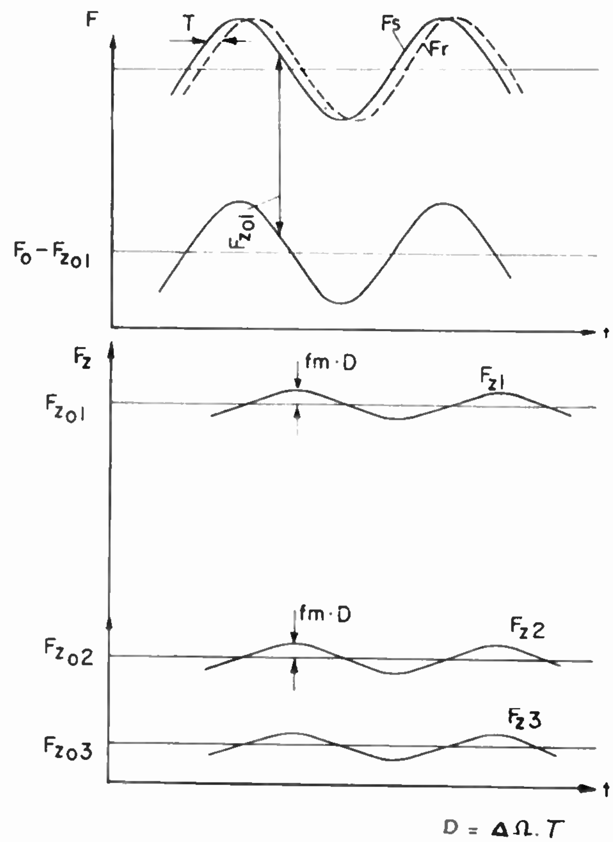

Fig. 4—Frequencies present in the new system.

$$
\begin{aligned}
u_s &= U_s \sin\left[\int \Omega_s dt\right] \\
&= U_s \sin\left[\Omega_0 t + \frac{\Delta\Omega}{\omega_m}\sin\omega_m t\right].
\end{aligned} \tag{11}
$$

For a certain target having a corresponding delay time T and a speed velocity f_v, the received signal u_r will be

$$
\begin{aligned}
u_r = U_r \sin\Big[&\Omega_0(t - T - T') \\
&+ \omega_v t + \frac{\Delta\Omega}{\omega_m}\sin\omega_m(t - T - T') + \phi_K\Big]
\end{aligned} \tag{12}
$$

where T' = the delay time corresponding to the feeding cables from the receiving antenna to mixer 2 in the receiver and from the transmitter to the transmitting antenna.

$\phi_K = a$ constant phase angle depending upon the reflection coefficient of the target.

In mixer 1 a signal from the transmitter at frequency $F_0 \pm \Delta F$ is heterodyned with a local signal at a fixed frequency $F_{Z_{01}}$. The mixer's output is then applied to band-pass filter 1 which passes only the lower sideband at a frequency $(F_0 - F_{Z_{01}}) \pm \Delta F$ and rejects strongly both the carrier at a frequency $F_0 \pm \Delta F$ and the upper sideband at frequency $(F_0 + F_{Z_{01}}) \pm \Delta F$. Frequency $F_{Z_{01}}$, however, should be made high enough to allow complete rejection of the last two components. Naturally, the passing of the required signal through filter 1 requires a certain delay time

$$
T'' = \frac{d\phi}{d\omega} \tag{13}
$$

so that $d\phi/d\omega$ is the slope of the linear portion of the phase characteristics of the filter, with the phase angle ϕ expressed in radians. The resulting output of the filter will, therefore, have the form

$$
\begin{aligned}
u_{F_1} = U_{F_1} \sin\Big[&(\Omega_0 - \Omega_{Z_{01}})(t - T'') \\
&+ \frac{\Delta\Omega}{\omega_m}\sin\omega_m(t - T'') + \phi_1\Big]
\end{aligned} \tag{14}
$$

where ϕ = a constant phase angle determined by the characteristics of mixer 1 and filter 1.

With filter 1 so designed as to have

$$
T'' = T' \tag{15}
$$

the result is

$$
\begin{aligned}
u_{F_1} = U_{F_1} \sin\Big[&(\Omega_0 - \Omega_{Z_{01}})(t - T') \\
&+ \frac{\Delta\Omega}{\omega_m}\sin\omega_m(t - T') + \phi_1\Big].
\end{aligned} \tag{16}
$$

Signal u_{F_1} is then mixed with u_r in mixer 2 giving

$$
\begin{aligned}
u_{Z_1} = U_{Z_1} \cos\Big[&(\Omega_{Z_{01}} + \omega_v)t \\
&- \frac{2\Delta\Omega}{\omega_m}\sin\frac{\omega_m T}{2}\cdot\cos\omega_m\left(t - \frac{T}{2} - T'\right) \\
&+ (\Omega_0 T - \Omega_{Z_{01}} T' + \phi_K - \phi_1)\Big].
\end{aligned} \tag{17}
$$

If $\omega_m T/2$ is substituted for $\sin\omega_m T/2$ [T is always $\ll (1/f_m)$], and the small phase angle $-\omega_m[T' + (T/2)]$ is neglected (it does not play any role in the following calculations), (17) reduces to

$$u_{Z_1} = U_{Z_1} \cos \overbrace{[(\Omega_{Z_{01}} + \omega_v)t - \Delta\Omega \cdot T \cos \omega_m t + \phi']}^{\phi_{Z_1}}. \quad (18)$$

From (18) it is seen that the IF signal u_{Z_1} is also sinusoidally fm at the lf f_m, and its instantaneous frequency can be calculated as

$$F_{Z_1} = \frac{1}{2\pi}\frac{d}{dt}\phi_{Z_1} = F_{Z_{01}} + f_v + f_m \cdot \Delta\Omega \cdot T \sin \omega_m t. \quad (19)$$

From (19) it is apparent that the frequency difference between the mean frequency of F_{Z_1} and $F_{Z_{01}}$ is the speed frequency f_v, which is directly proportional to the target's speed, while its maximum frequency deviation $f_m.\Delta\Omega.T$ is directly proportional to the target's range. Unfortunately, because f_v and $f_m.\Delta\Omega.T$ are always extremely small relative to $F_{Z_{01}}$, their determination and hence the indication of both the range and the speed of the target by means of a usual frequency discriminator, is practically impossible.

Example

$F_0 = 1500$ mc,	$\Delta F = 1$ mc,
$F_{Z_{01}} = 120$ mc,	$f_m = 120$ cps,
$d = 10$ m,	$v = 10$ m,
$f_m \cdot \Delta\Omega \cdot T = 50$ cps,	$= 0.000044$ per cent $F_{z_{01}}$

and

$f_v = 100$ cps,	$= 0.00009$ per cent $F_{z_{01}}$

In order to make possible the measuring of f_v and $f_m \cdot \Delta\Omega \cdot T$, the IF F_{Z_1} should be converted to a much lower frequency of some ten kc without affecting the values of the two quantities to be measured. Direct frequency conversion, *i.e.*, by mixing signal u_{Z_1} with a constant frequency signal having these few kc frequency differences from $F_{Z_{01}}$, does not lead to a successful result. Because the frequency stability of oscillators working at such high frequencies as $F_{Z_{01}}$ is usually low, the mean frequency of the signal after conversion will always be liable to appreciably large frequency changes. This makes it quite impossible to measure the speed frequency f_v and the maximum frequency deviation $f_m \cdot \Delta\Omega \cdot T$.

In the method used, the required frequency conversion is carried out quite successfully without being affected in any way by the stability of the hf oscillators. As shown in Fig. 3, a signal from a second local oscillator at frequency $F_{Z_{02}}$ is heterodyned with a signal at frequency $F_{Z_{01}}$ in mixer 3, whose output is applied to a band-pass filter 2 which passes only the lower sideband at frequency $(F_{Z_{01}} - F_{Z_{02}})$ and rejects completely other signals at frequencies $F_{Z_{01}}$ and $(F_{Z_{01}} + F_{Z_{02}})$. Frequency $F_{Z_{02}}$ is made much smaller than $F_{Z_{01}}$, with its minimum value determined by that which allows the proper design of filter 2. The output of filter 2 is then heterodyned with u_{Z_1} in mixer 4, giving another IF output signal u_{Z_2}, so that

$$u_Z = U_{Z_2} \cos [\Omega_{Z_{02}} t + \omega_v t - \Delta\Omega \cdot T \cos \omega_m t + \phi'']; \quad (20)$$

i.e.,

$$F_{Z_2} = F_{Z_{02}} + f_v + fm \cdot \Delta\Omega \cdot T \sin \omega_m t. \quad (21)$$

Hence, it is seen that the IF of u_{Z_2} has now become $F_{Z_{02}}$ instead of $F_{Z_{01}}$ for signal u_{Z_1}. If, for example, $F_{Z_{01}} = 15\ F_{Z_{02}}$, the IF will be decreased by 15 times, and the resulting IF will have a frequency stability equal to that of $F_{Z_{02}}$, which is absolutely independent of that of the high frequency $F_{Z_{01}}$.

Further frequency conversion can be carried out in similar steps (two or three such steps are sufficient) until a suitable low value is reached for the IF at which the two quantities f_v and $f_m.\Delta\Omega.T$ can be measured accurately and sensitively by normal frequency discriminators. Also, the frequency stability of the last IF can be made very high by using a crystal oscillator for the last oscillator.

In Fig. 3, only two steps of frequency conversion are made, producing a signal u_{Z_3}, so that

$$u_{Z_3} = U_{Z_3} \cos [\Omega_{Z_{03}} t + \omega_v t - \Delta\Omega T \cos \omega_m t + \phi''']; \quad (22)$$

i.e.,

$$F_{Z_3} = F_{Z_{03}} + f_v + f_m \cdot \Delta\Omega \cdot T \sin \omega_m t. \quad (23)$$

If, therefore, signal u_{Z_3} is amplified, limited, and then applied to a frequency discriminator having a center frequency $F_{Z_{03}}$, its output will contain two components:

An ac component at a frequency f_m, whose amplitude is proportional to the target's range, and a dc component whose value is proportional to the mean frequency shift f_v; *i.e.*, to the target's speed.

As shown in Fig. 3, a dc instrument calibrated in unit lengths/sec and an lf amplifier at frequency f_m feeding an ac instrument calibrated in unit lengths may be fed with the discriminator output to indicate directly the target's speed and range respectively.

According to (23), it is to be noticed that $f_m \cdot \Delta\Omega \cdot T$ changes linearly with the target's range and hence no fixed error exists. This makes it possible to achieve a high accuracy in measuring all target ranges (even if they are very short) by using relatively small frequency deviations of the transmitter frequency, which may be only a fraction of that used in the usual fm radar system. Also, both the range and the speed of the target can be always determined correctly with no fear whatsoever of any falsity in their indications.

Four different arrangements for the new system which afford the measuring of the target's range and speed in the same way described above, are shown in Figs. 5 through 8. In the block diagrams of Fig. 5 and Fig. 6 some modifications regarding the frequencies of the local oscillators and the way of mixing are carried out. The working of both block diagrams can easily be understood by referring to them.

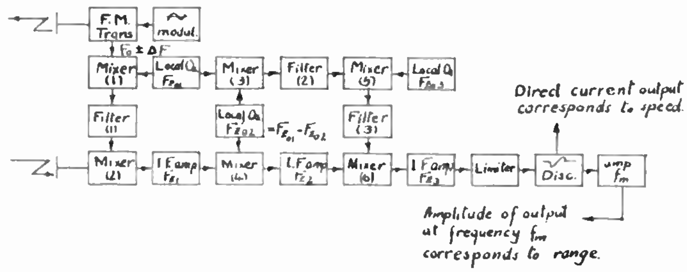

Fig. 5—A second arrangement for the new system.

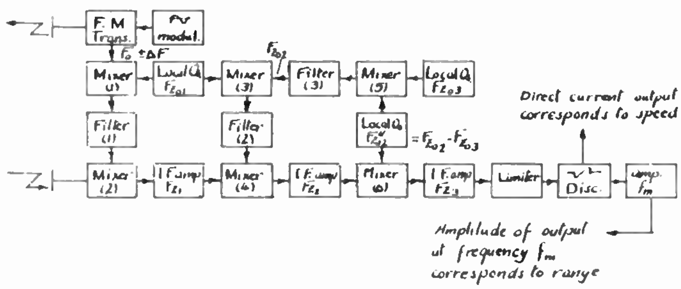

Fig. 6—A third arrangement for the new system.

The two arrangements of Fig. 7 and Fig. 8 are intended to facilitate the application of the new system at the highest radar frequencies used. This is done by eliminating the most difficult process at these frequencies; *i.e.*, mixing the high frequency $F_0 \pm \Delta F$ with $F_{Z_{01}}$ ($F_{Z_{01}}$ is always $\ll F_0$) and then selecting the lower sideband at frequency $F_0 - F_{Z_{01}} \pm \Delta F$ to pass through filter 1.

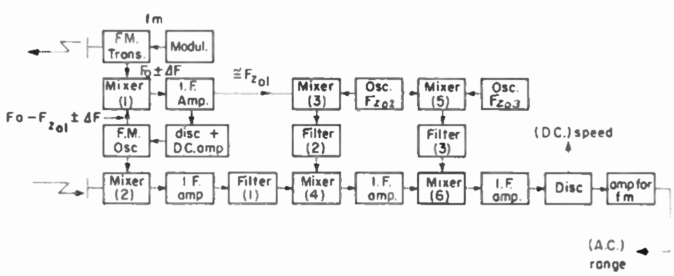

Fig. 7—A fourth arrangement for the new system.

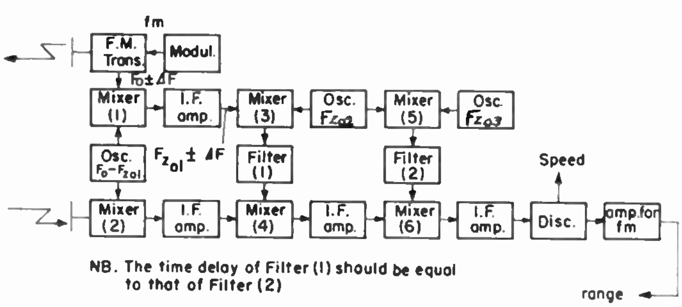

Fig. 8—A fifth arrangement for the new system.

In the arrangement of Fig. 7, the frequency of a local oscillator (which can also be fm) is made to follow automatically the variation of the transmitter frequency so that the frequency difference (this corresponds to $F_{Z_{01}}$ in the arrangement of Fig. 3) between them remains constant. The radar data are then obtained in the same way explained before.

In the arrangement of Fig. 8, a fixed frequency ($F_0 - F_{Z_{01}}$) of a local oscillator is heterodyned with both F_s and F_r producing two fm IF signals, $u_{Z_{(1s)}}$ and $u_{Z_{(1r)}}$ respectively, which are then amplified in two similar IF amplifiers having a pass band $>2\Delta F$. (Usually, $F_{Z_{01}}$ is more than 50 mc while ΔF does not exceed 1.5 mc in the new system.) In mixer 3, $u_{Z_{(1s)}}$ is heretodyned with a local oscillator's signal at frequency $F_{Z_{02}}$, and the higher sideband of frequency $F_{Z_{01}} + F_{Z_{02}} \pm \Delta F$ is passed through filter 1 to mix with $u_{Z_{(1r)}}$ in mixer 4, giving a signal u_{Z_2} (20). The radar data are then obtained as explained before.

Experimental Verification of the New System

In order to investigate the new system's performance,[5] an experimental radar apparatus has been built after the arrangement shown in Fig. 3. The following operating frequencies have been chosen.

$F_0 = 216$ mc, $f_m = 280$ cps,
$\Delta_F =$ arbitrary with a maximum value of about 1.5 mc,
$F_{Z_{01}} = 35$ mc, $F_{Z_{02}} = 3$ mc.
$F_{Z_{03}} = 170$ kc,

An exact and simple experimental method which simulates the working of the radar apparatus against a stationary target at a range d is to connect a section of ordinary rf cable having an electrical length of 2 d between the radar transmitter output and the receiver input. In Fig. 9 the indicated range (expressed in volts) in function of cable length at a transmitter frequency

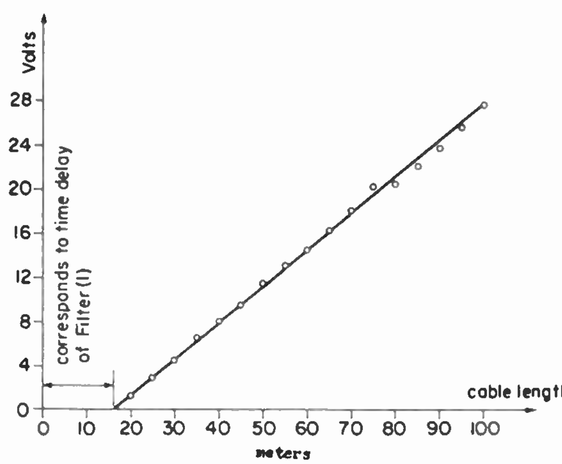

Fig. 9—The indicated range in function of cable length.

[5] This could only be available experimentally with respect to fixed ranges.

deviation ΔF of only 400 kc is drawn. This shows that the relation between them is a straight line which intersects the x axis at a cable length of about 16 m. This cable length should have, therefore, a time delay T'' equal to that of filter 1.

In practice, this time delay T'' should be compensated in order to make the zero range reading of the radar apparatus occur at zero range. As has been mentioned before, this can be achieved by making the time delay T' of the feeding cables from the transmitter to the transmitting antenna and from the receiving antenna to the receiver $\geqq T''$. If T' remains $<T''$, an additional band-pass filter at a mean frequency F_0 and having a pass band $>2\Delta F$ can be built at the receiver's input, giving a delay time T''' so that $T'+T'''\geqq T''$.

Conclusion

From the practical results represented by the straight line of Fig. 9 it can be concluded that the new fm radar system gives successful performance and is capable of indicating accurately and sensitively both the range and the speed of the target even at ranges as short as 50 cm.

After the experimental apparatus described above proved to be successful, the author built for Hasler Laboratories a prototype for a low level airplane altimeter, the photos of which are shown in Fig. 10 and Fig. 11. This altimeter is built after the block diagram of Fig. 5 and has the following characteristics.

Operating frequency $=440$ mc.

Modulating frequency $\cong 200$ cps.

$\Delta F \cong 1.5$ mc for 0–400 feet range.

$\Delta F \cong 0.15$ mc for 0–4000 feet range.

$F_{Z_{01}}=50$ mc.

$F_{Z_{02}}=47.8$ mc.

$F_{Z_{03}}=200$ kc.

Power consumption $=90$ watts.

Supply voltage $=24$ to 27 volts dc.

The apparatus was tested by different hf cable lengths and proved to be precise enough, so that a change in cable length of only one meter could accurately be read on the radio-altitude meter.

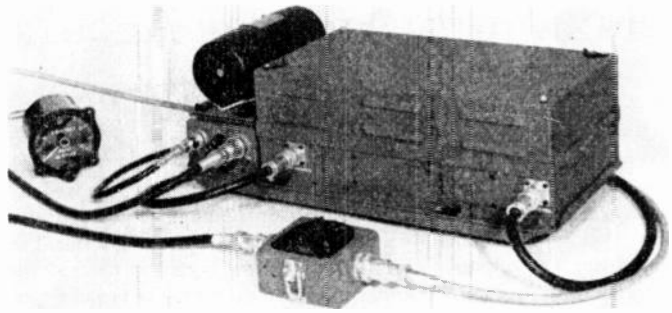

Fig. 10—The fm radar airplane altimeter designed for Hasler Laboratories, Bern, Switzerland.

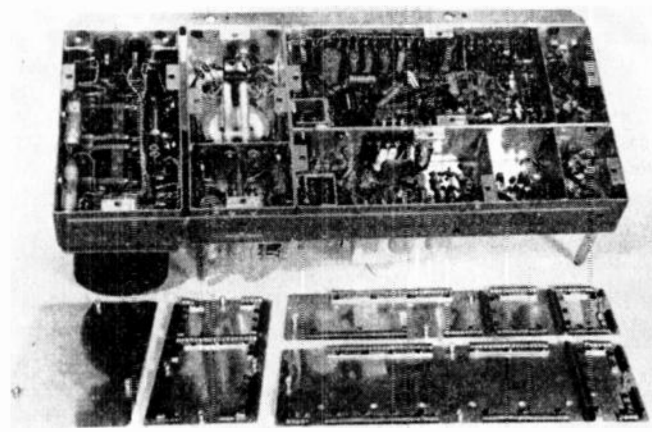

Fig. 11—Airplane altimeter with bottom shields removed.

Acknowledgment

The author is indebted to Prof. F. Tank, Head of the Institute of High Frequency Techniques at the Swiss Federal Institute of Technology, Zurich, for his encouragement during the carrying out of this work. Thanks are also extended to Hasler Laboratories for providing the necessary facilities to develop this new type of airplane altimeter.

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Maximum Angular Accuracy of a Pulsed Search Radar*

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Summary—**An investigation is made of the limits imposed by receiver noise on the accuracy with which the angular position of a target can be determined by a pulsed search radar. Using a result in the theory of statistical estimation, a lower bound is derived for the standard deviation of regular unbiased estimates of target angular position, for a large class of methods of angular position determination; the lower bound depends on scan rate, pulse repetition rate, beamwidth, beam shape, and signal-to-noise ratio. A similar analysis is made of the limits on angular accuracy imposed by a combination of receiver noise and one particular type of target cross section fluctuation.**

Operations which can be performed on the received signal to form an estimate of target angular position, the standard deviation of which approximately attains the theoretical lower bound, are discussed. The relation between the estimation of angular position and the problem of target detection is discussed.

A graphical presentation of the main results is given.

List of Symbols

A = angular sector scanned by the radar.

f = function describing the two-way power gain pattern of the beam.

N = number of pulses emitted as the beam travels through an angle 2β.

P_D = probability of detection.

X_0 = signal-to-noise power ratio at the input to the second detector, for a pulse emitted when the nose of the beam is pointed directly at the target (averaged over target fluctuations in the case of a fluctuating target).

β = parameter related to $\frac{1}{2}$-beamwidth [see (10) and ensuing discussion].

θ_T = angular position of the target.

$\hat{\theta}_T$ = estimate of θ_T.

$\tilde{\theta}_T$ = maximum likelihood estimate of θ_T.

$\sigma(\hat{\theta}_T)$ = standard deviation of $\hat{\theta}_T-\theta_T$.

$\sigma_{\min}$ = theoretical lower bound of $\sigma(\hat{\theta}_T)$.

$\Delta\theta$ = angle moved by the beam between successive pulses.

I. Statement of the Problem

THE ACCURACY with which the angular position of a target can be determined by a search radar is sometimes considered to be roughly equal to the width of the beam. It is, of course, well known that under certain conditions the target's angular position can be determined with an error that is, on the average, considerably smaller than the beamwidth. For example, on a ppi scope a target may appear as a small arc. If the target position is taken to be the center of the arc, the resulting error in the estimation of target angular position may be considerably smaller than the beamwidth. This is just one example of the manner in which the modulation of the signal returned from the target, caused by the varying gain of the beam as it sweeps across the target, can be utilized to give a more accurate estimate of target angular position. If the beam gain pattern were known exactly, and if no noise entered the system from any source, the angular position of the target could in many cases be determined exactly. Various sources of noise—such as receiver noise and target cross section fluctuation—will impose limits on the accuracy with which the target's angular position can be determined. The main purpose of this paper is to investigate the limits imposed by receiver noise on the accuracy with which the angular position of a target can be determined by a pulsed search radar, for a large class of methods of angular position determination.

Consider a pulsed radar scanning a certain angular sector A (which may be an entire circle). The scanning is assumed to be in only one angular coordinate. Also, only point targets which are stationary during any single sweep will be considered.[1] The coordinates of such a target may be given as (R, θ) where R is the range and θ is the angular coordinate measured from some arbitrarily chosen axis. The range coordinate is assumed to be divided into concentric range intervals, *e.g.*, by range gates, which may be treated as effectively independent of each other.

The main problem considered is the following: assuming a single target to be present in a given range interval, what limits are imposed by receiver noise on the accuracy with which its angular position can be determined on a single sweep (*i.e.*, without using information about its position on previous sweeps)? Furthermore, what operations can be performed upon the received signal pulses to actually approach the maximum theoretical accuracy? The relation between this problem and the detection problem, *i.e.*, the problem of deciding whether or not a target actually is present at the given range, will also be discussed. Also, in Section V, an analysis is made of the limits on angular accuracy imposed by a combination of receiver noise and a particular type of target fluctuation.

Let

θ_T = true angular position of the target,

$\Delta\theta$ = angle moved by the beam between successive pulses.

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[1] By "sweep" is meant one complete scan by the beam over the sector A.

Let A be divided into intervals of width $\Delta\theta$. It is assumed that A contains a large number K of such intervals, of the form[2]

$$i\text{th interval} = (\Delta\theta)_i = [(i-1)\Delta\theta,\ i\Delta\theta],\quad i = 1, \cdots, K.$$

Let the "position of a pulse" be the position in which the beam is pointing when the pulse is emitted. By "position in which the beam is pointing" is meant the angular position of some arbitrarily chosen reference line fixed with respect to the beam, such as a line through the nose of the beam.

Each interval $(\Delta\theta)_i$ will then contain exactly one pulse emitted in that interval. In any single sweeep of the sector A, all pulses will, of course, occupy the same relative position within their respective intervals. The position of the pulse within the ith interval will be denoted by θ_i. Thus, $\theta_i = \theta_1 + (i-1)\Delta\theta$.

It is assumed that, for a large succession of sweeps, θ_1 is distributed with constant probability density over the interval $(\Delta\theta)_1$; hence, each θ_i is, over a large number of sweeps, distributed with constant probability density within the interval $(\Delta\theta)_i$. This simply reflects the fact that the scan rate is not in general perfectly synchronized with the pulse repetition rate.

The block diagram of the radar receiver and angular position computer is given in Fig. 1.

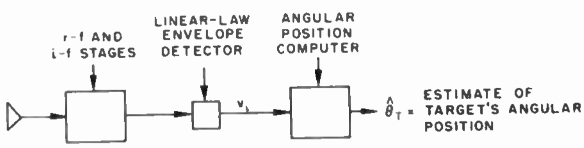

Fig. 1—Block diagram of receiver and angular position computer.

It is clear that the linear-law envelope detector could be replaced by an envelope detector having any strictly monotonic law, without affecting the results.

The input to the computer on each sweep (and for each independent range interval) is a series of pulses v_i, $i = 1, \cdots, K$. Here v_i is taken to be the pulse occupying the ith interval. These pulses consist of either signal plus circuit noise or noise alone. The circuit noise in the predetection stages of the receiver is assumed to be additive Gaussian noise, with a frequency spectrum shaped mainly by the IF. The IF bandwidth is assumed to be such that the noise components of successive pulses are statistically independent, and that each pulse can be assumed to present effectively a single independent voltage magnitude.

On each sweep, and for each independent range interval, the angular position computer assumes a single target to be present at the given range and operates on the received voltage pulses v_i in such a way as to form an estimate $\hat{\theta}_T$ of its position.

[2] Since, of course, A will not in general contain an integral number of such intervals, the above enumeration will leave out some very small portion of A; if K is large, the neglect of this portion will have no appreciable effect on the results.

The block diagram of Fig. 1 serves to delineate the class of angular position determination methods which is considered in this paper. It is clear that we are considering methods where only the information contained in the pulse amplitudes is used; methods which use phase data across the antenna aperture are not considered. Also, we will not consider radars using two receiving antennas with overlapping beams, or the equivalent.

The assumption that the circuit noise components of successive pulses are statistically independent implies that we are not considering systems employing comb-filter integration in the IF, or the equivalent.

The methods of analysis used in this paper can, however, probably be extended to systems using overlapping beams or IF integration; extension of the analysis to these cases would be a useful problem for further investigation.

In an actual radar, of course, a further operation would be performed on the received pulses v_i in such a way as to decide whether or not there actually is a target present at the given range; this will be discussed in Section IV.

Let the variance of the estimate $\hat{\theta}_T$ about the true target position θ_T be denoted by

$$\sigma^2(\hat{\theta}_T) = E\{(\hat{\theta}_T - \theta_T)^2\} \tag{1}$$

where $E\{\ \}$ denotes the expected value of the quantity in braces.[3]

The standard deviation will be denoted by $\sigma(\hat{\theta}_T)$.

An estimate $\hat{\theta}_T$ is called *unbiased* if

$$E(\hat{\theta}_T - \theta_T) = 0. \tag{2}$$

Also, an estimate $\hat{\theta}_T$ is called "regular" if it is a function of $v_1, \cdots, v_K$ satisfying certain differentiability conditions.[4]

The main problem to be considered can now be stated: for a given beam shape and given values of relevant parameters such as signal-to-noise ratio, pulse repetition rate, etc., find a lower bound for the quantity $\sigma(\hat{\theta}_T)$, for all regular unbiased estimates $\hat{\theta}_T$. If possible, find a lower bound which is approximately a greatest lower bound, in the sense that estimates $\hat{\theta}_T$ exist for which $\sigma(\hat{\theta}_T)$ nearly attains the lower bound.

II. A Lower Bound for $\sigma(\hat{\theta}_T)$—Nonfluctuating Targets

In this section it is assumed that the radar cross section of the target is nonfluctuating. This assumption is also retained in Sections III and IV.

To solve the problem formulated in Section I, one can apply results in mathematical statistics stated by Cramer.[5] The application of these results to the case at hand is as follows.

[3] All expected values are taken with respect to the joint probability distribution of $v_1, \cdots, v_K$.

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

[5] *Ibid.*, ch. 32.

If a target is present at position θ_T, and the pulses are emitted at certain positions θ_i, the joint probability density function for the K voltages $v_i \cdots, v_K$ (normalized by dividing by the rms noise voltage) is,[6,7]

$$L[v_1, \cdots, v_K \mid \theta_T; \{\theta_i\}] = \prod_{i=1}^{K} v_i \exp\left[-\left(\frac{v_i^2}{2} + x_i\right)\right] I_0[v_i\sqrt{2x_i}]. \quad (3)$$

Here

I_0 is the modified Bessel function of the first kind, order zero

x_i is the signal-to-noise power ratio (at the input to the detector) for the ith pulse.

Eq. (3) arises from the equation given by Rice[6] for the probability density of the envelope of a sine wave plus Gaussian noise. According to Rice, if the ratio of the power in the sine wave to the mean noise power is x, the density function for the envelope of the sine wave plus noise is given by the function

$$ve^{-(v^2/2+x)}I_0(v\sqrt{2x}).$$

Now x_i is a function of θ_i and θ_T:

$$x_i = x(\theta_i, \theta_T) = X_0\phi(\theta_i, \theta_T). \quad (4)$$

Here X_0 is defined to be the signal-to-noise ratio for a pulse emitted when $\theta = \theta_T$. X_0 depends on the various quantities entering into the radar range equation; since some of these quantities, such as target cross section, may be unknown, X_0 must in general be considered a parameter which is not known *a priori*.

The function $\phi(\theta_i, \theta_T)$ depends mainly on the beam gain pattern, and to some extent on the scan rate, the pulse repetition frequency, and the range R of the target.

Now, taking into account the fact $\theta_i = \theta_1 + (i-1)\Delta\theta$ and that, for a large number of sweeps, the quantity θ_1 is uniformly distributed over the interval $(\Delta\theta)_1$, the overall joint probability density function for the voltages $v_1, \cdots, v_K$ if the target is at θ_T is:

$$L[v_1, \cdots, v_K \mid \theta_T] = \int_{(\Delta\theta)_1} \prod_{i=1}^{K} v_i \exp\left[-\left(\frac{v_i^2}{2} + x_i\right)\right] I_0[v_i\sqrt{2x_i}]d\theta_1 \quad (5)$$

where $x_i = X_0\phi(\theta_i, \theta_T)$; $\theta_i = \theta_1 + (i-1)\Delta\theta$.

The quantity $\log_e L$ will be denoted by λ:

$$\lambda[v_1, \cdots, v_K \mid \theta_T] = \log_e L[v_1, \cdots, v_K \mid \theta_T]. \quad (6)$$

A theorem cited by Cramer[4] states that for any regular unbiased estimate $\hat{\theta}_T$,

$$\sigma^2(\hat{\theta}_T) \geqq \frac{1}{E\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2\right\}} \cdot \frac{1}{1-\rho^2} \quad (7)$$

where

$$\rho^2 = \frac{E^2\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)\left(\frac{\partial\lambda}{\partial X_0}\right)\right\}}{E\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2\right\}E\left\{\left(\frac{\partial\lambda}{\partial X_0}\right)^2\right\}}. \quad (8)$$

Here $E\{\ \}$ denotes the expected value of the quantity in braces, taken with respect to the joint probability distribution of $v_1 \cdots, v_K$.

This formula is for the case where X_0 is regarded as a parameter which is not known *a priori*.[8]

It is convenient to denote the square root of the quantity on the right side of (7) by $\sigma_{\min}$

$$\sigma_{\min} = \frac{1}{\sqrt{E\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2\right\}}} \cdot \frac{1}{\sqrt{1-\rho^2}}. \quad (9)$$

Applying (5) and (6) would in principle enable one to evaluate $\sigma_{\min}$. However, the mathematical difficulties are formidable unless some approximations are made. Several of the necessary approximations will follow from the following assumption: the beamwidth is large compared with $\Delta\theta$.

It is henceforth assumed that $\Delta\theta$ is sufficiently small for the motion of the beam between transmission and reception of any given pulse and for the averaging process of (5) to be neglected for purposes of evaluating $\sigma_{\min}$. (On the other hand, it is necessary, of course, to assume that $\Delta\theta$ is large enough so that the assumption of discrete pulses with statistically independent noise components can be maintained. This will be true as long as $\Delta\theta$ is larger than, say, the pulse width.)

It is also assumed that the function $x_i = x(\theta_i, \theta_T)$ is of the following form

$$x_i = x(\theta_i, \theta_T) = X_0 f\left[\frac{\theta_i - \theta_T}{\beta}\right] \quad (10)$$

where f is an even function of its argument and the derivative of f is an odd function of its argument. The function f can be identified with the two-way power gain pattern. Since X_0 is defined to be the signal-to-noise ratio for a pulse emitted when $\theta_i = \theta_T$, we have $f(0) = 1$.

The parameter β is a scale factor related to the beamwidth. Any actual beam pattern can be represented by

[6] S. O. Rice, "Mathematical analysis of random noise," *Bell Syst. Tech. J.*, vol. 23, pp. 282–332; July, 1944; vol. 24, pp. 46–156; January, 1945.

[7] J. I. Marcum, "A statistical theory of target detection by pulsed radar: mathematical appendix," The Rand Corp., Res. Memo. RM-753; July 1, 1948.

[8] Technically, the methods used by Cramer to derive (7) involve the assumption that an unbiased estimate of X_0 exists; however, his arguments can easily be generalized so that one need only assume the existence of an estimate of X_0, the bias of which is independent of θ_T; such an assumption is presumably justified.

an infinite number of choices of f and β.[9] However, it is most convenient to choose f in such a way that 2β is equal to the beamwidth, where "beamwidth" is defined in some reasonable manner. For example, suppose the beam pattern has the shape of the Gaussian curve. Then, if f is chosen to be $f(u)=e^{-u^2}$, 2β will be equal to the beamwidth, provided beamwidth is defined to be the angle between the $(1/e)$-power points of the two-way beam. It is also easy, once f is given, to relate β to any other conventional definition of beamwidth. In the example just cited, if beamwidth were defined to be, say, the angle between the $\frac{1}{2}$-power points of the one-way beam, then 2β would be equal to 0.85 times the beamwidth.

It is also convenient to define

$$N = \frac{2\beta}{\Delta\theta}\cdot \tag{11}$$

That is, N is the number of pulses emitted as the beam travels through an angle 2β.

It is assumed, then, that $\Delta\theta$ is small compared to the beamwidth. The answer to the question of how small, is simply small enough so that the necessary approximations are good.

Since, in particular, $\Delta\theta$ is assumed small enough so that the averaging process of (5) can be neglected, one can write

$$\lambda[v_i, \cdots, v_K \mid \theta_T]$$
$$\approx \sum_{i=1}^{K}\left(\log_e v_i - x_i - \frac{v_i^2}{2}\right) + \sum_{i=1}^{K} \log_e I_0(v_i\sqrt{2x_i}) \tag{12}$$

where the x_i are given by (10).

The additional assumption is made that the sector A is several times as large as the significant portion of the beam. Expressed in other words, it is assumed that the sector A can be considered effectively infinite with respect to the beam. Then, at least for θ_T not near the edges of A, edge effects can be neglected. In particular, at least for a certain range of θ_T (not near the edges of A), the quantity $\sum x_i$ will be approximately independent of θ_T.

Also, of course, the quantity $\sum[\log_e v_i - \frac{1}{2}v_i^2]$ is independent of θ_T. Therefore,

$$\frac{\partial\lambda}{\partial\theta_T} \approx \sum_{i=1}^{K} \frac{\partial}{\partial\theta_T}\log_e I_0(v_i\sqrt{2x_i})$$

$$\approx -\frac{1}{\beta}\sqrt{\frac{X_0}{2}}\sum_{i=1}^{K}\frac{I_1(v_i\sqrt{2x_i})}{I_0(v_i\sqrt{2x_i})}\frac{f'\left(\dfrac{\theta_i-\theta_T}{\beta}\right)}{\sqrt{f\left(\dfrac{\theta_i-\theta_T}{\beta}\right)}}v_i \tag{13}$$

where I_1 is the modified Bessel function of the first kind, of order one.

[9] For example, suppose the beam is Gaussian-shaped. A given Gaussian-shaped beam could be represented by taking $f(u) = \exp(-k^2u^2)$, where k^2 is any positive number, provided β is chosen so that the ratio k/β is always the same.

Let

$$w_i = \frac{I_1(v_i\sqrt{2x_i})}{I_0(v_i\sqrt{2x_i})}v_i \tag{14}$$

and

$$f_i = f\left(\frac{\theta_i-\theta_T}{\beta}\right), \qquad f_i' = f'\left(\frac{\theta_i-\theta_T}{\beta}\right)\cdot \tag{15}$$

Then

$$\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2 \approx \frac{X_0}{2\beta^2}\left\{\sum_{i=1}^{K} w_i^2\frac{f_i'^2}{f_i} + \sum_{i\neq j} w_iw_j\frac{f_i'f_j'}{\sqrt{f_if_j}}\right\}. \tag{16}$$

Now let

$$\text{Expected value of } w_i = \bar{w}_i$$
$$\text{Expected value of } (w_i-\bar{w}_i)^2 = \sigma^2(w_i). \tag{17}$$

Then, again denoting expected values by $E\{\ \}$, and using the fact that w_i and w_j are independent for $i\neq j$ so that $E(w_iw_j)=\bar{w}_i\bar{w}_j$, one obtains

$$E\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2\right\} \approx \frac{X_0}{2\beta^2}\left\{\left[\sum_i \bar{w}_i\frac{f_i'}{\sqrt{f_i}}\right]^2 + \sum_i \sigma^2(w_i)\frac{f_i'^2}{f_i}\right\}. \tag{18}$$

Since, however, f is assumed to be an even function of its argument, and f' an odd function of its argument, the first sum on the right side of (18) is approximately zero, so that

$$E\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2\right\} \approx \frac{X_0}{2\beta^2}\sum_{i=1}^{K}\sigma^2(w_i)\frac{f_i'^2}{f_i} \tag{19}$$

where w_i, f_i', and f_i are given by (14) and (15).

Also,

$$\frac{\partial\lambda}{\partial X_0} \approx -\sum_{i=1}^{K} f_i + \frac{1}{\sqrt{2X_0}}\sum_{i=1}^{K} w_i\sqrt{f_i}\,. \tag{20}$$

The quantity ρ can be evaluated in a manner similar to the evaluation of

$$E\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2\right\}.$$

The result is

$$\rho \approx 0. \tag{21}$$

The quantity $\sigma_{\min}$ has now been evaluated. However, it is still not in a form very suitable for computational purposes. The expression can be considerably simplified by assuming that $\Delta\theta$ is sufficiently small so that the sum in (19) can be replaced by an integral.

If this is done, one obtains finally

$$E\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2\right\} \approx \frac{X_0N}{4\beta^2}\int_{-\infty}^{\infty}\frac{f'^2(u)}{f(u)}G(u)du. \tag{22}$$

The function $G(u)$ is defined as follows.

Let

$$x(u) = X_0 f(u). \tag{23}$$

Then,

$$G(u) = \int_0^\infty v^3 \exp\left[-\left(\frac{v^2}{2} + x(u)\right)\right] \frac{I_1{}^2[v\sqrt{2x(u)}]}{I_0[v\sqrt{2x(u)}]}\, dv$$
$$- \left\{\int_0^\infty v^2 \exp\left[-\left(\frac{v^2}{2} + x(u)\right)\right] I_1[v\sqrt{2x(u)}]dv\right\}^2. \tag{24}$$

Putting (22) and (21) into (19), we get for all regular unbiased estimates θ_T,

$$\sigma(\hat{\theta}_T) \geqq \sigma_{\min},$$

where

$$\sigma_{\min} \approx \frac{2\beta}{\sqrt{NX_0}} \frac{1}{\sqrt{\int_{-\infty}^{\infty} \frac{f'^2(u)}{f(u)} G(u) du}} \tag{25}$$

where $G(u)$ is given by (24). It is assumed that the function $f(u)$ is such that the integral appearing in (25) exists.

Eq. (25) should be considered applicable, however, only if the width of A is several times as large as $\sigma_{\min}$. The reason for this is that the validity of certain of the above assumptions breaks down as $\sigma_{\min}$ approaches the width of A. It is not immediately apparent why this is so. The reason is mainly that, as $\sigma_{\min}$ approaches the width of A, it is unjustified to restrict oneself to consideration of unbiased estimates, since actual estimates will in general become significantly biased. This is, however, a minor consideration. In almost all cases of interest, $\sigma_{\min}$ will be no larger than, say, the beamwidth.

The validity of the various approximations used in deriving the above formulas depends on the form of f. For most functions f that would be met with in practice, the necessary approximations would probably hold good even if $\Delta\theta$ is not very small (say $\Delta\theta \approx 1/10$ beamwidth). However, it should be mentioned that there are some mathematically conceivable functions f for which the approximations used would not hold good, no matter how small $\Delta\theta$ is. An example of this is the "square beam": $f(u) = \text{constant}$, $|u| \leqq 1$; $f(u) = 0$, $|u| > 1$. In particular, in this case the averaging process of (5) is an essential element which cannot be neglected.

Caution should also be used in applying the above methods to cases where the probability density functions are not so well-behaved as in the case at hand.

Fig. 2 shows $\sigma_{\min}$ as a function of N, B, and X_0, for $f(u) = e^{-u^2}$.

It is of interest to evaluate the asymptotic values of $\sigma_{\min}$ for large X_0 and small X_0. One can utilize certain formulas for Bessel functions for this purpose, namely

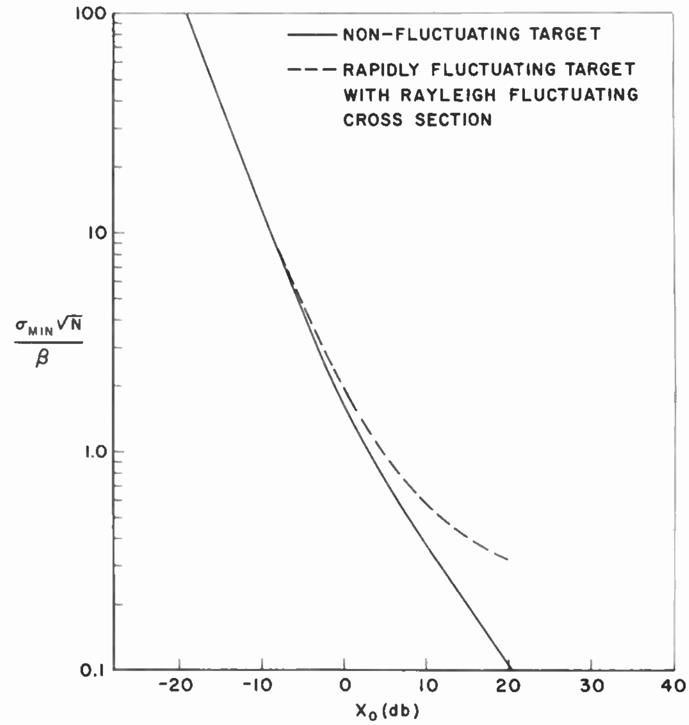

Fig. 2—$\sigma_{\min}$ as a function of X_0, N, and β for $f(u) = e^{-u}$.

$$\left.\begin{aligned} &1) \text{ for } x_i \ll 1, \quad \frac{I_1[v_i\sqrt{2x_i}]}{I_0[v_i\sqrt{2x_i}]} \approx v_i\sqrt{\frac{x_i}{2}}, \\ &2) \text{ for } x_i \gg 1, \quad \frac{I_1[v_i\sqrt{2x_i}]}{I_0[v_i\sqrt{2x_i}]} \approx 1. \end{aligned}\right. \tag{26}$$

Putting (26) into (24) and (25), one obtains 1) for $X_0 \ll 1$,

$$\sigma_{\min} \approx \frac{\beta}{X_0\sqrt{N}} \frac{1}{\sqrt{\frac{1}{2}\int_{-\infty}^{\infty} f'^2(u) du}}, \tag{27}$$

2) for $X_0 \gg 1$,

$$\sigma_{\min} \approx \frac{\beta}{\sqrt{NX_0}} \frac{2}{\sqrt{\int_{-\infty}^{\infty} \frac{f'^2(u)}{f(u)} du}}. \tag{28}$$

For $f(u) = e^{-u^2}$, this leads to

$$\left.\begin{aligned} &1) \ X_0 \ll 1: \\ &\qquad \sigma_{\min} \approx \frac{1.26\beta}{X_0\sqrt{N}} \\ &2) \ X_0 \ll 1: \\ &\qquad \sigma_{\min} \approx \frac{1.06\beta}{\sqrt{NX_0}} \end{aligned}\right\} \text{ for } f(u) = e^{-u}. \tag{29}$$

It is also of interest to consider the following question: suppose one chooses a certain value of X_0, say $X_0{}^*$, and attempts to optimize the estimate of θ_T for the particular chosen value $X_0{}^*$. Could one in this way obtain an estimate of θ_T, which, when X_0 actually happens to equal X_0^*, has standard deviation less than the value of $\sigma_{\min}$ given by (25)?

This question can be answered by applying the results of Cramer[4] to the case where X_0 is regarded as a known parameter; *i.e.*, to the case where θ_T is the only unknown parameter. Suppose we denote the lower bound thus derived for this case by $\sigma_{\min}'$. The applicable formula of Cramer states that $\sigma_{\min}'$ is just given by

$$\sigma_{\min}' = \frac{1}{E\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2\right\}}. \tag{30}$$

But, looking at (9), we see that

$$\sigma_{\min}' = \sigma_{\min}\sqrt{1-\rho^2} \tag{31}$$

where ρ is given by (8). But as we have seen, under the assumptions which were made in evaluating $\sigma_{\min}$, $\rho \approx 0$ [see (21)]. Therefore, under the assumptions which were made in evaluating $\sigma_{\min}$, the quantity $\sigma_{\min}$ given by (25) also gives a lower bound for the standard deviation of unbiased estimates of θ_T when X_0 is known *a priori*. This result, surprising at first sight, is further discussed in Section III.

III. Maximum Likelihood Estimation

In the preceding section, a lower bound for $\sigma(\hat{\theta}_T)$ for all regular unbiased estimates $\hat{\theta}_T$ was derived. This was denoted by $\sigma_{\min}$. So far nothing has been said about whether estimates exist whose standard deviation approximately attains the lower bound $\sigma_{\min}$.

It is at this point necessary to discuss the so-called maximum likelihood method of estimation. The likelihood function for the set of voltages $v_1, \cdots, v_K$ is

$$L(v_1, \cdots, v_K \mid \theta_T; X_0) = \int_{(\Delta\theta)_1} \prod_{i=1}^{K} v_i \exp\left[-\left(\frac{v_i^2}{2} + x_i\right)\right] I_0(v_i\sqrt{2x_i})d\theta_1 \tag{32}$$

where x_i is given by (4).

The so-called "maximum likelihood" method consists of finding, for any given set of voltages $v_1, \cdots, v_K$, the values $\tilde{\theta}_T$, $\tilde{X}_0$ which maximize L. It is assumed that the beam shape is such that this will give a unique answer with probability one. It can be shown that under the assumptions used in the evaluation of $\sigma_{\min}$, $\tilde{\theta}_T$ is approximately unbiased. Also, under these assumptions the integration in (32) can be neglected and $\tilde{X}_0$, $\tilde{\theta}_T$ will be, approximately, those values which maximize the quantity

$$\sum_{i=1}^{K}\left[\log_e v_i - \frac{v_i^2}{2} - x_i\right] + \sum_{i=1}^{K} \log_e I_0(v_i\sqrt{2x_i}). \tag{33}$$

If edge effects can be neglected and if x_i can be represented by (10), one obtains finally that $\tilde{X}_0$, $\tilde{\theta}_T$ are approximately the values which maximize

$$\mu(\theta_T, X_0) = -X_0\sum_{i=1}^{K} f\left(\frac{\theta_i-\theta_T}{\beta}\right) + \sum_{i=1}^{K}\log_e I_0\left[v_i\sqrt{2X_0 f\left(\frac{\theta_i-\theta_T}{\beta}\right)}\right]. \tag{34}$$

For large or small X_0, this amounts to the following: assume that for $X_0 \gg 1$, or $X_0 \ll 1$, the estimate $\tilde{X}_0$ has high probability of satisfying $\tilde{X}_0 \gg 1$ or $\tilde{X}_0 \ll 1$ respectively. The first term in (34) is approximately independent of θ_T (regardless of the value of X_0). In the second term, one can apply formulas for the function I_0 for large or small argument; if this is done, one finally obtains

1) $X_0 \ll 1$:

$$\tilde{\theta}_T = \text{value of } \theta_T$$

which maximizes

$$\sum_{i=1}^{K} v_i^2 f\left(\frac{\theta_i-\theta_T}{\beta}\right) \tag{35}$$

2) $X_0 \gg 1$:

$$\tilde{\theta}_T = \text{value of } \theta_T$$

which maximizes

$$\sum_{i=1}^{K} v_i\sqrt{f\left(\frac{\theta_i-\theta_T}{\beta}\right)}.$$

The joint maximum likelihood estimation of X_0, θ_T is introduced because it can be shown that $\tilde{\theta}_T$ thus obtained is an "asymptotically efficient" estimate of θ_T; *i.e.*, the ratio $\sigma_{\min}^2/\sigma^2(\tilde{\theta}_T)$ approaches unity as $N \to \infty$ (for fixed β). Thus, at least for large enough N, an estimate of target position can be made for which the standard deviation closely approaches $\sigma_{\min}$, assuming only receiver noise to be present.

Now let $\sigma_{\min}'$ be the lower bound for $\sigma(\hat{\theta}_T)$, for all unbiased estimates $\hat{\theta}_T$, derived by applying the results of Cramer[4] for the case where X_0 is known *a priori*. In the previous section, the result was obtained that under the assumptions used in evaluating $\sigma_{\min}$, $\sigma_{\min}' \approx \sigma_{\min}$. The reason for this (other than that the mathematics comes out with this result) can now be seen.

From (34) and (35) it can be seen that $\tilde{\theta}_T$ depends mainly on the range within which $\tilde{X}_0$ falls, and is relatively insensitive to the exact value of $\tilde{X}_0$. In fact, if $\tilde{X}_0$ falls with high probability in the correct range of values (for example if, with high probability, $\tilde{X}_0 \gg 1$ whenever the true value $X_0 \gg 1$) then $\tilde{\theta}_T$ obtained from joint maximum likelihood estimation of X_0 and θ_T will have nearly the same value as the maximum likelihood estimate of $\tilde{\theta}_T$ obtained for X_0 known *a priori*.

It is reasonable to suppose that the result $\sigma_{\min}' \approx \sigma_{\min}$ depends rather strongly on the assumptions used in evaluating $\sigma_{\min}$—mainly, the assumption of a symmetric beam pattern and N not too small.

The maximum likelihood estimate of θ_T has been mentioned here mainly to bring out the fact that, under certain conditions, the lower bound $\sigma_{\min}$ can be approximately attained. A detailed analysis of maximum likelihood angular position estimators is to be found in a paper by Bernstein.[10] He deals, however, only with the case where the target is assumed to have a rapidly Rayleigh fluctuating cross section. An analysis following the same lines could be made for the case of a nonfluctuating target.

One of the most interesting possibilities for further work along these lines is the analysis of various types of more easily mechanized angular position computers, *e.g.*, computers based on the center of gravity or on the peak of the pulse envelope, to see how nearly they approach the theoretical optimum.

IV. Decision as to Whether a Target Actually Is Present

The discussion has centered thus far around the problem of estimating the target's angular position on the assumption that a single target is present in any given range interval. In a practical system it would be necessary also to decide whether or not a target actually is present at the given range, or, to put it another way, whether any significance should be attributed to the result of the position estimation.

A black box, the function of which would be to make such a decision, *i.e.*, a target detector, can be considered either as operating separately from, and in parallel with, the angular position computer; or as being to a greater or lesser extent combined in the same circuitry.

One method, for example, of utilizing for detection purposes the output of a maximum-likelihood angular position computer would be as follows: the maximum-likelihood computer maximizes $\mu(\theta_T, X_0)$, as given by (34), with respect to θ_T, X_0. The maximum value is $\mu(\tilde{\theta}_T, \tilde{X}_0)$. This maximum value could then be subjected to a threshold test: a target would be declared present if $\mu(\tilde{\theta}_T, \tilde{X}_0)$ exceeds a preassigned threshold. (This is of course just one of a multitude of possible procedures.)

The various estimation and detection processes discussed thus far would not be applicable to all possible situations arising in radar detection and position estimation. For one thing, the above results depend on the *a priori* assumption that at most one target per range interval can be present in the sector A at any given time. In many cases, however, one could get around this limitation by, for example, the following procedure: by conventional detection methods, the presence and approximate position of each target in the sector A could be established. Then, if there were more than one target present in a given range interval, and if these targets were not too close together in angular position, the sector A could be divided into subsectors, each containing just one target. The more precise estimate of target position could then be made for each subsector. If the targets were separated by several beamwidths, the maximum accuracy of position estimation would not be significantly reduced, as compared with the case of only one target per range interval, since the main contribution to the position estimation is made by signal returns occurring when the target is near the central portion of the beam. For most beam-shapes, any sector which is several times as wide as the beam is for these purposes effectively infinite.

The results which have been obtained do not give any information about the degree to which targets at the same range, and separated by less then, say, one beamwidth, can be resolved. The problem of target resolution is, however, undoubtedly amenable to treatment by the same general methods.

Fig. 2 can be used in conjunction with probability of detection curves to provide information as to how accurately θ_T can be theoretically determined, if signal-to-noise ratio is sufficient to give some definite probability of detection.

An example of how this could be done is given in Figs. 3 and 4. Fig. 3 treats the case of a nonfluctuating target, and Fig. 4 that of a rapidly fluctuating target.[11] In these figures, the probability of detection curves were taken from Marcum[12] and Swerling.[13] These curves correspond to a detection process using a square law second detector, followed by addition of N_D pulses, the sum being required to exceed a voltage threshold.

In Figs. 3 and 4, $f(u)$ is taken to be e^{-u^2}. It is assumed that N_D, the number of pulses integrated for detection, is the number of pulses emitted during the time it takes for the beam to turn through an angle equal to the width between the one-way half power points. For $f(u)=e^{-u^2}$, the width between the one-way half power points is 2.35β. Since the parameter N was defined as $2\beta/\Delta\theta$, we have $N_D=1.18N$. This was the value used for the number of pulses integrated for detection, in deriving the detection curves of Figs. 3 and 4 from those of Marcum[12] and Swerling.[13]

Also, a beam shape loss of 1.6 db was assumed for the detection curves in Figs. 3 and 4. That is, the value of signal-to-noise ratio used in the curves[12,13] was taken to be 1.6 db less than X_0, the signal-to-noise ratio at the nose of the beam.

The detection curves were plotted for false alarm number, n, equal to 10^8. False alarm number, n, is defined as follows: n = false alarm time divided by pulse width (assuming the IF bandwidth to be roughly the reciprocal of pulse width). The false alarm time is defined to be the time in which the probability of occurrence of at least one false alarm is 50 per cent.

[10] Robert Bernstein, "An analysis of angular accuracy in search radar," 1955 IRE Conv. Record, part 5, pp. 61–78.

[11] $\sigma_{\min}$ for a rapidly fluctuating target is evaluated in Section V.

[12] J. I. Marcum, "A statistical theory of target detection by pulsed radar," The Rand Corp., Res. Memo. RM-754; December 1, 1947.

[13] P. Swerling, "Probability of detection for fluctuating targets," The Rand Corp., Res. Memo. RM-1217; March 17, 1954.

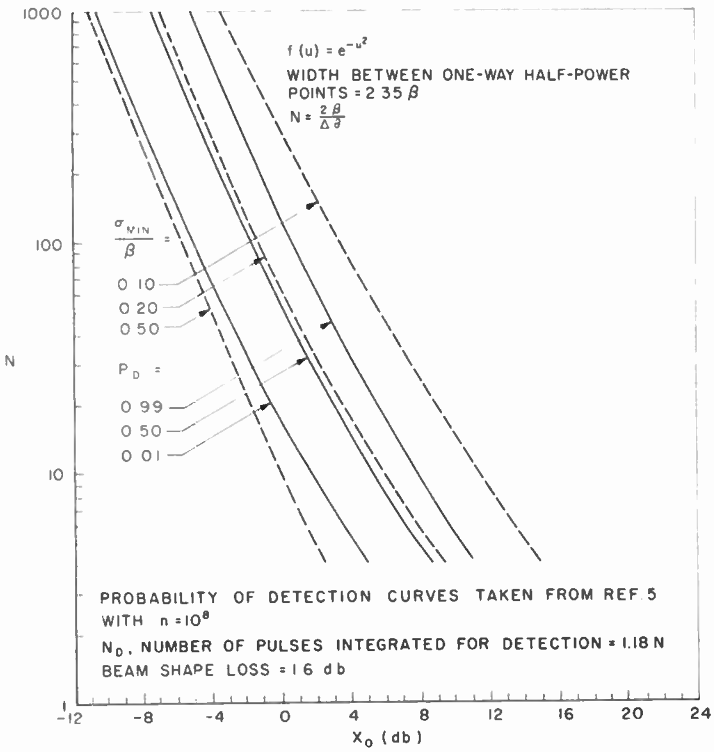

Fig. 3—Comparison of $\sigma_{\min}$ with probability of detection P_D for nonfluctuating targets.

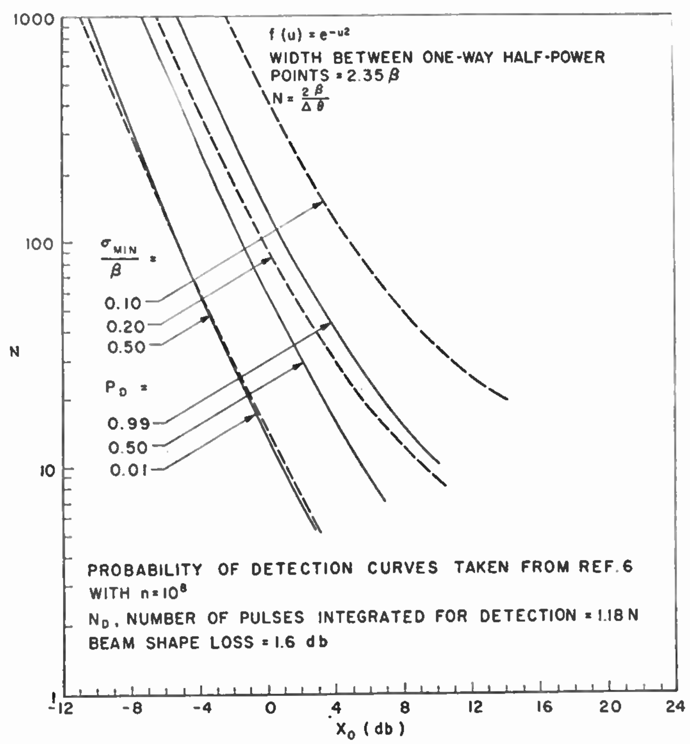

Fig. 4—Comparison of $\sigma_{\min}$ with probability of detection P_D for rapidly Rayleigh fluctuating targets.

V. A Lower Bound for $\sigma(\hat{\theta}_T)$—Rapidly Fluctuating Targets

It is almost obvious that the theoretical limits imposed on angular accuracy will be greatly affected if noise sources other than receiver noise are taken into account. As an illustration of this, a theoretical lower bound $\sigma_{\min}$ for the standard deviation of unbiased estimates will now be derived assuming the presence of both receiver noise and a particular type of fluctuation of the target cross section.

It will be assumed that the radar cross section $\sum$ of the target fluctuates according to the probability density function

$$p(\sum, \overline{\sum}) = \frac{1}{\overline{\sum}} \exp \frac{-\sum}{\overline{\sum}} \tag{36}$$

where $\overline{\sum}$ is the average of $\sum$ over the target fluctuations. The signal-to-circuit-noise power ratio of a pulse at any given beam position will fluctuate according to the same density function. It is assumed that the fluctuations are so rapid that the instantaneous cross section of the target is statistically independent from pulse to pulse.

The fluctuation expressed by (36), the so-called Rayleigh fluctuation, is the most commonly assumed type of fluctuation for certain types of radar targets, such as aircraft.

The assumption that the fluctuations are statistically independent from pulse to pulse is not very realistic in most practical cases; the analysis is carried out for this case because 1) the mathematical evaluation of $\sigma_{\min}$ is easy for this case, and 2) it serves well enough to illustrate the fact that the lower bound $\sigma_{\min}$ can be altered significantly if sources of noise other than circuit noise are considered.

Let x_i now represent the *average* over the target fluctuations of the signal-to-noise ratio for the ith pulse and let X_0 represent the *average* over the target fluctuations of the signal-to-noise ratio for a pulse emitted at $\theta = \theta_T$.

The basic system block diagram is taken to be the same as in Fig. 1, except that for the sake of mathematical convenience, it will be assumed that the second detector is a square law envelope detector.[14] Then, the joint probability density function for the voltages $v_1, \cdots, v_K$ emerging from the second detector is[13]

$$L(v_1, \cdots, v_K \mid \theta_T) = \int_{(\Delta\theta)_1} \prod_{i=1}^{K} \frac{1}{1 + x_i} \exp\left[\frac{-v_i}{1 + x_i}\right] d\theta_1 \tag{37}$$

where

$$x_i = X_0 f\left(\frac{\theta_i - \theta_T}{\beta}\right). \tag{38}$$

Also let

$$\lambda(v_1, \cdots, v_K \mid \theta_T) = \log_e L(v_1, \cdots, v_K \mid \theta_T). \tag{39}$$

[14] As pointed out in the explanation of Fig. 1, any monotonic detector law could be assumed without affecting the results.

If the same assumptions are made as in Section II, the quantity $\sigma_{\min}$ (9) can be evaluated exactly as in Section II. The result of such an evaluation is as follows:

$$E\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2\right\} \approx \frac{X_0^2}{\beta^2}\sum_{i=1}^{K}\frac{1}{(1+x_i)}f_i'^2. \tag{40}$$

Approximating this sum by an integral, one obtains

$$E\left\{\left(\frac{\partial\lambda}{\partial\theta_T}\right)^2\right\} \approx \frac{NX_0^2}{2\beta^2}\int_{-\infty}^{\infty}\frac{f'^2(u)}{[1+X_0f(u)]^2}du. \tag{41}$$

Also,

$$\rho \approx 0 \tag{42}$$

so that

$$\sigma_{\min} \approx \frac{\beta}{X_0\sqrt{N}}\frac{1}{\sqrt{\frac{1}{2}\int_{-\infty}^{\infty}\frac{f'^2(u)}{[1+X_0f(u)]^2}du}}. \tag{43}$$

For this case, $\sigma_{\min}$ is depicted by the dotted line in Fig. 2.

Eq. (43) is, however, not applicable for indefinitely large values of X_0. The reason is that for sufficiently large X_0, the finiteness of the sector A cannot be ignored, and one is not justified in extending the integral to infinity. The extension of the integral to infinity is, for most beam shapes, justified, provided

$$X_0f\left(\frac{\theta_i-\theta_T}{\beta}\right) \ll 1$$

for θ_i on the edges of the sector A. This will be true in most cases of interest.

Eq. (43) is also not applicable for indefinitely small X_0 (*i.e.*, for X_0 such that $\sigma_{\min} \approx$ width of A); this is true for the same reasons as were cited in Section II.

Comparing the results for the nonfluctuating and rapidly fluctuating targets, it is seen from Fig. 2 that, while $\sigma_{\min}$ goes to zero inversely as $\sqrt{X_0}$ for the nonfluctuating case, it decreases more slowly (for large X_0) in the rapidly fluctuating case. In fact, if the finite width of the sector A were taken into account, it would approach a nonzero asymptote.

This situation is a result of the particular target fluctuation model assumed in (36). In the nonfluctuating case, circuit noise is assumed to be the only thing which distorts the pattern of the returned pulses. The distortion caused by circuit noise can be indefinitely reduced by increasing X_0. In the rapidly fluctuating case, both circuit noise and target fluctuation distort the pattern. For the type of fluctuation assumed in (36) the standard deviation of the instantaneous signal-to-noise ratio is proportional to X_0, so that, roughly speaking, the target fluctuations cause about the same amount of distortion no matter how large X_0 becomes. If X_0 is large, a further increase in X_0 would only cause a small diminution in $\sigma_{\min}$, attributable to overcoming the effect of circuit noise on the edges of the beam.

The form of the maximum likelihood computer for the rapidly fluctuating case can readily be determined. $\tilde{\theta}_T$, $\tilde{X}_0$ are approximately the values which maximize

$$\mu(\theta_T, X_0) = -\sum_{i=1}^{K}\log_e\left[1+X_0f\left(\frac{\theta_i-\theta_T}{\beta}\right)\right] - \sum_{i=1}^{K}\frac{v_i}{1+X_0f\left(\frac{\theta_i-\theta_T}{\beta}\right)}. \tag{44}$$

Bernstein has made a detailed analysis of the maximum likelihood computer for the rapidly fluctuating case.[10]

Extension of this type of analysis to other types of target fluctuation would be an interesting subject for further research. It is probable, for example, that if the correlation time of the target fluctuations is of the order of, say, half the time on target of the beam, the theoretical lower bound for the standard deviation would be more severely affected than for the rapid fluctuation treated here.

VI. Application to a Hypothetical Search Radar

As an illustration of how to apply the foregoing results, consider a search radar with the following system parameters:

Peak transmitted power = 500 kilowatts.
Pulse length = 1 microsecond.
Antenna gain at nose of beam = 10^3.
Wavelength = 10 centimeters.
IF bandwidth = 1 megacycle.
Receiver noise figure = 16db.
Losses (atmospheric transmission loss, maintenance degradation loss, etc.) = 6 db.
Horizontal beamwidth (between one-way half power points) = 3.54 degrees.
Angular sector scanned (horizontal) = 360 degrees.
Scan rate = 10 rpm.
Pulse repetition rate = 500 pps.
Beam shape = Gaussian.

Suppose the target in question has a radar cross section of 10 square meters, and is at a range of 40 nautical miles. From the radar range equation, it can be readily determined that X_0, the power signal-to-noise ratio at the input to the second detector when the target is at the nose of the beam, is 1.0 db.

To apply the results given in Fig. 2, take $f(u) = e^{-u^2}$. Then $2.35\beta = 3.54$ degrees, or $2\beta = 3.54/1.18$ degrees $= 3.0$ degrees. The beam turns through 3 degrees in 1/20 of a second. Thus $N = 500/20 = 25$. Assuming a nonfluctuating target, one reads from Fig. 2 that

$\sigma_{\min}\sqrt{N}/\beta = 1.44$ for $X_0 = 1.0$ db. Since $\sqrt{N} = 5$, and $\beta = 1.5$ degrees, one finally get $\sigma_{\min} \approx .43$ degrees.

The results embodied in Figs. 3 and 4 can be applied as follows. Suppose one wishes to answer the following question: suppose the various system parameters, target cross section, and target range are such that the probability of detection for a single scan is, say, 50 per cent. What is $\sigma_{\min}$ in such a case?

Assuming a nonfluctuating target and a Gaussian beam pattern, one sees from Fig. 3 that $\sigma_{\min}/\beta \approx 0.21$. Since Fig. 3 assumed $f(u) = e^{-u^2}$, the beamwidth, if defined to be the width between one-way half power points, is equal to 2.35β. Thus,

$$\frac{2\sigma_{\min}}{\text{beamwidth}} \approx .18.^{15}$$

[15] It should be mentioned that the ratio of $2\sigma_{\min}$ to beamwidth is a more interesting parameter than, say, the ratio of $\sigma_{\min}$ to beamwidth. The reason for this is that, roughly speaking, for an angular position estimator with standard deviation σ, the uncertainty in angular position lies between $-\sigma$ and $+\sigma$. Thus the ratio of total angular uncertainty to beamwidth is more accurately represented by the ratio of 2σ to beamwidth than by the ratio of σ to beamwidth.

In the opinion of the author, the application of the results in this form is much more meaningful than the calculation of a definite value of $\sigma_{\min}$ for definite values of the system and target parameters. The reason is that there are usually great uncertainties in one's knowledge of some of the parameters such as target cross section and the various "losses"; such uncertainties have a great effect on the resulting calculated value of X_0, and hence of $\sigma_{\min}$. The same uncertainties enter into the the calculation of detection range, but the calculated value of detection range is less severely affected because of the fourth power law.

Since, however, uncertainties in the knowledge of the various parameters affect the calculation of probability of detection in almost exactly the same way as they affect the calculation of $\sigma_{\min}$, the statement that "the ratio of $\sigma_{\min}$ to beamwidth is a certain amount if the various system parameters are such as to give a probability of detection of a certain amount" is much more accurate than a statement of the absolute value of $\sigma_{\min}$ for a specified radar and target.

An 8-mm Klystron Power Oscillator*

R. L. BELL† AND M. HILLIER†

Summary—**The development of a cw klystron oscillator as a low-noise transmitter for the 8-mm band is described, and details are given of its performance. The power in the electron stream is 350 watts (0.1 ampere at 3,500 volts) and the output power 12 w. Special features include a design for a bakeout temperature of 700°C, using molybdenum-sealing aluminosilicate glass, and a high convergence electron-optical system, permitting the use of a sprayed-oxide cathode with 1000-hour life.**

INTRODUCTION

THE TUBE described in this paper was the outcome of work on a low-noise transmitter for a cw radar in the 8-mm band. Certain advantages may result from the use of cw rather than the more usual pulsed operation. The rate at which information is gathered under some conditions may be improved, or problems of presenting the information may be simplified. The property utilized is the long-term phase stability or *coherence* of the transmitted signal, discarded when an oscillator is pulsed. This quality of coherence is of course lost eventually, even with cw operation, because of random fluctuations in phase which beset every oscillator. Phase stability then is a criterion of fundamental importance for the present tube. At the outset of the work it seemed likely, on admittedly intuitive grounds, that the required stability would be more easily achieved in a klystron than in a cw magnetron of the same power output.

The vehicle finally chosen for the work was the floating-drift-tube oscillator[1] offering the efficient working of the two-cavity klystron but, having a single cavity, avoiding the inconvenience of having to tune one cavity to another. The present tube is without tuning. In order to make the tube useful as an experimental transmitter, a power output of the order of 10 w was aimed at.

The development of a millimeter klystron to this order of power raises the following problems.

A dc electron stream of some tens of amperes per square centimeter is required but no known cathode reliably approaches this in emission density within an order of magnitude. The cw emission density of the oxide cathode is inadequate by a factor of about 100, and the use of such a cathode therefore demands an

* Original manuscript received by the IRE, April 16, 1956; revised manuscript received, June 8, 1956.
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[1] M. Chodorow and S. P. Fan, "A floating-drift-tube klystron," Proc. IRE, vol. 41, pp. 25–31; January, 1953.

accurately aligned converging electron gun. Given a stream of this current density, the minimum operating voltage turns out to be several thousand volts and the dc power flow in the stream is thus an appreciable fraction of a megawatt per square cm. Since water-cooled copper anodes are known to melt under bombardment at one or two kilowatts/square cm it is necessary to protect the drift tube and cavity against so formidable a stream.

Aside from these difficulties, it is to be expected, from the high phase sensitivity of a klystron oscillator to electron stream potential, that space-charge neutralization of the electron stream by positive ions will lead to undesirably large phase fluctuations, through fluctuations in the numbers of ions present, and their distribution. Hence the whole tube, not excluding the cathode, must be processed so as to yield the highest possible vacuum under the intensely adverse conditions expected.

The possible use of the tube in experimental equipment meant that its microphonics should be as low as possible and its useful life some hundreds of hours at least.

Electrical Design

The main problem in designing a power tube at short wavelengths is to bring the maximum amount of current to interact with the rf fields at the lowest operating voltage, without burning out the structure supporting the field. The conventional solid cylindrical system usual with klystrons was adopted, although there may be advantages in flat strip or hollow cylindrical systems, so far insufficiently developed.

Electron Gun

The characteristics of the tube, voltage, current, and efficiency then turn on the types of electron gun available, and the emission density can then be extracted from available cathodes. Consideration shows that when emission is at a premium, high perveance may actually be an embarrassment. What is needed is a gun of modest perveance but a high ratio of stream current density to cathode emission density. The low electron stream resistance required is then obtained by working at a high voltage, thus easing the structure design and promoting a high power flow in the stream.

At the time of the development the only readily available cathode was the sprayed oxide cathode, with a cw emission of under 1 a/square cm. A gun was developed from the work of Pierce and others,[2,3] having a current density multiplication of about 100 times and a perveance of 0.5 10^{-6} a/$v^{3/2}$. With a cathode loading of 0.6 a/cm² it gives a working stream current density of about 50 a/cm². At the working voltage of 3,500 v, the power density in the electron stream is 175 kw/cm².

[2] J. R. Pierce, "Theory and Design of Electron Beams," D. Van Nostrand Co., Inc., New York, N. Y., 1949.

[3] K. Spangenberg, R. Helm, and L. M. Field, "Cathode design procedure," *Elec. Commun.*, vol. 24, pp. 101–107; March, 1947.

Cavity

Given the stream, an acceptable design for the required power can be calculated in advance with confidence. Trials and modifications of this design led to a tube with the parameters shown in Table I.

TABLE I

Frequency	34,000 mc
Electron stream voltage	3,500 v
Useful stream current	100 ma
Cathode loading	0.6 a/cm²
Drift length	0.104 inch (17.3 radians)
Tunnel diameter	0.023 inch (4.0 radians)
Gap lengths	0.010 inch (1.7 radians)
Unloaded cavity Q	1,800
Loaded cavity Q	500
Power output	12 w
Efficiency	3½ per cent
Electronic tuning range	30 mc

Efficiency

A notable feature of this tube as a power tube is its low efficiency. It seems unlikely that this could be much improved by changing the circuit arrangements—gap spacings, drift length, gap voltage ratio, and so on—which are about optimum for the stream used.

It can be shown that the low efficiency is due basically to the high copper losses encountered at millimeter wavelengths. These vary with the square of the rf fields (or in cases where rf defocusing bombardment is present, with some higher power than the square) and impose operating conditions on the tube which are far from optimum for the generation of rf power in the stream or its extraction. The optimum output gap voltage, for example, turns out to be much less than the stream voltage, so that only a fraction of the rf power available in the stream is in fact removed, the rest going to heat up the collector.

The remedy is to increase the stream intensity, which can only be done in this case by using a cathode with higher emission, a gun with a higher density multiplication, or a higher working voltage.

Construction

Such low efficiency in a power tube entails getting rid of large quantities of unwanted heat, and this in turn implies a solid metal construction with air or water cooling in good thermal contact with the sources of heat. In order to outgas such a structure on the pump it has to be baked out for a certain time at the highest possible temperature, usually determined by the collapse of any glass present. The time taken may be cut by a useful factor (1,000 to 10,000 times) by raising the baking temperature from the usual 400–500°C to 700 or 800°C.

The high melting properties of the aluminosilicate glasses were exploited in order to achieve this. These glasses seal to molybdenum and some work[4] was done on deep drawing of 0.010-inch molybdenum sheet as a means of fabricating the required shapes for sealing. The glass to metal seals can be seen in Fig. 1(a) and (b).

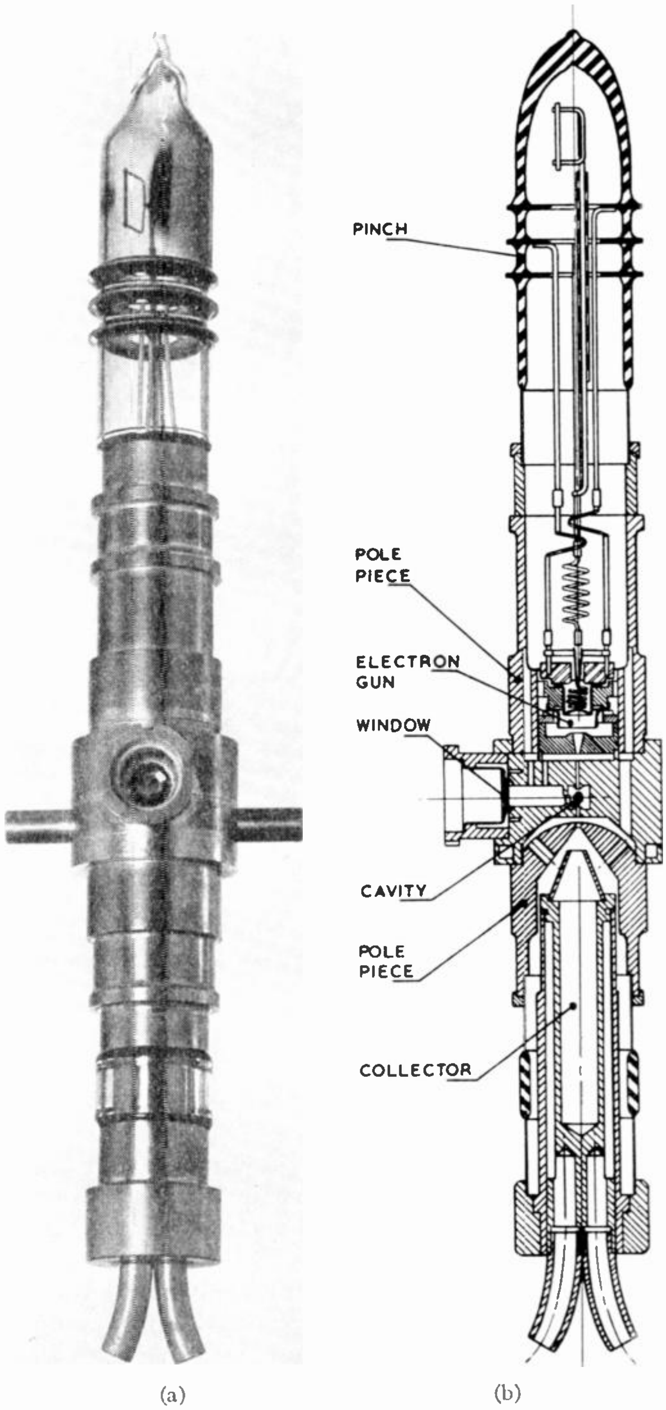

Fig. 1—(a) Photograph and (b) Cross section of 8-mm klystron.

They are: A stack of 3 disc seals for the "pinch": an 0.008-inch slip of glass sealed over a 5 mm hole in the end of a cup, for the window, and cylindrical butt-seals, on 20-mm diameter, elsewhere.

[4] F. Duckworth, "Deep drawing of molybdenum," *Machinery*, vol. 84, pp. 389–390; February, 1954.

The seals are made in a protective atmosphere of dry argon, by eddy-current heating the metal to 1200°C, when the glass flows, wets the metal and makes the seal. The molybdenum should preferably be slightly oxidized to start with. The glass is annealed by cooling slowly from 800°C in hydrogen, which process also reduces any oxide remaining on exposed metal, thus facilitating chemical cleaning of the molybdenum for plating at a later stage. Plating with nickel or copper makes it possible to silver solder the finished parts into the tube. No water penetration of the glass-metal seals is observed.

The tube [Fig. 1(b)] comprises a water-cooled cavity block and window assembly of OFHC copper with steel pole pieces silver brazed in either side, one housing the electron gun and the other a water-cooled collector. All joints are made by silver-copper brazing, being designed to come under compression when cold. It is advisable to have excess copper at every joint, providing this with a copper collar where necessary. Using this precaution, the whole structure, including the glass-to-metal seals, has given little trouble from leaks.

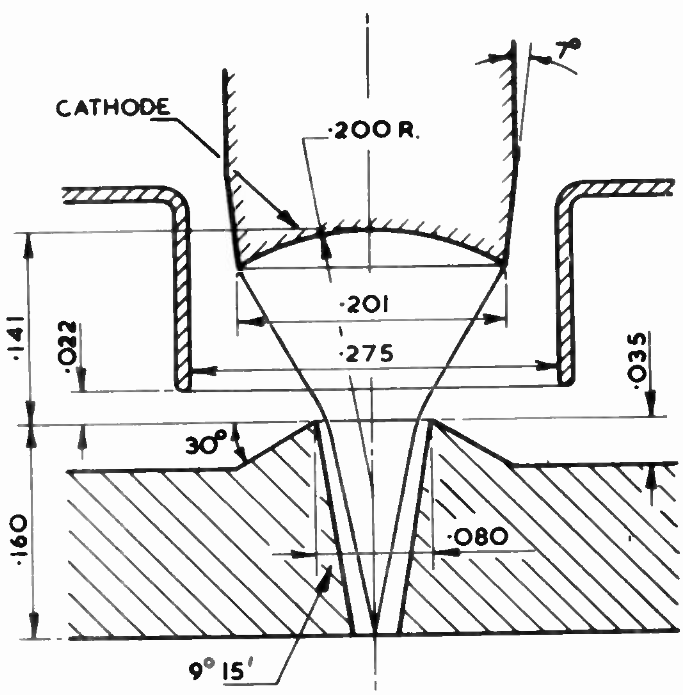

Fig. 2—Electron gun design for 8-mm klystron. All dimensions are in inches.

Electron Gun

Fig. 2 gives the essential dimensions of the electron gun. The stream starts out as a converging cone of a 30° half angle, is refracted by the diverging lens of the anode aperture, and finally converges to a crossover with a 10° half angle. The crossover diameter is determined by space charge, thermal spread, and lens aberrations in roughly equal proportions. More than 90 per cent of the current in the crossover is included within a diameter 1/10 that of the cathode. The location of the crossover along the gun axis can be adjusted within limits by varying the bias on the focusing cylinder surrounding the cathode, so compensating for tube to tube variations in gun assembly. The cylinder runs negative to the cathode by about one-fifth the voltage on the cathode.

In assembly the gun is spring-loaded against its location on the cavity block so that each piece-part transmits the compression to its neighbor. Hence, although their mutual fit may not be perfect over the range of operating temperatures, yet the parts are prevented from vibrating mechanically with respect to the cavity. This feature eliminates microphonics due to the gun structure.

Once formed, the stream is injected into an axial magnetic field of 3000 gauss. This is an optimum value, above which the transmission begins to fall off. It is about twice the *Wang value*[2] for the stream, the excess being partially compensated by a small leakage flux on the cathode.

The tunnel in the cavity block is 0.6 mm in diameter. The first few millimeters of its length are used for cleaning up the edges of the stream before injecting it into the drift space. In this region the stream encounters a gradually increasing magnetic field, which removes it from the vicinity of the tunnel wall and counteracts the scattering action of the rf fields. The magnetic field goes on increasing along the length of the drift tube and up to the collector-end pole piece, where it is 5500 gauss. The stream passes through a hole in the pole piece, spreads out in a field-free region, and is collected.

The nonuniformity of the field, obtained by shaping the collector-end pole piece, is found to be essential to the operation of the tube above the few hundred milliwatt level. Tubes without this feature run at very low efficiency, with the drift tube at bright red heat.

A typical tube with the shaped field in operation at 3500 v transmits 90 ma through to the collector and intercepts 30 ma elsewhere, 5 to 10 ma of the interception being due to rf defocussing.

Resonant Cavity

A sketch of the resonant cavity is shown in Fig. 3 and the essential dimensions are given in Table I. The gap dimensions will be recognized as being too large for good coupling to the beam, by longer wavelength standards. This makes an additional contribution to low efficiency, imposed on the design by the limited current density that can be attained in the stream. The shapes of drift tube and gaps represent an attempt to minimize the hot copper losses, that is to say, a compromise between minimum losses when cold and robust construction with good heat dissipation properties under bombardment.

Although a small increase in power might result from adjusting the dimensions of the input gap, the arrangement as it stands with identical gaps is advantageous from the point of view of microphonics: a first-order displacement of the drift tube in any direction generates no (first-order) change in the resonant frequency.

Machining tolerances are at their finest in the cavity, where the gap lengths are held to within 0.0002 inch. The tube to tube spread in resonant frequency is of the order of ± 1 per cent.

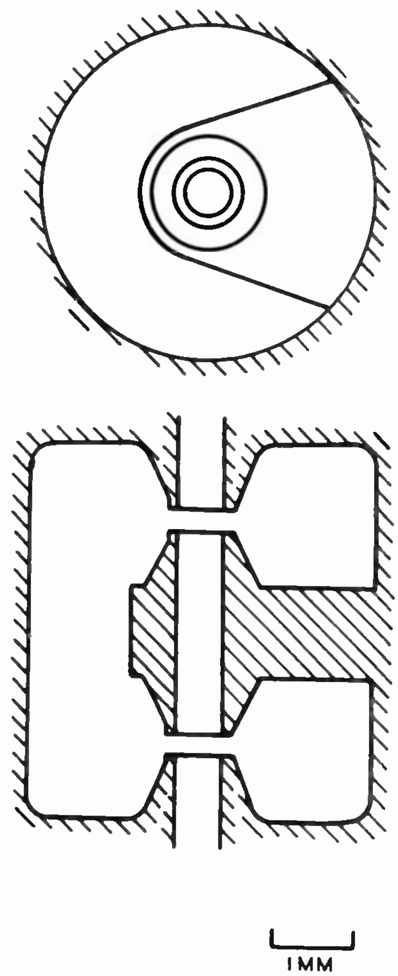

Fig. 3—8-mm resonant cavity with floating drift tube.

After assembly the cavity block is oxidized at 650°C in a current of air and reduced in hydrogen. This is found[5] to raise the surface conductivity of the cavity from around 50 per cent in the as-brazed condition to around 95 per cent of the theoretical maximum value for copper at 8-mm wavelength, and has a profound effect on the power output of the tube.

Cold measurements on the cavity usually indicate unloaded Q values between 1700 and 2000.

Processing

The use of aluminosilicate glass envelope materials potentially raises the bakeout temperature to 800°C or higher. Other considerations—softening of the springs

[5] J. S. Thorp, "R.F. conductivity at 8 mm wavelength," *Proc. IEE*, vol. 107, Part III, pp. 357–359; November, 1954.

of the gun mounting, silver evaporation, *etc.*—limit the bakeout on the present tube to 700°C. At this temperature the tube is noticeably porous to most gases, and whereas this high porosity promotes outgassing of the tube, it also permits contamination by diffusion of atmospheric gases into the tube through the walls.

The tube is therefore baked in a vacuum furnace, being provided with a molybdenum heater, heat shields, and a vacuum bell jar, pumped to 10^{-5} mm Hg by a high speed oil pump. It is meanwhile exhausted independently on a fractionating oil pump.

After the bakeout the processing follows conventional lines—the cathode is flashed, the tube run up to power on the pump, sealed off by melting a constriction in the aluminosilicate glass, and a getter fired. Work is proceeding on cathode activation during the bake, squeezing off from a copper pumping stem, and elimination of the getter, all of which are expected to improve the vacuum and the life.

Performance

The average efficiency of tubes made to date is $3\frac{1}{2}$ per cent, giving an average power output of 12 w. Powers as low as 3 w and as high as 30 w have been observed. The aluminosilicate window handles higher powers than these without forced cooling.

The power output of a given tube is a maximum for an optimum set of values of electron stream voltage, magnetic field, focussing bias and output match. None of these is critical. Thus the magnetic field is provided by a commercial permanent magnet, in most cases without adjustment. Optimum power results when the load is matched to an admittance which contains the stream loading loss of the tube. As this is a variable quantity, a matching stub is sometimes necessary to obtain maximum efficiency.

The tube is voltage-tunable over about 30 mc between half-power points, with an electronic tuning sensitivity of 100 kc/v on the cathode voltage. Thermal tuning of the same order is also observed, having a time constant of 0.01 second.

Measurements have been made of the phase-modulation noise spectrum of the tube output as a function of the vacuum. At 10^{-5}-mm Hg, the noise level is high and the spectrum contains discrete oscillations of a relaxation character. The noise level and the presence of oscillations depend on the voltage difference existing between collector and cavity, with the occurrence of *modes*, reminiscent of reflex klystron operation, for both positive and negative collector polarities. At 10^{-6}-mm, the oscillations have died out, but the noise level is still high. At 10^{-7} or 10^{-8}-mm Hg the noise has decreased to a value of the order of what might be expected from basic sources like shot noise or flicker noise in the electron flow. Tubes normally settle down to low noise operation in the first few hours of life, and remain quiet up to the end of life. Amplitude-modulation fluctuations are below the limits of detectability.

End of life, which normally occurs at about 1000 hours, is marked by a fall of emission and power output, when the cathode is found to be practically bare of coating. A *take-apart joint* is provided at the gun end, and new life can be breathed into the tube by replacing the gun and reprocessing.

Tubes baked at 450°C in air, using the aluminosilicate construction or the conventional Kovar-borosilicate construction showed signs of rapid depreciation of the vacuum on running, and failed at about 50 hours by sputtering of the cathode. No noise measurements were made in these tubes but in view of the results above they were almost certainly noisy.

Microphonics in the output from the tube were found to be low; none were noticed during noise measurements although no special anti-vibration precautions were taken. No quantitative measurements have been made.

Conclusion

Production of a number of samples of the tube described has shown the feasibility of making a cw power klystron for the 8-mm band, of the 10-w caliber. Experience in designing and testing the tube indicates that with types of high emission cathode now becoming available a substantial increase in power should be reliably attainable.

The development has shown that an increase in bakeout temperature alone may have a profound effect on the running pressure of such a tube, and so on the cathode life and other properties which may depend on the vacuum.

Although the use of aluminosilicate glasses was originally intended to simulate the effect which ceramic envelope construction might have in raising the bakeout temperature, the technique has proved to have clear advantages over current ceramic techniques in the matter of simple technology and reliable end products. In applications where very low dielectric loss is not important these may well prove to be overriding advantages.

Noise measurements have shown that samples processed for high vacuum approach in stability the fundamental limits set by shot noise and kindred effects in the electron stream.

Acknowledgment

The authors are indebted to the British Admiralty for permission to publish this paper.

Restrictions on the Shape Factors of the Step Response of Positive Real System Functions*

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Summary—**Given the maximum size of any overshoot or undershoot of the step response of a network whose system function is positive real, a lower bound on the rise time from zero to the final value is developed. Similarly lower bounds on the settling time are also derived. These results are improvements over previously published results. They are special cases of a general theorem which bounds the unit step response, $A(t)$, for $0 \leq t < \tau$ when this response is bounded by $(1 \pm \gamma)r$ for $t \geq \tau$ where γ is a positive real number and r is the final value of the unit step response.**

Introduction

It has been shown in some recent papers[1,2] that the transient responses of various types of networks are bounded so that such networks are not suitable for the synthesis of those circuits whose desired responses fall outside these bounds. In many cases it is desired that the output of a system reproduce the shape of the input. If the input is a step the best response is another step. A passive driving point impedance having some shunting capacity, C, across its two terminals cannot produce such a response, for its rise time from zero to one is never less than rC where r is the value of this impedance at zero frequency. Moreover this lowest value for rise time occurs only when the overshoot is infinite. The question remains as to how much larger this lower bound on rise time must be if the peak overshoot or undershoot is specified. An answer to this problem is obtained in this paper. Furthermore, lower bounds on the settling time of such systems have previously been shown to exist.[1] This result is also improved. These conclusions are stated as corollaries to the main theorem.

Specifications of System Functions

Only lumped, linear, fixed, and finite systems are considered whose system functions, $Z(s)$, are positive real and have no poles on the real frequency axis or in the right half plane, no zero at the origin, and one more pole than zero. This last restriction is a consequence of the stray shunting capacities existing across any pair of terminals coupled with the inevitable dissipation in any physical circuit. The real parts of these functions are nonnegative along the real frequency axis. Expanding such a system function into infinite series, the following two expressions may be obtained. Eq. (1) holds in the neighborhood of $s=0$ and (2) holds in the neighborhood of $s=\infty$.

$$Z(s) = r + k_1 s + k_2 s^2 + k_3 s^3 + \cdots. \tag{1}$$

$$Z(s) = \frac{1}{Cs} + \frac{K_2}{s^2} + \frac{K_3}{s^3} + \cdots. \tag{2}$$

It is assumed that the input unit step function is applied at time, $t=0$, so that the unit step response is zero for negative values of time. The unit impulse response, $W(t)$, and the real part of the system function along the real frequency axis, $R(\omega)$, are related by the following Fourier cosine transforms.

$$W(t) = \frac{2}{\pi}\int_0^\infty R(\omega)\cos\omega t\,d\omega, \qquad t \geq 0. \tag{3}$$

$$R(\omega) = \int_0^\infty W(t)\cos\omega t\,dt. \tag{4}$$

Bounds on the Step Response of Positive Real System Functions

If bounds on the unit step response for the stated type of system are specified beyond the time, τ, then bounds on this response before this time may be obtained. An immediate consequence of this is that restrictions on the unit step response during the initial transition period exist and they become stronger as the magnitude of the greatest overshoot or undershoot decreases. The theorem stated subsequently presents this result precisely and Fig. 1 illustrates it for the case where the unit step response remains within ten per cent of its final value, r, after the normalized time, τ/rC. This response must then lie within the bounds indicated for time between 0 and τ/rC. It should be noted that the factor τ/rC cannot be chosen arbitrarily since values that are too small may not be realizable. The value of 1.36 which was used for this illustration was obtained from the unit step response of the Doba network.[3] Furthermore these bounds are not best possible and may be improved. This matter is considered after the proof of the theorem.

The proof of this result depends upon two inequalities which the sine function satisfies and which are stated by lemma 3. The first two lemmas are needed in the proof of the third. Their proofs appear in Appendix I.

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[1] A. H. Zemanian. "Bounds existing on the time and frequency responses of various types of networks," Proc. IRE, vol 42, pp. 835–839; May, 1954.

[2] A. H. Zemanian, "Further bounds existing on the transient responses of various types of networks," Proc. IRE, vol. 43, pp. 322–326; March, 1955.

[3] R. C. Palmer and L. Mautner, "A new figure of merit for the transient response of video amplifiers," Proc. IRE, vol. 37, pp. 1073–1077; September, 1949.

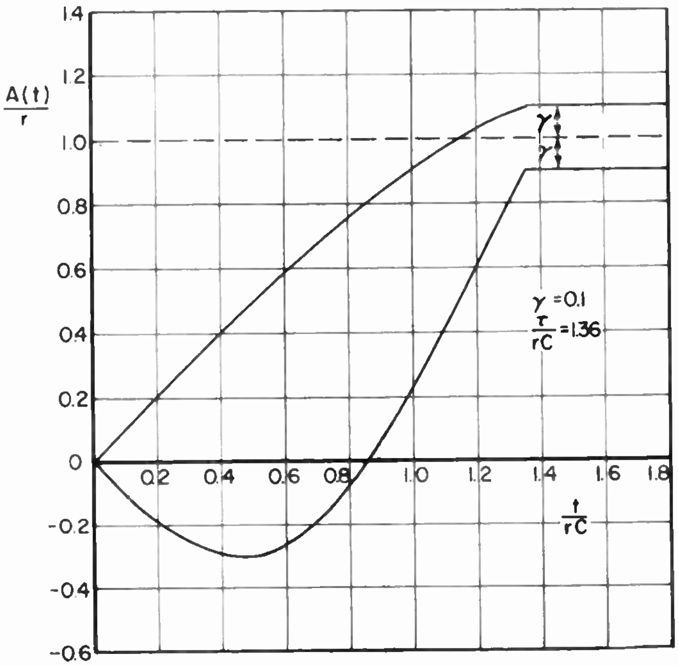

Fig. 1—Illustration of the bounds on unit step response as given in the theorem for the case where $\gamma=0.1$ and $\tau/rC=1.36$.

Lemma 1

If $v(u)$ is an even, periodic and integrable function whose period is $2\pi y$ where $0\leqq y<1$ and if $v(u)\geqq \cos u$ for $-\pi y\leqq u\leqq \pi y$, then

$$\int_0^x v(u)du \geqq \sin x \quad \text{for} \quad x\geqq 0.$$

Lemma 2

If $w(u)$ is an even, periodic and integrable function whose period is $2\pi y$ where $0\leqq y<1$ and if $w(u)\leqq \cos[u+\pi(1-y)]$ for $-\pi y\leqq u\leqq \pi y$ then

$$\int_0^x w(u)du \leqq \sin x \quad \text{for} \quad x\geqq 0.$$

Lemma 3

For $0\leqq y<1$, $x\geqq 0$ and N a positive integer

$$\begin{aligned}\sin x \leqq Q_0 x + yQ_2\sin\frac{x}{y} + \frac{yQ_4}{2}\sin\frac{2x}{y}\\ + \cdots + \frac{yQ_{2N}}{N}\sin\frac{Nx}{y}\end{aligned} \tag{5}$$

and

$$\begin{aligned}\sin x \geqq -Q_0 x + yQ_2\sin\frac{x}{y} - \frac{yQ_4}{2}\sin\frac{2x}{y}\\ + \cdots + (-1)^{N+1}\frac{yQ_{2N}}{N}\sin\frac{Nx}{y}\end{aligned} \tag{6}$$

where

$$Q_0 = 1 + \sum_{k=1}^{N}\frac{(-y^2)(1^2-y^2)\cdots[(k-1)^2-y^2]}{(k!)^2} \tag{7}$$

$$Q_{2p} = (-1)^p 2\sum_{k=p}^{N}\frac{(-y^2)(1^2-y^2)\cdots[(k-1)^2-y^2]}{(k-p)!(k+p)!} \tag{8}$$

$p = 1, 2, 3, \cdots, N$.

The main results of this paper are special cases of the following theorem.

Theorem

If the unit step response, $A(t)$ corresponding to the positive real system function which satisfies (1) and (2) is bounded by $(1\pm\gamma)r$ for $t\geqq\tau$ where γ is a positive number, then

$$\begin{aligned}A(y\tau) \leq r\frac{\sin\pi y}{\pi}\bigg\{\frac{\tau}{rC} + 2y^2\sum_{n=1}^{\infty}\frac{(-1)^{n+1}}{n(n^2-y^2)}\\ + \gamma 2y^2\sum_{n=1}^{\infty}\frac{1}{n(n^2-y^2)}\bigg\}\end{aligned} \tag{9}$$

$$A(y\tau) \geq -r\frac{\sin\pi y}{\pi}\left\{\frac{\tau}{rC} - 2y^2(1-\gamma)\sum_{n=1}^{\infty}\frac{1}{n(n^2-y^2)}\right\} \tag{10}$$

where $0\leqq y<1$.

Proof

The Fourier transform that relates the real part of the system function along the real frequency axis, $R(\omega)$, to the unit step response, $A(t)$, is given by the expression

$$A(t) = \frac{2}{\pi}\int_0^\infty \frac{R(\omega)}{\omega}\sin\omega t d\omega, \qquad t\geqq 0. \tag{11}$$

Since $R(\omega)\geqq 0$, $A(t)$ is less than or equal to the function obtained by inserting the right-hand side of (5) in place of the $\sin\omega t$ in (11). Integrating the finite sum term by term

$$\begin{aligned}A(t) \leq Q_0\frac{t}{C} + y\left[Q_2 A\left(\frac{t}{y}\right) + \frac{Q_4}{2}A\left(\frac{2t}{y}\right)\right.\\ \left. + \cdots + \frac{Q_{2N}}{N}A\left(\frac{Nt}{y}\right)\right].\end{aligned} \tag{12}$$

This inequality holds for all N and therefore for $N\to\infty$. Furthermore the Q's may be summed[4] as follows when $N\to\infty$. In these summations F is the hypergeometric function.

$$Q_0 \to q_0 = F(x, -x; 1; 1) = \frac{\sin\pi y}{\pi y}$$

$$\begin{aligned}Q_{2p}\to q_{2p} &= (-1)^p 2\frac{(-y^2)(1^2-y^2)\cdots[(p-1)^2-y^2]}{(2p)!}\\ &\quad\cdot F(p+y, p-y; 2p+1; 1)\\ &= (-1)^{p+1}\frac{2y\sin\pi y}{\pi(p^2-y^2)}.\end{aligned}$$

$p = 1, 2, 3, \cdots.$

[4] A. Erdelyi, W. Magnus, F. Oberhettinger, and F. G. Tricomi, "Higher Transcendental Functions," McGraw-Hill Book Co., Inc., New York, N. Y., vol. 1, eqs. 2.8(46) and 1.2(8); 1953.

Moreover $|A(t)| \leqq t/C$ since $R(\omega) \geqq 0$ as has been shown in a previous article.[1] This means that the double infinite series represented by

$$\lim_{N\to\infty}\left[Q_0\frac{t}{C}+y\sum_{p=1}^{N}\frac{Q_{2p}}{p}A\left(\frac{pt}{y}\right)\right]$$

converges absolutely for $0 \leq y < 1$ as can be seen from

$$\lim_{N\to\infty}\left[Q_0\frac{t}{C}+y\sum_{p=1}^{N}\frac{Q_{2p}}{p}A\left(\frac{pt}{y}\right)\right]$$
$$=\left\{1+\sum_{k=1}^{\infty}\frac{(-y^2)(1^2-y^2)\cdots[(k-1)^2-y^2]}{(k!)^2}\right\}\frac{t}{C}$$
$$+y\sum_{p=1}^{\infty}(-1)^p\frac{2}{p}$$
$$\cdot\sum_{k=p}^{\infty}\frac{(-y^2)(1^2-y^2)\cdots[(k-1)^2-y^2]}{(k-p)!(k+p)!}A\left(\frac{pt}{y}\right).$$

Replacing each term by its absolute value and $A(pt/y)$ by pt/yC, a double series of positive terms is obtained which converges to the sum of the absolute values of the q_{2p} multiplied by t/C as was shown above.

$$\lim_{N\to\infty}\left[Q_0\frac{t}{C}+y\sum_{p=1}^{\infty}\frac{Q_{2p}}{p}A\left(\frac{pt}{y}\right)\right]$$
$$\leq\frac{t}{C}\left[\frac{\sin\pi y}{\pi y}+\sum_{p=1}^{\infty}\frac{2y\sin\pi y}{\pi(p^2-y^2)}\right]$$
$$=\frac{t}{C}\left[2\frac{\sin\pi y}{\pi y}-\cos\pi y\right].$$

Thus,

$$A(t)\leq q_0\frac{t}{C}+y\left[q_2A\left(\frac{t}{y}\right)+\frac{q_4}{2}A\left(\frac{2t}{y}\right)\right.$$
$$\left.+\frac{q_6}{3}A\left(\frac{3t}{y}\right)+\cdots\right].$$

Since the q_{2p} are positive if p is odd and negative if p is even for $0 \leqq y < 1$, (9) may be obtained by setting t equal to $y\tau$ and replacing $A(pt/y)$ by $(1+\gamma)r$ if p is odd and by $(1-\gamma)r$ if p is even. Similarly replacing the sin ωt in (11) by the right-hand side of (6) and letting N go to infinity yields

$$A(t)\geq -q_0\frac{t}{C}+y\left[q_2A\left(\frac{t}{y}\right)-\frac{q_4}{2}A\left(\frac{2t}{y}\right)\right.$$
$$\left.+\frac{q_6}{3}A\left(\frac{3t}{y}\right)-\cdots\right].$$

The double series obtained in this case can again be shown to converge absolutely. Each coefficient of $A(pt/y)$ in this expression is positive and so (10) may be obtained from it by setting $t=y\tau$ and replacing each $A(pt/y)$ by $(1-\gamma)r$. This completes the proof.

The condition that $A(t)$ achieves its upper and lower bounds, $(1\pm\gamma)r$, at arbitrarily large values of time was used in the above proof. Such behavior is impossible for the type of system considered here since the unit step response will approach its final value, r, as t becomes arbitrarily large. Thus the inequalities (9) and (10) are too strong. However, since the series

$$\sum_{p=1}^{\infty}q_{2p}\left|A\left(\frac{pt}{y}\right)\right|$$

converges quite rapidly, the values of $A(pt/y)$ for large p contribute very little to the series. Thus the possible improvement in (9) and (10) is probably very small so that these results are quite likely very close to being best possible.

Restrictions on Rise Time-Overshoot Pairs

The theorem yields a lower bound on the rise time from zero to one of the step response, T_γ, which depends on the overshoots and undershoots and becomes stronger as the overshoots and undershoots are restricted to being smaller. As shown in Fig. 2, if the unit

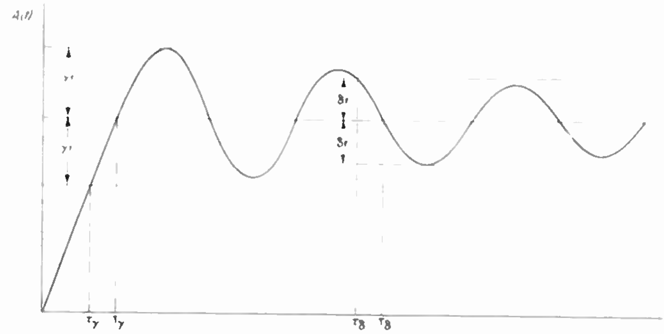

Fig. 2—Illustration of the shape factors for the unit step response.

step response is bounded by $(1\pm\gamma)r$ for $t \geqq T_\gamma$ where γ is the largest fractional overshoot or undershoot, then the rise time from zero to one must lie above the solid curve shown in Fig. 3. That is, the rise time-overshoot (or undershoot) pair for the unit step response of any positive real system function defines a point in the plane of Fig. 3 and this point must be above the solid curve. Furthermore this curve can be applied to find a lower bound on the time at which the unit step response crosses the final value line and beyond which it remains within the bounds $(1\pm\delta)r$ even though some of the overshoots and undershoots are greater than δr. Such a time, T_δ, is illustrated in Fig. 2. This result may be stated precisely as the following corollary.

Corollary 1

If the unit step response, $A(t)$, corresponding to the positive real system function which satisfies (1) and (2), is bounded by $(1\pm\gamma)r$ for $t \geqq T_\gamma$ where γ is a positive number and where $A(T_\gamma)=r$, then

$$T_\gamma > rC\left\{\frac{\pi y}{\sin \pi y} - 2y^3 \cdot\left[\sum_{n=1}^{\infty}\frac{(-1)^{n+1}}{n(n^2-y^2)} + \gamma\sum_{n=1}^{\infty}\frac{1}{n(n^2-y^2)}\right]\right\} \quad (13)$$

where $0 \leqq y < 1$.

It may be assumed that $A(y\tau)$ in (9) is such that $y\tau = T_\gamma$. This implies that $A(y\tau) = r$ and $(1-\gamma)r \leqq A(t) \leqq (1+\gamma)r$ for $t \geqq T_\gamma$. Replacing $A(y\tau)$ by r and τ by T_γ/y, (13) may be obtained after rearranging the result. Furthermore since $A(\infty) = r$, equality of the two sides of (13) can never be achieved.

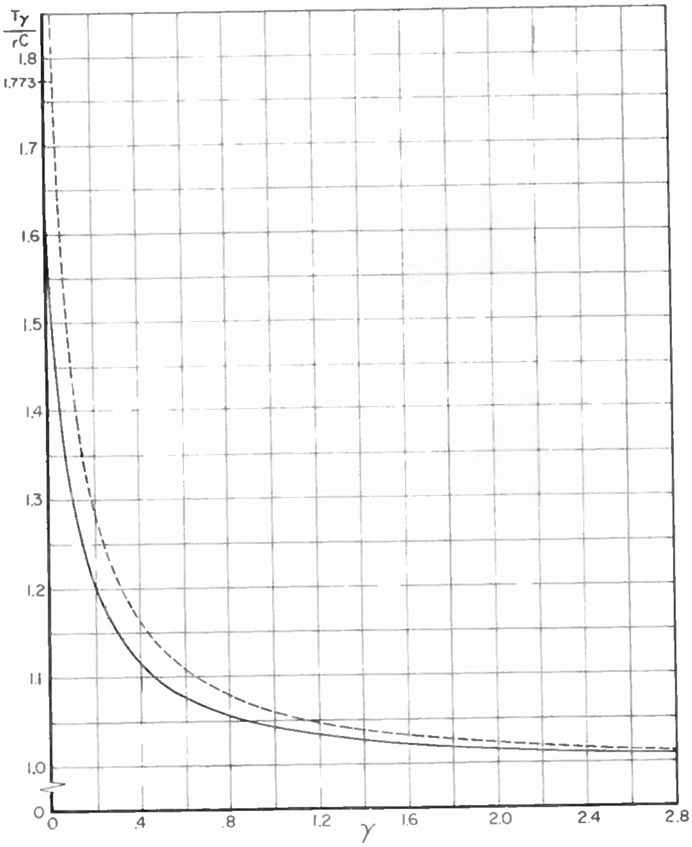

Fig. 3—The solid curve illustrates the lower bound on the rise time-overshoot pairs, and the dotted curve the rise time-overshoot pairs for the shunt peaked filter.

On the plane of Fig. 3, the right-hand side of the inequality (13) is the equation of a straight line whose position is a function of y. By varying y between zero and one, a family of straight lines may be obtained whose envelope is shown in Fig. 3 by the solid curve. As y approaches zero, the ordinate intercept of this straight line approaches one and its abscissa intercept approaches infinity. As y approaches one, the ordinate intercept approaches 1.773 and its abscissa intercept approaches zero. The rise time-overshoot pairs for the shunt peaked filter shown in Fig. 4 due to the variation of its damping is also indicated in Fig. 3 by the dotted curve. If the lower bound on T_γ is ever improved, the improvement cannot be greater, of course, than that indicated by this dotted curve. However, as stated before, this possible improvement is probably much less than that indicated by the difference between the two curves of Fig. 3.

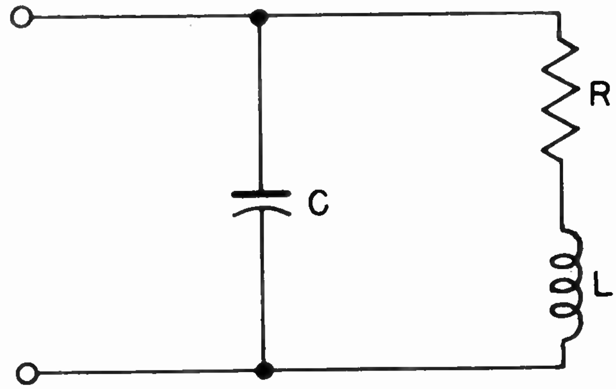

Fig. 4—The shunt peaked filter.

Restriction on Settling Time

Eq. (9) also yields a lower bound on the settling time for the unit step response of positive real system functions. Defining the settling time, τ_δ, as the least time beyond which the unit step response remains within the bounds $(1\pm\delta)r$, it is found that τ_δ must remain above the curve shown in Fig. 5. This settling time, τ_δ, and the

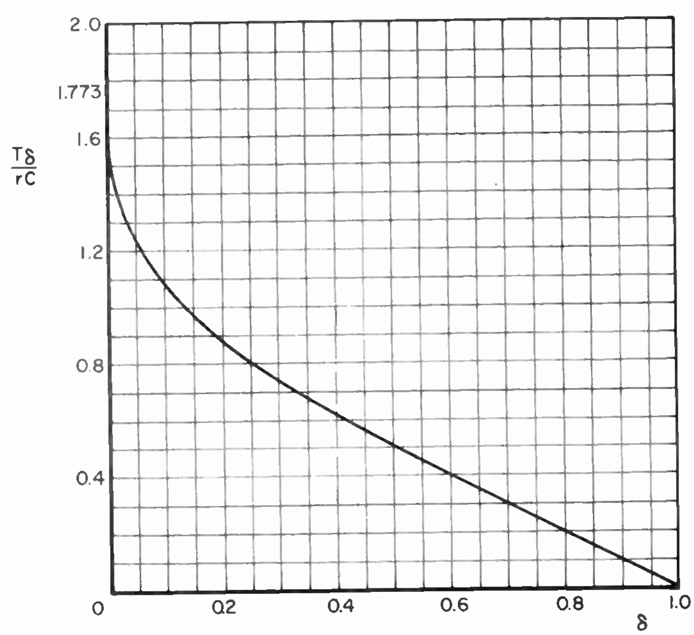

Fig. 5—A lower bound on the settling time of the unit step response.

one, τ_γ, corresponding to the weaker bounds, $(1\pm\gamma)r$ are illustrated in Fig. 2. The definition of settling time used here is different than the one used previously[1] and has the advantage of being defined in terms of the step response only. This result is stated precisely below.

Corollary 2

If the unit step response, $A(t)$, corresponding to the positive real system function which satisfies (1) and (2), is bounded by $(1\pm\delta)r$ for $t\geqq\tau_\delta$ where δ is a positive number between zero and one, then

$$\tau_\delta > rC\left\{\frac{\pi y}{\sin \pi y} - 2y^3\sum_{n=1}^{\infty}\frac{(-1)^{n+1}}{n(n^2-y^2)} - \delta\left[\frac{\pi y}{\sin\pi y} + 2y^3\sum_{n=1}^{\infty}\frac{1}{n(n^2-y^2)}\right]\right\} \tag{14}$$

where $0\leqq y<1$.

In this case, τ_δ is the least τ where τ is defined in the theorem. Assuming that $y\tau=\tau_\delta$, $A(y\tau)$ may be replaced by $(1-\delta)r$, τ by τ_δ/y and γ by δ in (9). Eq. (14) follows immediately upon rearrangement. Again the fact that $A(\infty)=r$ indicates that equality between the two sides of (14) can never be achieved. Also the envelope of the family of straight lines defined by the right-hand side of (14) and resulting from the variation of y between zero and one is shown in Fig. 5. As y approaches zero, both the ordinate and abscissa intercepts approach one and as y approaches one, the ordinate intercept approaches 1.773 whereas the abscissa intercept approaches zero.

Appendix I

Proofs of Lemmas

Proof of Lemma 1

First assume that the point x is such that $\cos x\leqq\cos \pi y$. For n equal to zero or a positive integer.

$$\begin{aligned}\sin x &= \int_0^x \cos u\,du \\ &= \int_0^{\pi y}\cos u\,du + \int_{\pi y}^{\pi(2-y)}\cos u\,du + \int_{\pi(2-y)}^{\pi(2-y)}\cos u\,du \\ &\quad + \int_{\pi(2+y)}^{\pi(4-y)}\cos u\,du + \cdots + \int_{\pi(2n+y)}^{x}\cos u\,du \\ &= \int_0^{(2n+1)\pi y} g(u)\,du + \int_0^{x-(2n+1)\pi y} h(u)\,du\end{aligned}$$

where $g(u)$ is periodic with a period of $2\pi y$ and equals $\cos u$ for $-\pi y\leqq u\leqq\pi y$ and $h(u)$ is periodic with a period of $2\pi(1-y)$ and equals $\cos(u+\pi y)$ for $0\leqq u\leqq 2\pi(1-y)$. But $g(u)\leqq v(u)$ and $h(u)\leqq\cos\pi y\leqq v(u)$ for all u by hypothesis. Thus

$$\int_0^x \cos u\,du \leqq \int_0^x v(u)\,du \quad \text{for all } x\geqq 0.$$

The proof proceeds in the same way for the case where $\cos x\geqq\cos\pi y$.

Proof of Lemma 2

Again the proof will be given only for the case where $\cos x\geqq\cos\pi(1-y)$ since the proof proceeds in the same way when $\cos x\leqq\cos\pi(1-y)$. For n a positive integer

$$\begin{aligned}\sin x &= \int_0^x \cos u\,du = \int_0^{\pi(1-y)}\cos u\,du \\ &\quad + \int_{\pi(1-y)}^{\pi(1+y)}\cos u\,du + \int_{\pi(1+y)}^{\pi(3-y)}\cos u\,du \\ &\quad + \int_{\pi(3-y)}^{\pi(3+y)}\cos u\,du + \cdots + \int_{\pi(2n-1+y)}^{x}\cos u\,du \\ &= \int_0^{2n\pi y} p(u)\,du + \int_0^{x-2n\pi y} q(u)\,du\end{aligned}$$

where $p(u)$ is periodic with a period of $2\pi y$ and equals $\cos[u+\pi(1-y)]$ for $0\leq u\leq 2\pi y$ and $q(u)$ is periodic with a period of $2\pi(1-y)$ and equals $\cos u$ for $-\pi(1-y)\leq u\leq\pi(1-y)$. But since $w(u)\leq p(u)$ and $w(u)\leq\cos\pi(1-y)\leq q(u)$, this lemma holds when $x\geq\pi(1-y)$. For $0\leq x\leq\pi(1+y)$, the proof is trivial.

Proof of Lemma 3

Consider the following expansion[5] for the cosine function where $-\pi/2\leq\phi\leq\pi/2$ and m is a real number.

$$\cos m\phi = 1 - \frac{m^2}{2!}\sin\phi - \frac{m^2(2^2-m^2)}{4!}\sin^4\phi - \frac{m^2(2^2-m^2)(4^2-m^2)}{6!}\sin^6\phi - \cdots$$

If m is restricted to the interval, $0\leqq m<2$, then every term after the first is a nonpositive quantity and so $\cos m\phi$ is less than or equal to any finite sum obtained by terminating this series at any term after the first. Terminating after the $(N+1)$th term, replacing the powers of the sine function by their expansions[6] in terms of the sums of cosines of multiples of ϕ and using the change of variable, $y=m/2$, the following expression may be obtained where $0\leqq y<1$ and the Q's are given by (7) and (8).

$$\cos 2y\phi\leq Q_0+Q_2\cos 2\phi+Q_4\cos 4\phi+\cdots+Q_{2N}\cos 2N\phi.$$

Letting $u=2y\phi$ and then invoking lemma 1, the inequality (5) will follow.

Similarly the negative of the function obtained by replacing ϕ in the right-hand side of the last inequality by $[(u/2y)-(\pi/2)]$ satisfies the conditions of lemma 2 and therefore (6) is true.

[5] E. W. Hobson, "A Treatise on Plane Trigonometry," Cambridge University Press, Cambridge, Eng., 7th Ed., p. 276, eq. (5); 1939.

[6] *Ibid.*, p. 54, eq. (44).

Appendix II

List of Symbols

Those symbols which have physical significance are listed below.

$A(t)$ = The response to a unit step function applied at time, $t=0$.

C = The capacity that a system simulates as frequency approaches infinity.

γ = The least upper bound on all the fractional overshoots and undershoots of the unit step response.

δ = The least upper bound on $|A(t)-r|/r$ for $t \geq \tau_\delta$.

$R(\omega)$ = The real part of a system function for real frequencies.

r = The resistance of a system function under dc conditions.

s = The complex frequency variable.

T_γ = The rise time from zero to one of the step response.

t = The time variable.

τ = Any time beyond which the unit step response remains within the bounds $(1 \pm \gamma)r$.

τ_δ = The least time beyond which the unit step response remains within the bounds $(1 \pm \delta)r$ where $0 \leqq \delta \leqq 1$.

$W(t)$ = The response to a unit impulse function applied at time, $t=0$.

$\omega = (2\pi)(\text{frequency})$ = angular frequency.

$Z(s)$ = A system function.

Acknowledgment

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Correction

Sol Sherr, author of the article "Generalized Equations for RC Phase-Shift Oscillators," which appeared on pages 1169–1172 of the July, 1954 issue of Proceedings, has requested that the following equation, from which the last three terms were omitted in error in printing, should read:

$$A = -8 - \frac{R_1}{R}\left(\frac{11}{K} + \frac{4}{K^2} + 8\right) - \left(\frac{R_1}{R}\right)^2\left(\frac{2}{K} + 2\right) - \frac{12}{K} - \frac{7}{K^2} - \frac{2}{K^3} \quad (3b).$$

IRE Standards on Electronic Computers: Definitions of Terms, 1956*

(56 IRE 8.S1)

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Introductory Note

Boldface is used for alphabetic entries, and, in the body of an essay, at the point where the **boldface** term is being defined.

Italics are used to show that the *italicized* term is defined elsewhere in the list of definitions.

Access Time. A time interval which is characteristic of a storage unit, and is essentially a measure of the time required to communicate with that unit. Many definitions of the beginning and ending of this interval are in common use.

Accumulator. A device which stores a number and which, on receipt of another number, adds it to the number already stored and stores the sum.

Note: The term is also applied to devices which function as described but which also have other properties.

Accuracy. The quality of freedom from mistake or error, that is, of conformity to truth or to a rule. Accuracy is distinguished from *Precision* as in the following example: A six-place table is more precise than a four-place table. However, if there are errors in the six-place table, it may be either more or less accurate than the four-place table.

Adder. A device which can form the sum of two or more numbers or quantities.

Address. An expression, usually numerical, which designates a particular location in a *Storage* or *Memory* device or other source or destination of information. See also *Instruction Code*.

Address Part. In an instruction, any part that is usually an *Address*. See also *Instruction Code*.

Analog (in Electronic Computers). A physical system in which the performance of measurements yields information concerning a class of mathematical problems.

Analog Computer. A physical system together with means of control for the performance of measurements (upon the system) which yield information concerning a class of mathematical problems.

And-Circuit. Synonym for *And-Gate*.

And-Gate. A gate whose output is energized when and only when every input is in its prescribed state. An *And-Gate* performs the function of the logical "and."

Arithmetic Element. Synonym for *Arithmetic Unit*.

Arithmetic Unit. That part of a computer which performs arithmetic operations.

Automatic Check. See *Check, Automatic*.

Band (in Electronic Computers). A group of *Tracks* on a magnetic drum.

Base. See *Positional Notation*.

Binary. See *Positional Notation*.

Binary Cell. An elementary unit of storage which can be placed in either of two stable states.

Binary-Coded-Decimal System. A system of number representation in which each decimal digit is represented by a group of binary digits (*e.g.*, *Excess-Three Code*).

Binary Number System. See *Positional Notation*.

Binary Point. See *Point*.

Bit (in Electronic Computers). 1) An abbreviation of "binary digit." 2) A single *Character* of a *Language* employing exactly two distinct kinds of characters. 3) A unit of storage capacity. The capacity, in bits, of a storage device is the logarithm to the base two of the number of possible states of the device. See also *Storage Capacity*.

Block. A group of *Words* considered as a unit.

Borrow. See *Carry*.

Branch. Synonym for *Conditional Jump*.

Break Point. A place in a *Routine* at which a special instruction is inserted which, if desired, will cause a digital computer to stop for a visual check of progress.

Buffer. 1) An isolating circuit used to avoid reaction of a driven circuit on the corresponding driving circuit. 2) A storage device used to compensate for a difference in rate of flow of information or time or occurrence of events when transmitting information from one device to another.

Bus (in Electronic Computers). One or more conductors which are used as a path for transmitting information from any of several sources to any of several destinations.

Carry. 1) A signal, or expression, produced as a result of an arithmetic operation on one digit place of two or more numbers expressed in *Positional Notation* and transferred to the next higher place for processing there. 2) Usually a signal or expression as defined in 1) above which arises in adding, when the sum of two digits in the same digit place equals or exceeds the *Base* of the number system in use. If a carry into a digit place will result in a carry out of the same digit place, and if the normal adding circuit is bypassed when generating this new carry, it is called a **High-Speed Carry,** or **Standing-on-Nines Carry.** If the normal adding circuit is used in such a case, the carry is called a **Cascaded Carry.** If a carry resulting from the addition of carries is not allowed to propagate (*e.g.*, when forming the partial product in one step of a multiplication process), the process is called a **Partial Carry.** If it is allowed to propagate, the process is called a **Complete Carry.** If a carry generated in the most significant digit place is sent directly to the least significant place (*e.g.*, when adding two negative numbers using nines complements) that carry is called an **End-Around Carry.** 3) In direct subtraction, a signal or expression as defined in 1) above which arises when the difference between the digits is less than zero. Such a carry is frequently called a **Borrow.** 4) The action of forwarding a carry. 5) The command directing a carry to be forwarded.

Cascaded Carry. See *Carry.*

Cell. An elementary unit of storage (*e.g.*, binary cell, decimal cell).

Channel (in Electronic Computers). That portion of a storage medium which is accessible to a given reading station. See also *Track.*

Character (in Electronic Computers). One of a set of elementary marks or events which may be combined to express information.

Note: A group of characters, in one context, may be considered as a single character in another, as in the *Binary-Coded-Decimal System.*

Check. A process of partial or complete testing of 1) the correctness of machine operations, 2) the existence of certain prescribed conditions within the computer, or 3) the correctness of the results produced by a *Routine.* A check of any of these conditions may be made automatically by the equipment or may be programmed. See also *Marginal Checking; Verification.*

Check, Automatic. A *Check* performed by equipment built into the computer specifically for that purpose, and automatically accomplished each time the pertinent operation is performed. Sometimes referred to as a built-in check. **Machine Check** can refer to an automatic check, or to a *Programmed Check* of machine functions.

Check Digits. See *Check, Forbidden-Combination.*

Check, Forbidden-Combination. A *Check* (usually an *Automatic Check*) which tests for the occurrence of a nonpermissible code expression. A **Self-Checking Code** (or **Error-Detecting Code**) uses code expressions such that one (or more) error(s) in a code expression produces a forbidden combination. A **Parity Check** makes use of a self-checking code employing binary digits in which the total number of 1's (or 0's) in each permissible code expression is always even or always odd. A check may be made for either even parity or odd parity. A **Redundancy Check** employs a self-checking code which makes use of redundant digits called **Check Digits.**

Check Problem. See *Check, Programmed.*

Check, Programmed. A *Check* consisting of tests inserted into the program of the problem and accomplished by appropriate use of the machine's instructions. A **Mathematical Check** (or *Control*) is a programmed check of a sequence of operations which makes use of the mathematical properties of that sequence.

A **Check Routine** or **Check Problem** is a routine or problem which is designed primarily to indicate whether a fault exists in the computer, without giving detailed information on the location of the fault. See also *Diagnostic Routine; Test Routine.*

Check Routine. See *Check, Programmed.*

Check, Selection. A *Check* (usually an *Automatic Check*) to verify that the correct register, or other device, is selected in the performance of an instruction.

Check, Transfer. A *Check* (usually an *Automatic Check*) on the accuracy of the transfer of a word.

Circulating Register (or **Memory**). A register (or memory) consisting of a means for delaying information and a means for regenerating and reinserting the information into the delaying means.

Clear. To restore a storage or memory device to a prescribed state, usually that denoting zero. See also *Reset.*

Clock. A primary source of synchronizing signals.

Code (in Electronic Computers). 1) A system of *Characters* and rules for representing information. 2) Loosely, the set of characters resulting from the use of a code. 3) To prepare a *Routine* in *Machine Language* for a specific computer. 4) To encode; to express given information by means of a code. See also *Language.*

Column. Synonym for *Place.*

Command. 1) One of a set of several signals (or groups of signals) which occurs as a result of an *Instruction;* the commands initiate the individual steps which form the process of executing the instruction. 2) Synonym for *Instruction.*

Complement. 1) A number whose representation is derived from the finite *Positional Notation* of another by one of the following rules: a) True complement—Subtract each digit from one less than the base; then add 1 to the least significant digit, executing all carries required. b) Base minus one's complement—Subtract each digit from one less than the base (*e.g.*, "9's complement" in the base 10, "1's complement" in the base 2, etc.). 2) To form the complement of a number.

Note: In many machines, a negative number is represented as a complement of the corresponding positive number.

Complete Carry. See *Carry.*

Computer. 1) A machine for carrying out calculations. 2) By extension, a machine for carrying out specified transformations on information.

Conditional Jump. An instruction which will cause the proper one of two (or more) addresses to be used in obtaining the next instruction, depending upon some property of one or more numerical expressions or other conditions.

Conditional Transfer Of Control. Synonym for *Conditional Jump.*

Control. 1) Usually, those parts of a digital computer which effect the carrying out of instructions in proper sequence, the interpretation of each instruction, and the application of the proper signals to the arithmetic unit and other parts in accordance with this interpretation. 2) Frequently, one or more of the components in any mechanism responsible for interpreting and carrying out manually-initiated directions. Sometimes called manual control. 3) In some business applications of mathematics, a *Mathematical Check.*

Copy. See *Transfer.*

Correction. See *Error.*

Counter. 1) A device capable of changing from one to the next of a sequence of distinguishable states upon each receipt of an input signal. 2) Less frequently, an *Accumulator.*

Counter, Ring. A loop of interconnected bistable elements such that one and only one is in a specified state at any given time and such that, as input signals are counted, the position of the one specified state moves in an ordered sequence around the loop.

Cyclic Shift. An operation which produces a *Word* whose *Characters* are obtained by a cyclic permutation of the characters of a given word.

Decimal Number System. See *Positional Notation.*

Decimal Point. See *Point.*

Decoder. A network or system in which a combination of inputs is excited at one time to produce a single output. Sometimes called *Matrix.*

Delay Line (in Electronic Computers). 1) Originally, a device utilizing wave propagation for producing a time delay of a signal. 2) Commonly, any device for producing a time delay of a signal.

Delay-Line Memory. Synonym for *Delay-Line Storage.*

Delay-Line Storage. A storage or memory device consisting of a delay line and means for regenerating and reinserting information into the delay line.

Diagnostic Routine. A *Routine* designed to locate either a malfunction in the computer or a mistake in coding. See also *Check, Programmed.*

Differentiator (in Electronic Computers). A device, usually of the analog type, whose output is proportional to the derivative of an input signal.

Digit. See *Positional Notation.*

Digital Computer. A computer which operates with information, numerical or otherwise, represented in a digital form.

Double-Length Number. A number having twice as many digits as are ordinarily used in a given computer.

Double-Precision Number. Synonym for *Double-Length Number.*

Encoder. A network or system in which only one input is excited at a time and each input produces a combination of outputs. Sometimes called *Matrix.*

End-Around Carry. See *Carry.*

Error. 1) In mathematics, the difference between the true value and a calculated or observed value. A quantity (equal in absolute magnitude to the error) added to a calculated or observed value to obtain the true value is called a **Correction.** 2) In a computer or data-processing system, any incorrect step, process, or result. In addition to the mathematical usage, in the computer field the term is also commonly used to refer to machine malfunctions as "machine errors" and to human mistakes as "human errors." It is frequently helpful to distinguish between these as follows: errors result from approximations used in numerical methods; **Mistakes** result from incorrect programming, coding, data transcription, manual operation, etc.; **Malfunctions** result from failures in the operation of machine components such as gates, flip-flops, amplifiers, etc.

Error-Detecting Code. See *Check, Forbidden-Combination.*

Excess-Three Code. A number *Code* in which the decimal digit n is represented by the four-bit binary equivalent of $n+3$.
See also *Binary-Coded-Decimal System.*

Extract. To form a new *Word* by juxtaposing selected segments of given words.

Fixed-Point System. See *Point.*

Flip-Flop. 1) A device having two stable states and two input terminals (or types of input signals) each of which corresponds with one of the two states. The circuit remains in either state until caused to change to the other state by application of the corresponding signal. 2) A similar bistable device with an input which allows it to act as a single-stage binary *Counter.*

Floating-Point System. See *Point.*

Flow Diagram (in Electronic Computers). A graphical representation of a *Program* or a *Routine.*

Forbidden-Combination Check. See *Check, Forbidden-Combination.*

Four-Address Code. See *Instruction Code.*

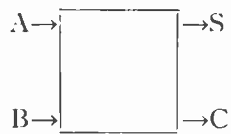
Gate (in Electronic Computers). A circuit having an output and a multiplicity of inputs so designed that the output is energized when and only when certain input conditions are met. See also *And-Gate; Or-Gate.*

Note: Sometimes "gate" is used for "and-gate."

Half Adder. A circuit having two input and two output channels for binary signals (0, 1) and in which the output signals are related to the input signals according to Table I.

TABLE I

Input To		Output From	
A	B	S	C
0	0	0	0
0	1	1	0
1	0	1	0
1	1	0	1



(So called because two half adders can be used in the construction of one binary *Adder*.)

Hexadecimal. See *Positional Notation.*

High-Speed Carry. See *Carry.*

Inhibiting Input. A *Gate* input which, if in its prescribed state, prevents any output which might otherwise occur.

Instruction. See *Instruction Code.*

Instruction Code. An artificial *Language* for describing or expressing the instructions which can be carried out by a digital computer. In automatically sequenced computers, the instruction code is used when describing or expressing sequences of **Instructions,** and each instruction word usually contains a part specifying the operation to be performed and one or more *Addresses* which identify a particular location in storage. Sometimes an *Address Part* of an instruction is not intended to specify a location in storage but is used for some other purpose.

If more than one address is used, the code is called a **Multiple-Address Code.** In a typical instruction of a **Four-Address Code** the addresses specify the location of two operands, the destination of the result, and the location of the next instruction in the sequence. In a typical **Three-Address Code,** the fourth address specifying the location of the next instruction is dispensed with and the instructions are taken from storage in a preassigned order.

In a typical **One-Address** or **Single-Address Code,** the address may specify either the location of an operand to be taken from storage, the destination of a previously prepared result, or the location of the next instruction. The arithmetic element usually contains at least two storage locations, one of which is an accumulator. For example, operations requiring two operands may obtain one operand from the main storage and the other from a storage location in the arithmetic element which is specified by the operation part.

Integrator (in Electronic Computers). 1) A device whose output is proportional to the integral of an input signal. 2) In certain digital machines, a device for numerically accomplishing an approximation to the mathematical process of integration.

Jump. To (conditionally or unconditionally) cause the next instruction to be selected from a specified storage location.

Language (in Electronic Computers). 1) A system consisting of a) a well defined, usually finite, set of characters; b) rules for combining characters with one another to form words or other expressions; and c) a specific assignment of meaning to some of the words or expressions, usually for communicating information or data among a group of people, machines, etc. 2) A system similar to the above but without any specific assignment of meanings. Such systems may be distinguished from 1) above, when necessary, by referring to them as formal or uninterpreted languages. Although it is sometimes convenient to study a language independently of any meanings, in all practical cases at least one set of meanings is eventually assigned. See also *Code; Machine Language.*

Logic. See *Logical Design.*

Logical Design. 1) The planning of a computer or data-processing system prior to its detailed engineering design. 2) The synthesizing of a network of *Logical Elements* to perform a specified function. 3) The result of 1) and 2) above, frequently called the **Logic** of the system, machine, or network.

Logical Diagram. In *Logical Design,* a diagram representing the *Logical Elements* and their interconnections without necessarily expressing construction or engineering details.

Logical Element. In a computer or data-processing system, the smallest building blocks which can be represented by operators in an appropriate system of symbolic logic. Typical logical elements are the and-gate and the flip-flop, which can be represented as operators in a suitable symbolic logic.

Logical Operation. 1) Any nonarithmetical operation. Examples are: *Extract,* logical (bit-wise) multiplication, *Jump,* data transfer, etc. 2) Sometimes, only those nonarithmetical operations which are expressible bitwise in terms of the propositional calculus or a two-valued Boolean algebra.

Logical Symbol. A symbol used to represent a *Logical Element* graphically.

Machine Check. See *Check, Automatic.*

Machine Language. 1) A *Language,* occurring within a machine, ordinarily not perceptible or intelligible to persons without special equipment or training. 2) A translation or transliteration of 1) above into more conventional characters but frequently still not intelligible to persons without special training.

Major Cycle. In a storage device which provides *Serial* access to storage positions, the time interval between successive appearances of a given storage position.

Malfunction. See *Error.*

Marginal Checking. A preventive maintenance procedure in which certain operating conditions, *e.g.*, supply voltage or frequency, are varied about their normal values in order to detect and locate incipient defective units. See also *Check.*

Marginal Testing. Synonym for *Marginal Checking.*

Master Routine. See *Subroutine.*

Mathematical Check. See *Check, Programmed.*

Matrix (in Electronic Computers). 1) Any logical network whose configuration is a rectangular array of intersections of its input-output leads, with elements connected at some of these intersections. The network usually functions as an *Encoder* or *Decoder.* 2) Loosely, any encoder, decoder, or *Translator.*

Memory. See *Storage.*

Memory Capacity. Synonym for *Storage Capacity.*

Minor Cycle. In a storage device which provides *Serial* access to storage positions, the time interval between the appearance of corresponding parts of successive words.

Mistake. See *Error.*

Multiple-Address Code. See *Instruction Code.*

Multiplier. A device which has two or more inputs and whose output is a representation of the product of the quantities represented by the input signals.

Number. 1) Formally, an abstract mathematical entity which is a generalization of a concept used to indicate quantity, direction, etc. In this sense a number is independent of the manner of its representation. 2) Commonly: A representation of a number as defined above (*e.g.*, the binary number "10110," the decimal number "3695," or a sequence of pulses). 3) An expression composed wholly or partly of digits which does not necessarily represent the abstract entity mentioned in the first meaning.

Note: Whenever there is a possibility of confusion between meaning 1) and meaning 2) or 3), it is usually possible to make an unambiguous statement by using "number" for meaning 1) and "numerical expression" for meaning 2) or 3).

Number System. See *Positional Notation.*

Octal. See *Positional Notation.*

Octonary. See *Positional Notation.*

One-Address Code. See *Instruction Code.*

Operation Code. 1) The list of *Operation Parts* occurring in an *Instruction Code*, together with the names of the corresponding operations (*e.g.*, "add," "unconditional transfer," "add and clear," etc.). 2) Synonym for *Operation Part* of an instruction.

Operation Part. In an instruction, the part that usually specifies the kind of operation to be performed, but not the location of the operands. See also *Instruction Code.*

Or-Circuit. Synonym for *Or-Gate.*

Order. 1) Synonym for *Instruction.* 2) Synonym for *Command.* 3) Loosely, synonym for *Operation Part.*

Note: The use of "order" in the computer field as a synonym for terms similar to the above is losing favor owing to the ambiguity between these meanings and the more common meanings in mathematics and business.

Or-Gate. A gate whose output is energized when any one or more of the inputs is in its prescribed state. An or-gate performs the function of the logical "inclusive-or."

Overflow. 1) The condition which arises when the result of an arithmetic operation exceeds the capacity of the number representation in a digital computer. 2) The *Carry* digit arising from this condition.

Parallel (in Electronic Computers). Pertaining to simultaneous transmission of, storage of, or logical operations on the parts of a word, character, or other subdivision of a word, using separate facilities for the various parts.

Parallel Digital Computer. One in which the digits are handled in parallel. Mixed serial and parallel machines are frequently called serial or parallel according to the way arithmetic processes are performed. An example of a parallel digital computer is one which handles decimal digits in parallel although it might handle the bits which comprise a digit either serially or in parallel. See also *Serial Digital Computer.*

Parity Check. See *Check, Forbidden-Combination.*

Partial Carry. See *Carry.*

Place. In *Positional Notation*, a position corresponding to a given power of the base. A digit located in any particular place is a coefficient of a corresponding power of the base.

Point. In *Positional Notation*, the *Character*, or the location of an implied symbol, which separates the integral part of a numerical expression from its fractional part. For example, it is called the **Binary Point** in binary notation and the **Decimal Point** in decimal notation. If the location of the point is assumed to remain fixed with respect to one end of the numerical expressions, a **Fixed-Point System** is being used. If the location of the point does not remain fixed with respect to one end of the numerical expressions, but is regularly recalculated, then a **Floating-Point System** is being used.

Note: A fixed-point system usually locates the point by some convention, while a floating-point system usually locates the point by expressing a power of the base.

Positional Notation. One of the schemes for representing numbers, characterized by the arrangement of digits in sequence, with the understanding that successive digits are to be interpreted as coefficients of successive powers of an integer called the **Base** of the **Number System.**

In the **Binary Number System** the successive digits are interpreted as coefficients of the successive powers of the base two just as in the **Decimal Number System** they relate to successive powers of the base ten.

In the ordinary number systems each **Digit** is a *Character* which stands for zero or for a positive integer smaller than the base.

The names of the number systems with bases from 2 to 20 are: **Binary, Ternary,** quaternary, quinary, senary, septenary, **Octonary,** (also **Octal**), novenary, decimal, undecimal, duodecimal, terdenary, quaterdenary, quindenary, **Sexadecimal** (also **Hexadecimal**), septendecimal, octodenary, novendenary, and vicenary. The sexagenary number system has the base 60. The commonly used alternative of saying "base-3," "base-4," etc., in place of tenary, quaternary, etc., has the advantage of uniformity and clarity.

Note: In the most common form of positional notation the expression

$$\pm\ a_n a_{n-1} \cdots a_2 a_1 a_0 \cdot a_{-1} a_{-2} \cdots a_{-m}$$

is an abbreviation for the sum

$$\pm \sum_{i=-m}^{n} a_i r^i$$

where the *Point* separates the positive powers from the negative powers, the a_i are integers $(0 \leqq a_i < r)$ called "digits," and r is an integer, greater than one, called the "base."

Note: For some purposes special rules are followed. In one such usage the value of the base, r, is not constant. In this case, the digits are coefficients of successive products of a nonconstant sequence of integers.

Precision. The quality of being exactly or sharply defined or stated. A measure of the precision of a representation is the number of distinguishable alternatives from which it was selected, which is sometimes indicated by the number of significant digits it contains. See also *Accuracy*.

Program. 1) A plan for the solution of a problem. 2) Loosely, a synonym for *Routine*. 3) To prepare a program.

Programmed Check. See *Check, Programmed*.

Radix. Synonym for *Base*.

Read. To acquire information, usually from some form of storage. See also *Write*.

Redundancy Check. See *Check, Forbidden-Combination*.

Regeneration (in Electronic Computers). In a storage device whose information storing state may deteriorate, the process of restoring the device to its latest undeteriorated state. See also *Rewrite*.

Register. A device capable of retaining information, often that contained in a small subset (*e.g.*, one *Word*) of the aggregate information in a digital computer. See also *Storage*.

Register Length. The number of characters which a register can store.

Reset. 1) To restore a storage device to a prescribed state. 2) To place a binary cell in the initial or "zero" state. See also *Clear*.

Rewrite. In a storage device whose information storing state may be destroyed by reading, the process of restoring the device to its state prior to reading.

Ring Counter. See *Counter, Ring*.

Routine. A set of instructions arranged in proper sequence to cause a computer to perform a desired operation, such as the solution of a mathematical problem.

Selection Check. See *Check, Selection*.

Self-Checking Code. See *Check, Forbidden-Combination*.

Serial. Pertaining to time-sequential transmission of, storage of, or logical operations on the parts of a word, using the same facilities for successive parts.

Serial Digital Computer. One in which the digits are handled serially. Mixed serial and parallel machines are frequently called serial or parallel according to the way arithmetic processes are performed. An example of a serial digital computer is one which handles decimal digits serially although it might handle the bits which comprise a digit either serially or in parallel. See also *Parallel Digital Computer*.

Set. 1) To place a storage device in a prescribed state. 2) To place a binary cell in the "one" state.

Sexadecimal. See *Positional Notation*.

Shift. Displacement of an ordered set of characters one or more places to the left or right. If the characters are the digits of a numerical expression, a shift may be equivalent to multiplying by a power of the base.

Sign Digit. A character used to designate the algebraic sign of a number.

Single-Address Code. See *Instruction Code.*

Standing On-Nines Carry. See *Carry.*

Staticizer. A storage device for converting time sequential information into static parallel information.

Storage. 1) The act of storing information. (See also *Store.*) 2) Any device in which information can be stored, sometimes called a **Memory** device. 3) In a computer, a section used primarily for storing information. Such a section is sometimes called a **Memory** or a *Store* (British).

Note: The physical means of storing information may be electrostatic, ferroelectric, magnetic, acoustic, optical, chemical, electronic, electrical, mechanical, etc., in nature.

Storage Capacity. The amount of information that can be retained in a storage (or memory) device, often expressed as the number of *Words* that can be retained (given the number of digits, and the base, of the standard word).

When comparisons are made among devices using different bases and word lengths, it is customary to express the capacity in *Bits.* This number is obtained by taking the logarithm to the base 2 of the number of distinguishable states in which the storage can exist.

Note: The "storage (or memory) capacity of a computer" usually refers only to the principal internal storage section.

Store. 1) To retain information in a device from which it can later be withdrawn. 2) To introduce information into such a device. 3) British synonym for *Storage* 3).

Subroutine. 1) In a *Routine,* a portion that causes a computer to carry out a well-defined mathematical or logical operation. 2) A routine which is arranged so that control may be transferred to it from a **Master Routine** and so that, at the conclusion of the subroutine, control reverts to the master routine. Such a subroutine is usually called a closed subroutine. A single routine may simultaneously be both a subroutine with respect to another routine and a master routine with respect to a third. Usually control is transferred to a single subroutine from more than one place in the master routine and the reason for using the subroutine is to avoid having to repeat the same sequence of instructions in different places in the master routine.

Ternary. See *Positional Notation.*

Test Routine. 1) Usually a synonym for *Check Routine.* 2) Sometimes used as a general term to include both check routine and *Diagnostic Routine.*

Three-Address Code. See *Instruction Code.*

Track (in Electronic Computers). That portion of a moving-type storage medium which is accessible to a given reading station; *e.g.*, as on film, drum, tapes, or discs. See also *Band.*

Transcriber. Equipment associated with a computing machine for the purpose of transferring input (or output) data from a record of information in a given language to the medium and the language used by a digital computing machine (or from a computing machine to a record of information).

Transfer. 1) To transmit, or **Copy,** information from one device to another. 2) To *Jump.* 3) The act of transferring.

Transfer Check. See *Check, Transfer.*

Transfer Control. Synonym for *Jump.*

Translator. A network or system having a number of inputs and outputs and so connected that signals representing information expressed in a certain code, when applied to the inputs, cause output signals to appear which are a representation of the input information in a different code. Sometimes called *Matrix.*

Unconditional Jump. An instruction which interrupts the normal process of obtaining instructions in an ordered sequence, and specifies the address from which the next instruction must be taken.

Unconditional Transfer of Control. Synonym for *Unconditional Jump.*

Unit. A portion or subassembly of a computer which constitutes the means of accomplishing some inclusive operation or function, as: *Arithmetic Unit.*

Verification. The process of checking the results of one data transcription against the results of another data transcription. Both transcriptions usually involve manual operations. See also *Check.*

Volatile. A term descriptive of a storage medium in which information cannot be retained without continuous power dissipation.

Note: Storage devices or systems employing nonvolatile media may or may not retain information in the event of planned or accidental power removal.

Williams-Tube Storage. A type of electrostatic storage.

Word (in Electronic Computers). An ordered set of *Characters* which is the normal unit in which information may be stored, transmitted, or operated upon within a computer.

Word Time. Synonym for *Minor Cycle.*

Write. To introduce information, usually into some form of storage. See also *Read.*

P-N-P-N Transistor Switches*

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Summary—**The design, fabrication, and electrical characteristics of silicon *p-n-p-n* transistors with $\alpha > 1$ for use as switches is discussed. The increase of alpha with injection level can be used to construct two terminal *p-n-p-n* switches. The high impedance characteristic has an impedance determined chiefly by the capacitance of the junctions. This capacity is of the order of tens of micromicrofarads. The low impedance portion of the switching characteristics has a slope resistance of a few ohms and a total voltage drop of approximately one volt. Methods of fabrication include suitable combinations of solid diffusion and alloying. Possible applications of *p-n-p-n* switches include function generators, photorelay, and talking path switches.**

Introduction

Devices capable of exhibiting differential negative resistance have long held an important place in the electronic technology. The differential negative resistance as such is useful in many applications, but at least as important are the two stable dc steady states of operation which are implied by the dc negative resistance.

In principle a combination of amplifying elements such as vacuum tubes or transistors can be connected in suitable circuit arrangement to result in bistable operation. For many specialized applications, however, it is desirable to synthesize the negative resistance function in a single device rather than to synthesize it from several devices and a relatively more complicated circuit.

The physical mechanisms which have produced negative resistances have in common an internal multiplication process which is a function of applied current and voltage. In the gaseous discharge, this process is the ionization of atoms by impact from electrons. At small current densities the ionization process increases in efficiency with increasing current. This results in a decrease in the total voltage as the current increases—hence a differential negative resistance [see Fig. 1(a)].

The "avalanche transistor"[1] is another example of negative resistance. In this case the multiplication process itself is a function only of applied voltage, but the relative number of minority carriers available for the multiplication process is a function of current. The dependence of the number of minority carriers on current can be tailored by an external resistor connected between emitter and base [see Fig. 1(b)]. In this connection, the current to the emitter is shunted by the resistor at low currents but, because of the nonlinearity of the emitter impedance, is not effectively shunted at high currents—hence the effective ionization or multiplication is increased with increasing current. Any transistor with $\alpha > 1$ can be connected in a suitable circuit to result in a differential dc negative resistance.[2] The avalanche transistor, point contact transistors, filamentary transistors, and *p-n-p-n* "hook collector" type junction transistors[3] can all be designed to have $\alpha > 1$ and hence can be used as bistable elements or, in the general sense, as switches.

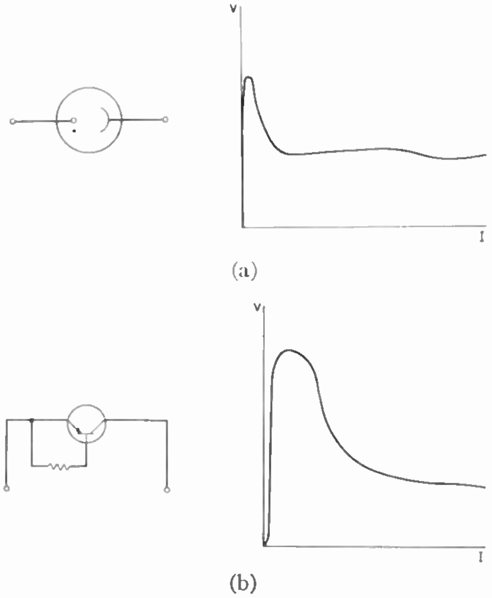

Fig. 1—(a) Gas tube negative resistance. Typical V-I characteristic of a hollow cathode gas-filled diode. Features of interest are the high impedance in the off state, the ratio of breakdown to sustain voltage (about 2:1), and the negative resistance in the conducting state. (b) Avalanche transistor characteristic. The schematic on the left indicates how the characteristic on the right may be obtained.

It is the purpose of this paper to discuss the design and performance of *p-n-p-n* transistors with $\alpha > 1$ as switching, or two-state, devices. Properly designed, the *p-n-p-n* transistor switch is a particularly simple circuit element for it is a two-terminal element or diode.

Mode of Operation

Fig. 2 shows the general type of V-I characteristic obtainable from the *p-n-p-n* transistor, if properly designed. This type of characteristic can be obtained under the following conditions. Let a voltage V be placed across the terminals of a *p-n-p-n* structure as in Fig. 3. The junctions J_1 and J_3 will be called the emitter

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[1] S. L. Miller and J. J. Ebers, "Alloyed junction avalanche transistors," *Bell Syst. Tech. J.*, vol. 34, pp. 883–902; September, 1955.

[2] A. E. Anderson, "Transistors in switching circuits," Proc. IRE, vol. 40, pp. 1541–1558; November, 1950.

[3] W. Shockley, "Theories of high values of alpha for collector contacts on germanium," *Phys. Rev.*, vol. 78, pp. 294–295; April 15, 1950. See also W. Shockley, M. Sparks, and G. K. Teal, "*P-N* junction transistors," *Phys. Rev.*, vol. 83, pp. 151–162; July 1, 1951.

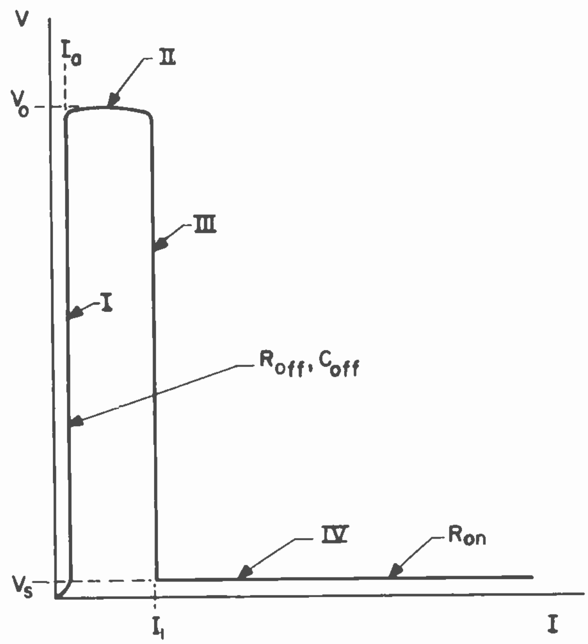

Fig. 2—The V-I characteristic of the p-n-p-n transistor. The four regions of primary interest are I, the off impedance; II, the region of small negative resistance; III, the region of large negative resistance; and IV, the on resistance.

junctions and J_2 will be called the collector junction.

If V is positive as shown in Fig. 3, the collector junction J_2 becomes reverse biased, and the emitters J_1 and J_3 become slightly forward biased. A forward current I flows through J_1 and J_3 and of course equal current crosses J_2. Now for J_2 reverse biased, the current at J_2 will be

$$I = IM_p\alpha_{1N} + IM_n\alpha_{2N} + I_{c0} \tag{1}$$

where α_{1N} is the fraction of the current at J_1, which is collected at J_2 as minority carrier current, and α_{2N} is the fraction of the current at J_3 which is collected at J_2.

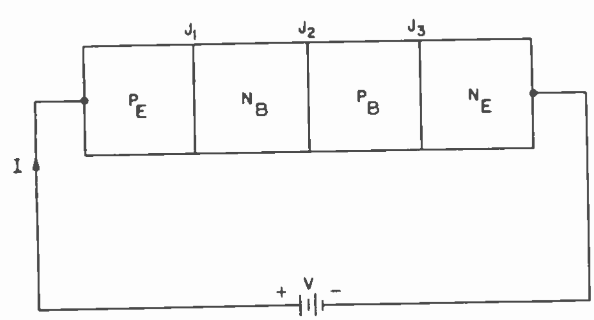

Fig. 3—Schematic of the p-n-p-n transistor. The end regions, as indicated by the subscripts, are emitters while the center regions act dually as base and collector. The center junction, J_2, is the collector, and both outside junctions, J_1 and J_3, are emitters when voltage is applied as shown.

The avalanche multiplication factors for holes and electrons are given by M_p and M_n respectively.[4] I_{c0} is the current that would flow through J_2 if J_2 were reverse biased, and isolated, *i.e.*, if α_{1N} and α_{2N} were zero.

[4] S. L. Miller, "Avalanche breakdown in germanium," *Phys. Rev.*, vol. 99, pp. 1234–1241; August 15, 1955.

From (1),

$$I = \frac{I_{c0}}{1 - M_p\alpha_{1N} - M_n\alpha_{2N}} = \frac{I_{c0}}{1 - \alpha_T}. \tag{2}$$

For purposes of discussion, let $M_p\alpha_{1N} + M_n\alpha_{2N}$ be called α_T, the total alpha. The high impedance portion or "off state" of the V-I curve in Fig. 2 (*i.e.*, region I) arises if the low current low voltage alpha satisfies the relation

$$\alpha_{1N} + \alpha_{2N} < 1. \tag{3}$$

At voltages appreciably less than V_B, the breakdown voltage of the center junction, the multiplication factors are essentially unity. In this case the current that flows is of the order of magnitude of I_{c0}.

If, however,

$$\alpha_{1N} + \alpha_{2N} \geqq 1, \tag{4}$$

then relation 2 is invalid and the current that flows is limited essentially by the external circuit. The total current at the collector is equal to the total current at each of the emitters. A total current I crosses each emitter and a current of $\alpha_{1N}I$ holes and $\alpha_{2N}I$ electrons reach the collector. If (4) is satisfied, the current at the collector is greater than I. To maintain the circuit condition that the total current at the collector is equal to I, the collector must become forward biased so as to emit electrons and holes back into the base layers. Hence (4) implies a low impedance or "on state" characteristic (region 4).

The transistor requirement for obtaining the V-I characteristic of Fig. 2 is that

$$\begin{aligned} \alpha_{1N} + \alpha_{2N} &< 1 \qquad I < I_1 \\ \alpha_{1N} + \alpha_{2N} &\geqq 1 \qquad I_1 < I \end{aligned} \tag{5}$$

where the α_{1N} and α_{2N} are the alphas at voltages low enough so that the multiplication factors are unity. Changes of α with current density are well known to transistor technology, even though they are not well understood, and it has been possible to design and construct p-n-p-n structures which satisfy the condition of (5). In Fig. 2, when V reaches V_0, the total alpha has reached unity by virtue of collector junction avalanche multiplication, and the current begins to increase (section II). This is a region of slight negative resistance where the total low voltage alpha is nearly unity due to increased current flow and the amount of avalanche multiplication required to keep the total high voltage alpha at unity decreases as current increases. As the current approaches I_1, the total low voltage alpha becomes essentially unity and a region of very high negative resistance (region III) is traversed. The high negative resistance persists until the collector junction J_2 begins to be forward biased and a new set of conditions apply. For currents greater than I_1, the total low voltage α is greater than unity, and a small positive resistance (region IV) is obtained.

Design Parameters

From the standpoint of utility, the device parameters of greatest interest include

I_0 = breakover current.
R-off = slope of section I, high impedance characteristic.
V_0 = breakover voltage.
C-off = capacitance of off device.
V_s = voltage drop in "on" region.
R-on = slope of section IV, low impedance characteristic.
I_1 = turn-on current.
Factors affecting speed.

I_0 and R-off

Most of these device parameters can be calculated from the device geometry or are insignificant in their effect on circuit operation. Those which can be made insignificant in their effect on circuit operation will be discussed first. This classification includes I_0, R-off. The slope resistance R-off can be made as high as 10^9 or 10^{10} ohms in a properly cleaned silicon junction at 300°K. This resistance is so high that in usual circumstances it can be neglected in its effect. The breakover current I_0 is easily much less than a microampere at room temperature and can be in the range of 10^{-8} amperes. Of course, as temperature is increased, I_0 increases and R-off will tend to decrease, but both parameters remain negligible up to about 100°C.

The parameters calculable from device geometry include V_0, breakover voltage; C-off, capacitance of the device in the off (region I) range; R-on, slope of on region; V_s, voltage drop in the on region.

Breakover Voltage

The breakover voltage is the voltage where the product of the junction avalanche multiplication and the low voltage alphas is unity. That is

$$M_p\alpha_{1N} + M_n\alpha_{2N} = 1 \tag{6}$$

where M_p is the multiplication for holes and M_n is the multiplication for electrons. In general, M_p and M_n are different, but to see the qualitative effect of multiplication on breakover voltage V_0, we will assume that they are equal and characterized, as in germanium[4] by

$$M = \frac{1}{1 - \left(\dfrac{V}{V_B}\right)^n} \tag{7}$$

where n is a parameter which is a function of breakdown voltage, V_B, and in the range of 2 or 3[5] for silicon in the range of $15 < V_B < 150$ volts. Then, at $V = V_0$, from (6) and (7)

$$M = \frac{1}{1 - \left(\dfrac{V_0}{V_B}\right)^n} = \frac{1}{\alpha_{1N} + \alpha_{2N}} \tag{8}$$

or

$$\frac{V_0}{V_B} = (1 - \alpha_{1N} - \alpha_{2N})^{1/n}. \tag{9}$$

If the low current alphas ($I < 10^{-6}$ amps) are of the order of 0.1 to 0.2 then for $n = 2$, or 3, V_0 is substantially V_B the breakdown voltage of the isolated collector junction J_2. If the multiplication for one type of carrier is much greater than the other type, then the alpha associated with this carrier is most significant in determining V_0/V_B.

Off-Capacitance

The off-capacitance is the series combination of the transition region capacitance of the three junctions J_1, J_2, and J_3. These transition region capacitances are functions of the impurity gradients at the junction, the applied voltage, and area. The total capacitance is a maximum at zero applied volts. The calculation of this capacitance has been adequately considered elsewhere and will not be considered in detail here.[6]

V-on, R-on

In the appendix the V-I characteristic of the two-terminal p-n-p-n is derived. In this derivation it is assumed that the multiplication factors are unity and the result is therefore valid only at voltages much less than V_0, the breakover voltage. The effects of multiplication can be included by multiplying the alphas by the appropriate factor. However, this formulation is intended to calculate the sustain voltage V_S, where the multiplication factors are unity. At low voltages the V-I characteristic of the two terminal p-n-p-n is given by

$$V = \frac{1}{\beta} \ln \frac{I_{s2}}{I_{s1}I_{s3}} \frac{(IA_1 + I_{s1})(IA_3 + I_{s3})}{(IA_2 + I_{s2})} \tag{10}$$

where $\beta = q/kT$.

I_{s1}, I_{s2}, I_{s3} = saturation current of J_1, J_2, J_3, respectively with the other two junctions shorted.

$$A_1 = \frac{1 + \alpha_{1I}\alpha_{2N} - \alpha_{2N}\alpha_{2I} - \alpha_{1I}}{1 - \alpha_{2N}\alpha_{2I} - \alpha_{1N}\alpha_{1I}}, \tag{11}$$

$$A_2 = \frac{\alpha_{1N} + \alpha_{2N} - 1}{1 - \alpha_{2N}\alpha_{2I} - \alpha_{1N}\alpha_{1I}}, \tag{12}$$

$$A_3 = \frac{1 + \alpha_{2I}\alpha_{1N} - \alpha_{1N}\alpha_{1I} - \alpha_{2I}}{1 - \alpha_{2N}\alpha_{2I} - \alpha_{1N}\alpha_{1I}}, \tag{13}$$

[5] S. L. Miller, private communication.

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

where

α_{1N} = the alpha with junction J_1 emitting, J_2 collecting,
α_{2N} = the alpha with junction J_3 emitting, J_2 collecting,
α_{1I} = the alpha when J_1 is collecting, J_2 emitting,
α_{2I} = the alpha when J_3 is collecting, J_2 emitting.

Note that

$$\alpha_{1I} + \alpha_{2I} \leqq 1 \tag{14}$$

since the total collected minority current cannot be more than the total emitted current.

Also, when

$$\alpha_{1N} + \alpha_{2N} < 1, \tag{15}$$

A_1 and A_3 are positive and A_2 is negative. This requires that

$$I < \frac{I_{s2}}{|A_2|} = \frac{I_{s2}(1 - \alpha_{2N}\alpha_{2I} - \alpha_{1N}\alpha_{1I})}{1 - \alpha_{2N} - \alpha_{2I}}. \tag{16}$$

This is in agreement with the earlier result that under the conditions prescribed I is of the order of a saturation current.

If

$$\alpha_{1N} + \alpha_{2N} > 1, \tag{17}$$

all of the A's are positive and I can increase without limit at a finite voltage. Furthermore, if the A's are substantially different from zero, say greater than 10^{-2}, and I is much larger than the saturation current we may write

$$V \cong \frac{1}{\beta} \ln \frac{I_{s2}I}{I_{s1}I_{s3}(1 - \alpha_{1N} - \alpha_{2N})} \tag{18}$$

or in other words, the voltage across the three junctions and two center regions is essentially that of a single forward biased diode. In practical *p-n-p-n* switches, A_1 and A_3 are always of the order of magnitude of unity, and A_2 changes its sign at $I = I_1$. Hence there is a small range of current where A_2 is very nearly zero and can make a large contribution to the voltage (18). However, it is unlikely that A_2 remains close to zero over a wide range of currents. The formulation leading to (18) neglects the resistance of the two end emitter regions. The total voltage drop across the device is

$$V = \frac{1}{\beta} \ln \frac{I}{I^*} + R_0 I \tag{19}$$

where R_0 is the ohmic series resistance of the end regions, and

$$I^* = \frac{I_{s1}I_{s3}}{I_{s2}}(1 - \alpha_{1N} - \alpha_{2N}). \tag{20}$$

The contribution of the logarithm in (19) should result in 0.6 to 0.8 volts so if R_0 is made of the order of 10^{-2} or less ohms (as can be done with diffused or alloyed contacts) the voltage drop at large fractions of amperes can be as low as one volt.

Turn-on Current

The parameters least amenable to calculation are I_1, the turn-on current, and the factors affecting speed of response. The turn-on current is determined by the current density at which the sum of the low voltage alphas reach unity. Thus, the current I_1 is a function of the device area, layer widths, initial lifetime, and the way carrier lifetime changes with injected current density. The recombination mechanism is not well enough understood to subject the turn-on current to exact analysis, but it can be empirically controlled.

Switching Speed

As to the possible switching speed, this is somewhat related to the turn-on current I_1. If the static current reaches I_1, the device turns on at a rate limited only by the external circuit parameters. However, if the device is turned on with a fast pulse, capacitive current flows which can affect the low voltage alpha and the device may turn on at a voltage less than V_0. Also, turn-off, which is attained by reducing the total current to less than I_1, requires a finite amount of time during which the excess carrier densities are decreasing. When the carrier density in the two floating base layers reaches a value corresponding to $\alpha < 1$ the device is turned off. The relations between turn-on and turn-off times and the circuit conditions make it difficult to specify these times in a simple way. However, using a series resistance and parallel capacitance in analogy to the gas tube circuit, a saw tooth oscillator was constructed which operated with a repetition rate of 2 mc. This indicates switching time of the order of a microsecond or less in the actual devices.

EXPERIMENTAL

Silicon *p-n-p-n* structures have been made which display the electrical characteristics shown in Fig. 2. The structures were produced by a combination of diffusion and alloying techniques. Three somewhat different designs will be described here which illustrate some of the major points of design flexibility and help verify the design theory of the preceding section.

The structures are shown schematically in Fig. 4, which shows also the manner in which the doping impurities are distributed in each structure. Structure A shown in Fig. 4(a) is produced by diffusing two impurities which produce opposite conductivity types in silicon into one surface of an n-type wafer. The fourth region is produced by alloying a p-type impurity into the diffused n layer. Structure B shown in Fig. 4(b) is produced by diffusing an n-type impurity into both

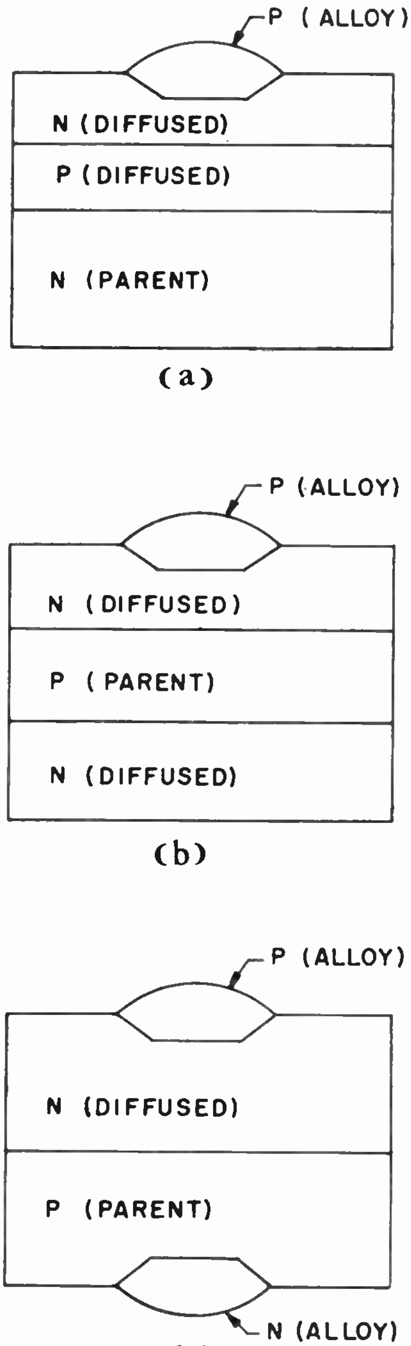

Fig. 4—Physical structure of *p-n-p-n* transistors. Structure *A* is obtained by a double diffusion of donors and acceptors into an *n*-type crystal to produce an *n-p-n*. The fourth region results from alloying an acceptor element into the diffused *n* layer. Structure *B* is obtained by diffusion of donors into both sides of a *p*-type wafer. The fourth region is obtained by alloying an acceptor element into one of the diffused *n*-layers. Structure *C* is similar to *B* except that one of the *n* layers is removed and both emitters are produced by alloying.

surfaces of a *p*-type silicon wafer. The fourth region is again produced by alloying a *p*-type impurity into one of the diffused *n* layers. Structure *C* [Fig. 4(c)] is produced by diffusing an *n*-type impurity into both surfaces, lapping off the back surface and alloying a *p*-type impurity into the *n* layer and an *n*-type impurity into the *p* layer.

The diffusion and alloying techniques have been described in an earlier paper.[7] The diffusant impurities employed in structure A were aluminum and antimony. The diffusions were performed in quartz tubes at 1250°C. from the vapor phase of the impurity elements. The *p*-type impurity used for alloying to form the outer *p* layer was metallic aluminum. The aluminum was vaporized in vacuum onto the surface of the diffused wafer. The thickness of the vaporized film and the subsequent conditions of alloying were controlled so as to alloy into but not through the diffused *n* layer. The diffusant in structure B was antimony. The *p*-type alloying agent in this case was also vaporized aluminum. The diffusant in structure C was also antimony: aluminum and gold-antimony alloy formed the vaporized contacts.

For the purposes of discussion the two outer layers of the *p-n-p-n* as shown in Fig. 4 will be referred to as P_E and N_E. The two inner layers will be called N_B and P_B so that reading from top to bottom the structure is P_E-N_B-P_B-N_E. It is evident from the figure that the P_E and N_B layers of the three structures are identical. In structure A, the P_B layer is a relatively heavily doped diffused layer and the N_E layer is the relatively lightly doped *n*-type original wafer. In structures B and C, the P_B layer is the original crystal. In structure B the N_E layer is a heavily doped diffused layer, and in structure C it is the alloyed gold-antimony contact. In all of the structures the P_E-N_B junction is a step junction on the more heavily doped P_B side and graded with an error-function-complement type of impurity distribution on the N_B side. In structure C the P_B-N_E junction is a step, while all other junctions in all of the structures are graded on both sides as can be seen from Fig. 4.

Electrical Characteristics

Fig. 5 shows the measured current-voltage characteristic of a typical unit with the structure shown in Fig. 4(a) at 25°C. and at 165°C. Fig. 6 shows similar characteristics of a typical unit with the structure of Fig. 4(b) or 4(c) at 22°C. and 102°C. All of the devices showed switching action in the two terminal connection. In the Introduction it was pointed out that this type of behavior requires an internal dependence of alpha on emitter current in at least one of the composite transistor structures which make up the *p-n-p-n*. By a slight modification of the fabrication procedure, it is possible to study the alpha variation in at least one of the two pertinent structures.

If in structure A, for example, enough aluminum is used, it is possible to alloy through the N_B layer and thus make contact to the P_B layer. The resulting structure is the diffused emitter and base silicon transistor.[7] Now if the N_B layer is used as the collector and the N_E layer as the emitter, it is possible to observe independ-

[7] M. Tanenbaum and D. E. Thomas, "Diffused emitter and base silicon transistors," *Bell Syst. Tech. J.*, vol. 35, pp. 1–23; January, 1956.

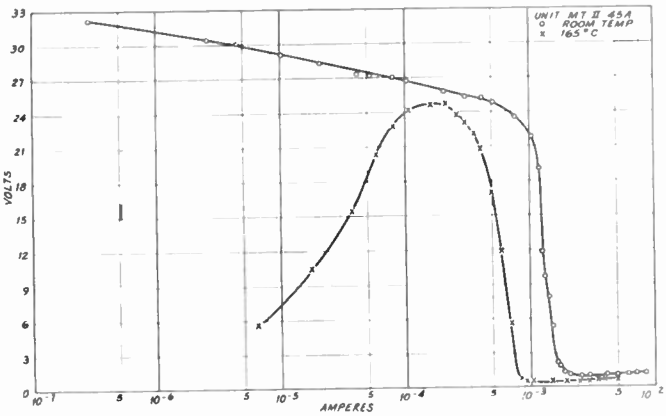

Fig. 5—V-I characteristic of a silicon p-n-p-n transistor of the structure shown in Fig. 4(a) at room temperature and 165°C.

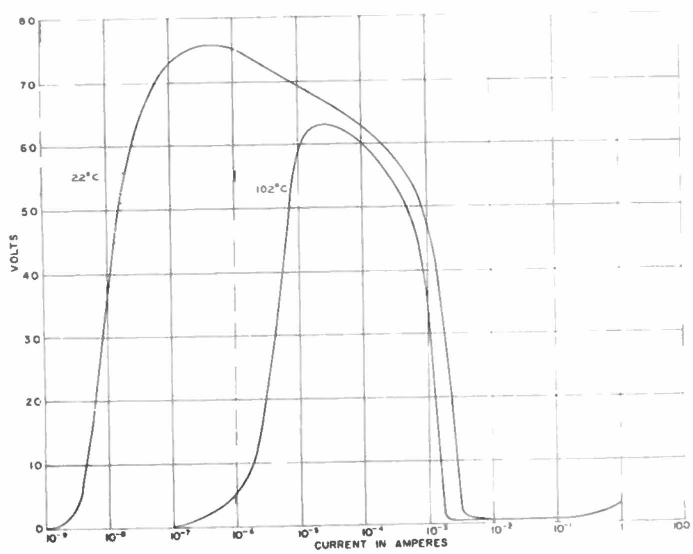

Fig. 6—V-I characteristic of a silicon p-n-p-n transistor of the structure shown in Fig. 4(b) at 22°C and 102°C.

ently one of the alphas which determines the behavior of the p-n-p-n. Fig. 7 is a plot of the alpha of this transistor as a function of emitter current. The measurements were made at a collector voltage of 2 volts since the breakdown voltage between the aluminum alloy contact and the n layer occurred at about 4 volts. It can be seen that alpha decreases rapidly at low emitter currents in just the manner required for the two-terminal operation of a p-n-p-n switch. The exact cause of this variation of alpha with emitter current has not been completely established. However, the variation has been observed in all silicon n-p-n diffused emitter and base transistors.[7] A similar variation of alpha has also been observed in silicon p-n-p transistors where the base layer was produced by diffusion and the emitter region was produced by alloying aluminum into the diffused base. This is also shown in Fig. 7. The observed variation could be caused by a recombination center in the base layer which is effective at low emitter currents but becomes saturated and relatively inactive at high injection levels.[8] This would lead to an effective increase in lifetime and thus an increase in alpha. It has been observed that a strong dc light will produce a considerable increase in alpha in these transistors at low emitter currents but has relatively little effect at large emitter currents. In addition, an increase in temperature will increase the alpha at low emitter currents but not produce an appreciable effect at large currents. Both of these observations are consistent with the proposal of a saturable recombination center.[9]

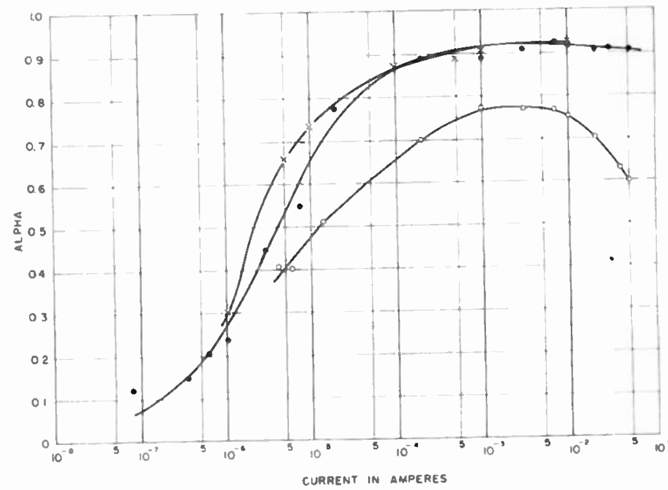

Fig. 7—Alpha as a function of emitter current in silicon diffused transistors. The solid circles are the α of a double diffused n-p-n as shown in Fig. 4(a) with the n (diffused) layer emitting into the p (diffused) layer. The open circles show the α of this transistor with the n (parent) layer emitting into the p (diffused) layer. The X's show the alpha of a diffused base alloyed emitter p-n-p as found in the structure in Fig. 4(b) and 4(c).

In all cases, the room temperature saturation current was masked by leakage current, and the slope resistance of the "off" characteristic was of the order of 10^8 or 10^9 ohms in properly cleaned and dried junctions. At elevated temperatures the "off" resistance is appreciably lowered. It was mentioned above that the alpha of the composite transistor structures at low emitter currents is observed to increase with increasing temperature. This would be reflected in an increase in current through the unit as shown by (2). This increase is noticeable at temperatures as low as 100°C. where the saturation current of the N_B-P_B junction, if it were isolated, should not have shown a measurable increase. The other parameter of importance in region I, the capacitance, is shown in Fig. 8 as a function of bias voltage. The unit is the one whose static characteristic is shown in Fig. 6. Note the agreement with theory in that the capacity falls off rapidly as bias of either polarity is applied. In addition, it is to be noted that the capacity at zero bias is 43 $\mu\mu f$ whereas the calculated value for the individual junctions themselves is 75 $\mu\mu f$ in this case.

[8] G. Bemski, "Lifetime of electrons in p-type silicon," *Phys. Rev.*, vol. 100, pp. 523–524; October 15, 1955.

[9] W. Shockley and W. T. Read, Jr., "Statistics of the recombination of holes and electrons," *Phys. Rev.*, vol. 87, pp. 835–842; September 1, 1952.

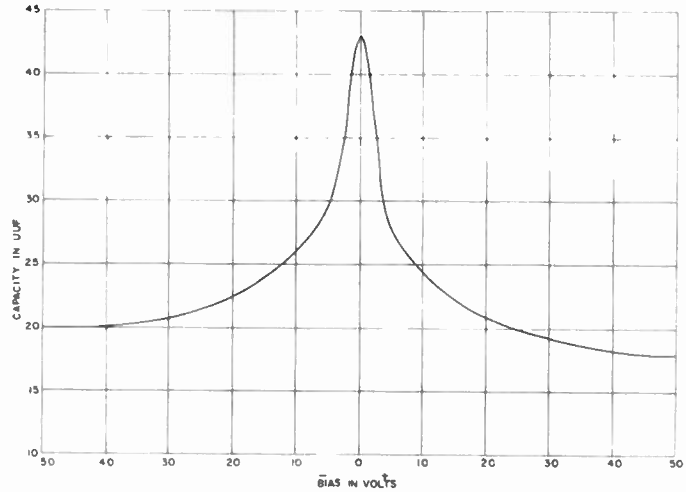

Fig. 8—Capacitance as a function of voltage of a silicon p-n-p-n transistor of the structure of 4(b).

The breakover voltage of the units is significantly different. In unit A, the reverse biased N_B-P_B junction is a junction between two rather heavily doped diffused layers. Thus, the breakover voltage of units with structure A will be independent of the original wafer resistivity as long as the wafer resistivity is high compared with that of the P_B layer. In units of structure B or C, however, the reverse biased junction is the junction between the diffused N_B layer and the P_B layer which is the original wafer. Thus the breakover voltage here is dependent on the resistivity of the original wafer, but can, of course, be controlled by the diffusion process.[10]

The measured breakover voltages for units with structure A are in the range of 25 to 35 volts when aluminum and antimony are used as diffusants and for units with structure B and C in the range of 25 to 100 volts, depending on the resistivity of the silicon from which the units were constructed. In each case the breakover voltage was approximately equal to the avalanche breakdown voltage of the P_B-N_B junction at room temperature and decreases somewhat as temperature is increased. The avalanche breakdown voltage increases with temperature, but so do the low current alphas. Eq. (9) shows that as the low current alphas increase, the ratio V_0/V_B decreases. The fact that the breakover voltage decreases with temperature indicates that the alpha change is the dominant temperature effect.

After the P_B-N_B junction breakdown, the p-n-p-n displays a range of small negative resistance (region II). However, at current I_1 (and sensibly voltage V_0) the negative resistance becomes very large (region III) and the unit switches to the "on" state. The range of small negative resistance (region II) extends from approximately 10^{-7} amperes to approximately 10^{-3} amperes. The upper limit is essentially the turn-on current I_1, and increases with the area of the device and the thickness of the base layers. The turn-on current also decreases as the lifetime of minority carriers in the base material increases.

It is difficult to determine if the observed values of negative resistance are consistent with the proposed model. The exact dependence of the current multiplication on voltage is not known in silicon. In addition the exact variation of alpha with current in the two composite transistor structures which form the p-n-p-n is not known. However, the fact that the switching action (region III) begins at a current near one milliampere coupled with the fact that in Fig. 7 the alpha of one of the composite transistors reaches a value near its maximum at approximately the same current would appear to substantiate the proposed model.

It can be seen from Figs. 5 and 6 that the turn-on current decreases with increasing temperature. This is in agreement with the earlier statement that the low current alpha of the composite transistor structures shows a marked increase in alpha at constant emitter current with an increase in temperature.

In the section on Mode of Operation it was shown that the sustain voltage depends on the values of the composite alphas in the "on" condition but that its likely variation would be between 0.6 and 0.8 volts. The observed values are within this range.

Another parameter of major importance is the "on" resistance. Here again there is a large difference between the structures. It was pointed out in the same section that this parameter is determined by the series resistance in the outer layers, *i.e.*, P_E and N_E in the terminology of this section. The P_E layer is identical in all three structures. However, in structure A the N_E layer is the lightly doped original wafer while in structure B it is the heavily doped diffused n layer. In structure C the N_E layer is the alloyed Au-Sb contact. Thus the "on" resistance of structures B and C would be expected to be appreciably less than that of A. This is observed and the "on" resistance of unit A is 50 ohms while that of units B or C is in the range of 1 to 3 ohms.

Comparison of Designs

The characteristics of the p-n-p-n switch that are calculable agree with the theoretical values. Structures B and C are very nearly identical in their properties with the exception that structure C has slightly lower "off" capacitance. The "on" resistance of structures B and C are much less than that of structure A—however, much thinner wafers of silicon must be used in the fabrication of B and C as compared to structure A. The "on" resistance is a very important factor in determining the current-handling capability of the switch and should be made as low as possible in high current applications.

[10] H. S. Veloric, M. B. Prince, and M. J. Eder, "Avalanche breakdown voltage in silicon diffused junctions," *J. Appl. Phys.*, vol. 27; August, 1956.

A very high speed switch requires thin base layers and low carrier lifetime. Structure A is most adaptable to fabrication with very thin base layers since the silicon wafer can be many times thicker than the base layers. Structures B and C are most adaptable to fabrication with very low "on" resistance. Clearly, the importance of any particular device parameter will dictate to some extent the method of fabrication. The three structures described do not, of course, exhaust the possible schemes of fabricating suitable *p-n-p-n* structures.

Applications

In many respects the V-I characteristic of the *p-n-p-n* is similar to the two-terminal gas tube, and a great many applications include precisely those things for which gas tubes are presently employed. In addition the *p-n-p-n* semiconductor device has characteristics peculiar to itself which suggest new applications. Some of the applications are listed below.

Talking Path Switch

The characteristics of a gas tube talking path switch have been described,[11] but the *p-n-p-n* offers an alternate possibility.

Thyratron

By making a third terminal connection to one of the base layers the *p-n-p-n* can be turned on by a small amount of control power much in the manner of a gas tube thyratron. If the series resistance is close to zero the structure will conduct several amperes at one volt drop.

Function Generation

In a circuit completely analogous to the gas tube circuit, the *p-n-p-n* can be used to generate saw-tooth waves, etc.

Photorelay

The *p-n-p-n* junctions are photosensitive as are other semiconductor junctions and the *p-n-p-n* can be triggered by light injected carriers as well as by electrically injected carriers. As a photorelay, the *p-n-p-n* is capable of operating directly electrical equipment requiring large fractions of amperes.

Appendix

Sustain Voltage of Two-Terminal *P-N-P-N* Transistors

The practical embodiment of the *p-n-p-n* transistor switch operates as a two terminal device, using an internal variation of α with current to give the switching characteristic. It is the purpose of this appendix to present a calculation of the sustain voltage of such a device in the "on" state. Fig. 9 shows schematically the *p-n-p-n* structure. For purposes of analysis, it will be supposed that electrical contact has been made to each region.

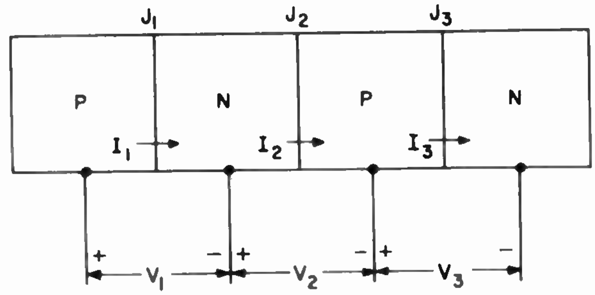

Fig. 9—Schematic of the *p-n-p-n* transistor. For purposes of analysis, it is supposed that connections have been made to all four regions.

Let the voltages V_1, V_2, V_3 be defined as the voltage across the junctions J_1, J_2, and J_3 respectively and taken as positive from left to right. Also, let I_1, I_2, and I_3 be the respective currents taken as positive to the right. The general problem is to relate the voltages V_1, V_2, and V_3 to the currents I_1, I_2, and I_3. If the problem is inverted, and the question is asked, what currents flow if V_1, V_2, V_3 are applied separately to the four regions, the principle of superposition is easily applied.[12] Thus, if $V_1>0$, $V_2=V_3=0$ then the left-hand *p-n-p* in Fig. 9 is simply a transistor with normal emitter bias and collector volts$=0$.

Let I_{s1} be the saturation current of junction 1 under these conditions.

Then

$$I_1 = I_{s1}(e^{\beta V_1} - 1) \tag{21}$$

$$I_2 = \alpha_{1N} I_{s1}(e^{\beta V_1} - 1) \tag{22}$$

$$I_3 = 0. \tag{23}$$

$\beta = q/kT$ and α_{1N} is the normal alpha of the left-hand *p-n-p*.

If $V_3>0$ and $V_1=V_2=0$, using similar notation as in (21), (22), (23)

$$I_1 = 0 \tag{24}$$

$$I_2 = \alpha_{2N} I_{s3}(e^{\beta V_3} - 1) \tag{25}$$

$$I_3 = I_{s3}(e^{\beta V_3} - 1). \tag{26}$$

α_{2N} is the alpha for junction 3 emitting and junction 2 collecting.

If $V_1=V_3=0$ $V_2<0$, junction 2 is emitting and junctions 1 and 3 are collecting. Then

$$I_1 = -\alpha_{1I} I_{s2}(e^{-\beta V_2} - 1) \tag{27}$$

$$I_2 = -I_{s2}(e^{-\beta V_2} - 1) \tag{28}$$

$$I_3 = -\alpha_{2I} I_{s2}(e^{-\beta V_2} - 1). \tag{29}$$

[11] M. A. Townsend and W. A. Depp, "Cold cathode tubes for transmission of audio signals," *Bell Syst. Tech. J.*, vol. 32, pp. 1371–1391; November, 1953.

[12] J. J. Ebers and J. L. Moll, "Large-signal behavior of junction transistors," Proc. IRE, vol. 42, pp. 1761–1772; December, 1954.

α_{1I} and α_{2I} are the alphas respectively for 1) junction 1 collecting and junction 2 emitting and 2) junction 3 collecting and junction 2 emitting. Note that for unity collection efficiency the total collected current must be less than the total emitted current so that

$$\alpha_{1I} + \alpha_{2I} \leqq 1. \tag{30}$$

For the case V_1, V_2, V_3 simultaneously different from zero, superposition applies and we have

$$I_1 = I_{s1}(e^{\beta V_1} - 1) - \alpha_{1I} I_{s2}(e^{-\beta V_2} - 1) \tag{31}$$

$$I_2 = \alpha_{1N} I_{s1}(e^{\beta V_1} - 1) - I_{s2}(e^{-\beta V_2} - 1) + \alpha_{2N} I_{s3}(e^{\beta V_3} - 1) \tag{32}$$

$$I_3 = -\alpha_{2I} I_{s2}(e^{-\beta V_2} - 1) + I_{s3}(e^{\beta V_3} - 1). \tag{33}$$

Our object is to solve for V_1, V_2, V_3 in terms of the currents, and this is done by inverting (31), (32), (33).

There results

$$I_{s1}(e^{\beta V_1} - 1) = \frac{I_1(1 - \alpha_{2N}\alpha_{2I}) - \alpha_{1I} I_2 + \alpha_{1I}\alpha_{2N} I_3}{A_0} \tag{34}$$

where

$$A_0 = 1 - \alpha_{2N}\alpha_{2I} - \alpha_{1N}\alpha_{1I} \tag{35}$$

$$I_{s2}(e^{-\beta V_2} - 1) = \frac{\alpha_{1N} I_1 - I_2 + \alpha_{2N} I_3}{A_0} \tag{36}$$

$$I_{s3}(e^{\beta V_3} - 1) = \frac{\alpha_{1N}\alpha_{2I} I_1 - \alpha_{2I} I_2 + I_3(1 - \alpha_{1N}\alpha_{1I})}{A_0}. \tag{37}$$

With two-terminal operation,

$$I = I_1 = I_2 = I_3 \quad \text{and} \quad V_s = V_1 + V_2 + V_3.$$

From (34), (36), (37),

$$V_s = \frac{1}{\beta} \ln \frac{I_{s2}}{I_{s1} I_{s3}} \frac{(IA_1 + I_{s1})(IA_3 + I_{s3})}{(IA_2 + I_{s2})} \tag{38}$$

where

$$A_1 = \frac{1 + \alpha_{1I}\alpha_{2N} - \alpha_{2N}\alpha_{2I} - \alpha_{1I}}{A_0} \tag{39}$$

$$A_2 = \frac{\alpha_{1N} + \alpha_{2N} - 1}{A_0} \tag{40}$$

$$A_3 = \frac{1 + \alpha_{1N}\alpha_{2I} - \alpha_{1N}\alpha_{1I} - \alpha_{2I}}{A_0}. \tag{41}$$

The interpretation of (38) requires an estimate of the sizes of A_1, A_2, A_3. Consider first the case where

$$\alpha_{1N} + \alpha_{2N} < 1. \tag{42}$$

In this case $A_2 < 0$, $A_1 > 0$, $A_3 > 0$ and (38) makes sense only for

$$IA_2 + I_{s2} > 0 \quad \text{or} \quad (\text{since } A_2 < 0) \tag{43}$$

$$I < \frac{I_{s2}}{|A_2|} = \frac{(1 - \alpha_{2N}\alpha_{2I} - \alpha_{1N}\alpha_{1I}) I_{s2}}{1 - \alpha_{1N} - \alpha_{2N}}. \tag{44}$$

Thus, if (42) holds, the current that flows at moderate voltages is a multiple of the saturation current and is small.

The other possibility

$$\alpha_{1N} + \alpha_{2N} > 1 \tag{45}$$

results in

$$A_0 > 0, \qquad A_1 > 0, \qquad A_2 > 0, \qquad A_3 > 0.$$

In this case the current can increase without limit at finite voltage. If the current is large compared to the saturation currents,

$$V_s \cong \frac{1}{\beta} \ln \frac{I_{s2}}{I_{s1} I_{s3}} I \frac{A_1 A_3}{A_2}. \tag{46}$$

Eq. (46) is essentially the voltage drop in a single forward biased junction if A_1, A_2, A_3 are of the order of $0.01-1$. For practical transistors, the only significant contribution of the A's to the voltage drop will be when $\alpha_{1N}+\alpha_{2N}$ is close to unity. If $\alpha_{1N}+\alpha_{2N}-1$ is of the order of 10^{-4} then V_s is about $10/\beta = 1/4$ volt greater than a forward biased diode.

Acknowledgment

The authors are pleased to acknowledge the aid of many of their associates who helped in carrying out the work reported in this paper. Particular thanks are due to C. S. Fuller who supplied some of the diffused material used in the experiments and contributed much of the necessary fundamental information on diffusion constants and solubilities of impurities in silicon. C. J. Frosch also made notable contributions to the diffusion technique.

Two-Terminal *P-N* Junction Devices for Frequency Conversion and Computation*

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Summary—**Design principles for semiconductor diodes are derived from the analysis of idealized *p-n* junctions. The analysis gives the superheterodyne conversion matrix and the large-signal admittance in terms of the small-signal diffusion admittances.**

Structures that minimize minority-carrier storage give minimum conversion loss under matched conditions in converting a high frequency to a low frequency, and are useful in logic circuits of computers. Examples are the emitter-base diode of a transistor and a small bonded or point contact.

Amplification and improved power-handling capabilities can be obtained in converting a low frequency to a high frequency, if the geometry favors storage of minority carriers near the junction. Such structures can also be used as pulse amplifiers.

Introduction

FREQUENCY-CONVERSION operations are of great importance in electrical communication. Information represented by a low-frequency electrical signal is usually converted to a high frequency for convenience in transmission. Conversion from a low frequency to a high frequency will be called "up-conversion." At the receiving end, the high-frequency signal must be converted to a low frequency. Conversion from a high frequency to a low frequency will be called "down-conversion."

Semiconductor diodes are now commonly used as down-converters, but it can be predicted that they will find increasing use as up-converters as their full possibilities become realized through appropriate design. Amplification of the signal power is possible in up-conversion; diodes designed for this amplification will also have relatively good power-handling capabilities.

Device theory opens the way to mathematical design procedures for the circuitry. This is particularly desirable in microwave circuits, where cut-and-try methods can mean large expenditures in the machine shop. The theory teaches how to make measurements that are economical in time and equipment and easy to interpret. The hypothesis that microwave diodes are *p-n* junction devices is encouraged by plausible explanations of some of the empirical findings.

In this paper, the concepts underlying *p-n* junction analysis are reviewed briefly, along with the approximations appropriate to the "strongly extrinsic" case. As long as these approximations are valid, the conversion matrix for superheterodyne operation depends rather simply upon the small-signal admittances. The maximum available gains for a few idealized structures illustrate the range of performance to be expected from *p-n* junction frequency converters. The pulse-circuit "turn-off" behavior of the idealized structures will be given. Equations for the idealized structures are in Appendix I. Appendix II is a list of symbols.

Principles of *P-N* Junction Analysis

The definition of a *p-n* junction is: A piece of semiconductor in which the fixed charge density changes sign. The fixed charge in "bulk" *p-n* junctions consists of impurity atoms in the semiconductor crystal. However, fixed charges on the surface of a semiconductor, if opposite in sign to the fixed charges in the bulk, can lead to *p-n* junctions with all the qualitative aspects of bulk *p-n* junctions. As far as is now known, the approximations to be used are just as valid for surface *p-n* junctions as for bulk junctions.

As an example, though, consider the bulk *p-n* junction illustrated in Fig. 1. The junction line is shown curved

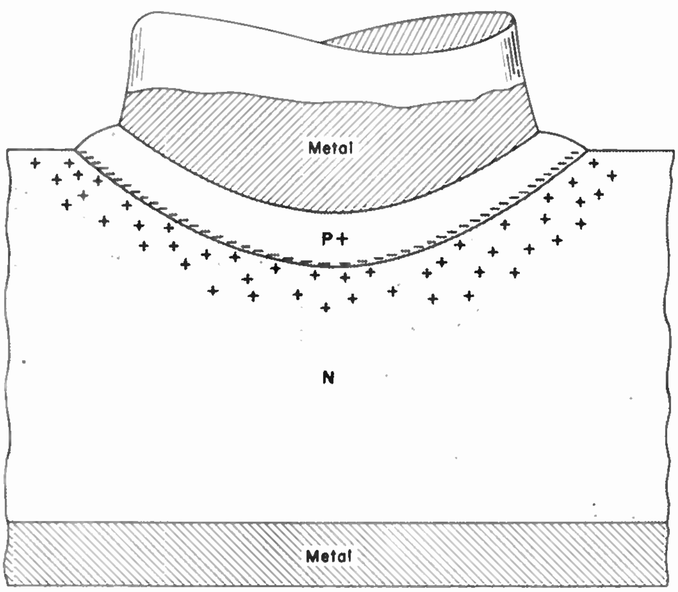

Fig. 1—Schematic diagram of n-$p+$ junction, showing space charge region.

because nonplanarity of the junction can be of importance. On either side of the junction, impurity charges are revealed because holes and electrons are swept out of this transition region. $P+$ is a symbol to indicate that the *p* region is very heavily doped with acceptor impurities, in comparison to the concentration of donors in the *n* region. In such a situation it is usually found to be permissible to neglect the flow of electrons across the junction and to consider only the hole current. This

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simplification is not essential to the analysis, but facilitates discussion without seriously limiting the generality of the conclusions.

The concentration of holes at the edge of the transition region depends in a nonlinear way upon the voltage drop across the transition region. This nonlinearity is responsible for frequency conversion in *p-n* junctions. To a good approximation, the holes meander at random in the neutral part of the *n* region; that is, they diffuse. Some holes are generated in the *n* region, others originate in the *p*+ region. Some recombine with electrons in the *n* region; others return to the *p*+ region. If a steady electric field is present, the diffusion equation may be generalized to take the field into account.

The design of the diffusion region will be discussed in this article almost to the exclusion of some considerations which have been emphasized elsewhere.[1] One of these is the displacement current across the junction, which can be represented by a transition region capacity C_T. The series "spreading resistance" R_S between the junction and the metal contacts is also important. Fig. 2 shows how these "parasitics" might be treated as

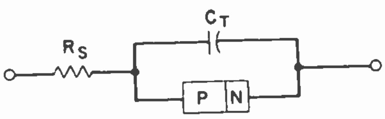

Fig. 2—Equivalent circuit of diode.

lumped elements in an approximate equivalent circuit for the diode (in general, they must be treated as distributed elements). These parasitics vary noticeably with operating conditions and act somewhat as frequency converters. But this action is much less than the frequency conversion of the "works" of the diode (the nonlinear diffusion admittance, denoted by a symbolic *p-n* junction), in the type of operation that is contemplated here.

Strongly-Extrinsic Approximation

Certain simplifying assumptions (instantaneous transition-region action, constant spreading resistance, constant transition-region capacity, and the implicit assumption that the edge of the transition region does not move) tend to become more accurate as the impurity doping in both the *p*+ and *n* regions is increased. The assumptions may therefore be said to apply to the "strongly-extrinsic" case, which is of interest for two reasons. First, the strongly-extrinsic analysis is quite accurate for many practical devices and can be used in the quantitative design of device and circuit. Second, quantitative studies of an idealized situation can be expected to give rise to qualitative concepts which retain some measure of validity in situations that depart from the ideal.

What is perhaps the most important such concept will be developed with the aid of an idealization that goes beyond the strongly-extrinsic approximations: The spreading resistance and the transition-region capacity will be neglected. In other words, the symbolic *p-n* junction of Fig. 2 will be treated as if it were a circuit element with accessible terminals. The results of such an analysis set upper limits on device performance, since R_S and C_T can degrade this performance but, as long as they are regarded as constants, cannot contribute any beneficial effects that could not be achieved with external passive elements.

An approximation will be adopted that has been used with much success in analyzing *p-n* junction transistors.[2] The hole concentration p_T in the *n*-region just at the edge of the transition region is given by

$$p_T = p_{NT} e^{\beta v}, \tag{1}$$

where v is the deviation of the barrier voltage from its equilibrium value. p_{NT} is the equilibrium concentration, and $\beta \equiv q/kT$. This relation is assumed to be instantaneous in comparison to the diffusion of carriers in the neutral region; thus, effects of the slowness of the diffusion process will be simply illustrated.

The exact requirements for the validity of (1) are not known, although its limitations in the steady-state case have been discussed.[3] It is plausible that thinness of the transition region favors the assumption of instantaneous action. In this connection, it may be remarked that anything ""instantaneous" invites the device designer to sacrifice some speed to obtain other more desirable properties. For example, a layer of high-resistivity semiconductor widens the transition region, thereby increasing reverse breakdown voltage and lowering capacity, in *p*-intrinsic-*n* power rectifiers[4] and *p-n-i-p* transistors.[5] The generalized diffusion equation is[6]

$$\frac{\partial p}{\partial t} = -\frac{p - p_N}{\tau} + D\nabla^2 p - \mu \vec{\nabla}\cdot(p\vec{E}). \tag{2}$$

The hole concentration is p; the diffusion constant D, mobility μ, and lifetime τ are for holes. It will be assumed that τ and $\vec{E}$ are independent of p and t, so that (2) is a linear partial differential equation. However, the equation may be resolved into separate equations for a dc component of p and one or more ac components. If variations in τ and $\vec{E}$ make the dc equation nonlinear, it is still possible that the ac equations, for a given dc bias, will be linear for moderately large ac values of p.

[1] H. C. Torrey and C. A. Whitmer, "Crystal Rectifiers," McGraw-Hill Book Co., Inc., New York, N. Y., sec. 4.6; 1948.

[2] W. Shockley, "The theory of *p-n* junctions in semiconductors and *p-n* junction transistors," *Bell Sys. Tech. J.*, vol. 28, pp. 435–489; July, 1949.

[3] M. Cutler, "Flow of electrons and holes through the surface barrier region in point contact rectification," *Phys. Rev.*, vol. 96, pp. 255–259; October 15, 1954.

[4] R. N. Hall, "Power rectifiers and transistors," Proc. IRE, vol. 40, pp. 1512–1518; November, 1952.

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

[6] Shockley, *op. cit.*, pp. 459 and 476. A list of symbols used in the present article is given in Appendix II.

One boundary condition arises from (1), which specifies the hole concentration at the edge of the transition region. For the other boundary condition, it is prudent to assume that p-p_N is proportional to its own gradient on some surface. That is to say, the boundary is characterized by a "surface-recombination velocity" which may be zero or infinite in the limiting cases of reflecting or absorbing surfaces. Then solutions for p-p_N may be multiplied by constants and added to each other to obtain other solutions.

In (1), one starts with the instantaneous voltage. The calculation is essentially completed by getting the instantaneous current density $\vec{J}$ from a solution of (2). Thus,

$$\vec{J} = -qD(\vec{\nabla}p)_T + q\mu(p\vec{E})_T \tag{3}$$

where the subscript T indicates a quantity evaluated at the edge of the transition region.

In frequency conversion, one is concerned with hole concentrations whose time dependence is given by

$$p = p_N + p_0 + \mathrm{Re}\sum_\omega p_\omega e^{j\omega t} \tag{4}$$

in which case (2) separates into equations of the form

$$0 = -(j\omega) + 1/\tau)p_\omega + D\nabla^2 p_\omega - \mu\nabla\cdot(p_\omega\vec{E}). \tag{5}$$

By solving these equations and applying (3), one can calculate the complex current i_ω at each frequency. For any $(P+)N$ junction, the current may be written

$$i_\omega = AqDp_{\omega T}/L_\omega \tag{6}$$

where $p_{\omega T}$ is the complex amplitude of the hole concentration at the edge of the transition region. The quantity L_ω may be called an "effective diffusion length" at frequency ω, for the structure in question. Values of L_ω for some ideal structures are given in Appendix I. In general, L_ω is a complex quantity.

The admittance measured with an ac voltage small compared to kT/q is called the small-signal admittance and is given by

$$Y_\omega = \beta AqDp_{NT}e^{\beta v_0}/L_\omega \tag{7}$$

where v_0 is the dc voltage. It will be convenient to write the large-signal results in terms of Y_ω.

Large-Signal AC Response

When an ac voltage comparable to or larger than kT/q is applied to a p-n junction, the current is not proportional to the voltage. Appreciable voltages and/or currents will then be present at harmonics of the fundamental frequency. The analysis of large-signal ac problems is complicated by the fact that the harmonic content is usually not given explicitly and must be determined from the properties of the p-n junction and the external circuit.

Some insight into these problems can be had by considering special cases. The harmonics are said to be short-circuited if all the harmonic voltages are zero. Then

$$v = v_0 + v_\omega \cos \omega t \tag{8}$$

with suitable choice of time zero. The Fourier expansion of (1) is then[7]

$$p_T = p_{NT}e^{\beta v_0}[I_0(\beta v_\omega) + 2I_1(\beta v_\omega)\cos\omega t + 2I_2(\beta v_\omega)\cos 2\omega t + \text{etc.}]. \tag{9}$$

The large-signal admittance is

$$Y_\omega(v_0, v_\omega) \equiv i_\omega/v_\omega = 2Y_\omega(v_0)I_1(\beta v_\omega)/\beta v_\omega \tag{10}$$

in terms of the small-signal admittance for the same dc voltage. For ordinary diodes (as opposed, for example, to one junction of a transistor when bias is applied to the other junction), (5) with $\omega=0$ may be used to calculate p_0. Then the dc characteristic, as affected by the ac voltage, is

$$i_0 = I_s[I_0(\beta v_\omega)e^{\beta v_0} - 1] \tag{11}$$

where $I_s = AQTDp_{N_T}/L_0$. Eqs. (10) and (11) may be regarded as describing the conversion to dc of a signal at frequency ω, or the slow modulation of a carrier frequency ω. In the first case, suppose v_ω is held constant. Then the dc output is fixed, according to (11). Also, for fixed v_ω, the in-phase ac current that must be taken from the source is then proportional to the small-signal conductance G_ω and should be as small as possible. The best type of diode is called a "variable-resistor" in the discussion of possible structures. On the other hand, slow modulation can best be accomplished by "variable-capacitor" structures, which will also be described.

For the case of open-circuited harmonics, it is found that

$$Y_\omega(v_0, v_\omega) = Y_\omega(v_0) \tag{12}$$

and

$$i_0 = I_s\{[1 + \tfrac{1}{4}(\beta v_\omega)^2]e^{\beta v_0} - 1\}. \tag{13}$$

The maximum value of βv_ω compatible with open-circuited harmonics is 2. (The maximum excursion of the instantaneous voltage is not limited but must be built up out of harmonics.)

For small values of βv_ω, the nature of the harmonic terminations is not very important and (11) and (13) become approximately equivalent. The open-circuit dc voltage is then

$$v_0 \approx -\tfrac{1}{4}\beta v_\omega^2 \quad (\beta v_\omega \ll 1) \tag{14}$$

and provides a simple check, in small-signal measurements, that v_ω is indeed small compared to $1/\beta$.

Superheterodyne Conversion Matrix

Superheterodyne operation is of great practical importance and lends itself to theoretical analysis because the signals are linearly related. The signals are treated

[7] Torrey and Whitmer, *op. cit.*, sec. 5.11

as having arbitrarily small amplitudes and are applied to or withdrawn from the diode in the presence of a local-oscillator drive at angular frequency b. The local oscillator is regarded as a power supply, not a signal. The signal (angular) frequencies are the upper sideband $b+s$, the lower sideband $b-s$, and the "intermediate" frequency s. When s is a much lower frequency than b, a down-converter results if the input signal is at frequency $b+s$ or $b-s$ and the output is at s. Microwave receivers usually contain such a down-converter. An up-converter can be used in a transmitter; the input would be at a low frequency s and the output is at either or both of the high frequencies.

Superheterodyne conversion has been analyzed[7] on the assumption that the instantaneous current is a function of the instantaneous voltage according to a relation of the form

$$i = I_s(e^{\beta v} - 1). \tag{15}$$

The present objective is to assume only an instantaneous relation, (1), between carrier concentration and voltage. The previous mathematical development can be used by making the substitutions

$$\begin{aligned} i + I_s &\to p_T \\ I_s &\to p_{NT}. \end{aligned} \tag{16}$$

Then the small-signal components of p_T can be written

$$\begin{Vmatrix} p_{b+s} \\ p_s \\ p_{b-s}^* \end{Vmatrix} = \beta p_{NT} e^{\beta v_0} \Big|_{\text{at } T} \begin{Vmatrix} I_0 & I_1 & I_2 \\ I_1 & I_0 & I_1 \\ I_2 & I_1 & I_0 \end{Vmatrix} \cdot \begin{Vmatrix} v_{b+s} \\ v_s \\ v_{b-s}^* \end{Vmatrix}. \tag{17}$$

The I's are modified Bessel functions of the first kind; their argument is βv_b, where v_b is the zero-to-peak value of the local-oscillator voltage. The assumption underlying this expression is that the harmonics of the local oscillator and the side bands of these harmonics are short-circuited.

Eqs. (6), (7), and (17) lead at once to the conversion matrix, from which the transmission properties of the converter can be calculated. One finds

$$\begin{Vmatrix} i_{b+s} \\ i_s \\ i_{b-s}^* \end{Vmatrix} = \begin{Vmatrix} Y_{b+s} & 0 & 0 \\ 0 & Y_s & 0 \\ 0 & 0 & Y_{b-s}^* \end{Vmatrix} \cdot \begin{Vmatrix} I_0 & I_1 & I_2 \\ I_1 & I_0 & I_1 \\ I_2 & I_1 & I_0 \end{Vmatrix} \cdot \begin{Vmatrix} v_{b+s} \\ v_s \\ v_{b-s}^* \end{Vmatrix}. \tag{18}$$

Thus, the conversion matrix is the product of two matrices. For open-circuited harmonics. the Bessel function matrix in (18) is replaced by

$$\begin{Vmatrix} 1 & \xi & 0 \\ \xi & 1+\xi^2 & \xi \\ 0 & \xi & 1 \end{Vmatrix}$$

where $\xi = \frac{1}{2}\beta v_b$. The effect of the series resistance R_s and transition-region capacity C_T is not included; it is a routine network calculation to do this for the signal frequencies.

However, R_s precludes exact short-circuiting of the harmonics. Open-circuited harmonics cannot be realized in practice, because of the transition-region capacity C_T. An exact calculation is therefore quite tedious. But for any terminations of the harmonics, the conversion matrix for the symbolic p-n junction of Fig. 2 will still be a product of two matrices. The first matrix will be the diagonal matrix of the small signal admittances—representing in a sense the linear diffusion process. The second matrix represents the nonlinear but instantaneous barrier action. If harmonic terminations are specified in terms of impedances, this matrix will depend upon the small-signal admittances at the harmonic frequencies, as well as upon the local-oscillator drive. If there are significant differences in the effective resistance in series with different locations on the junction (that is, if the equivalent circuit shown in Fig. 2 is not valid), each element of junction area must be considered as a separate frequency converter with a particular dc bias and harmonic termination.

To show what can, in principle, be done with p-n junctions, R_s and C_T will be neglected. The maximum available gain (MAG) will be used as a figure-of-merit. The MAG is the ratio of output signal power to available input signal power, for simultaneous conjugate matching of source and load. The concept of gain refers to a two-terminal-pair network; such can be made from the three-terminal-pair frequency converter by passive connections. For example, one of the sidebands can be short-circuited; this simple case was used in calculating the MAG's quoted below for the exemplary structures (the same results are obtained for open-circuited harmonics with one sideband open-circuited). The method of calculating MAG and matching admittances for a two-terminal-pair active network is given in the literature.[8]

The use of MAG as a figure-of-merit can be criticized from two essentially opposite points of view. One is that the unilateral gain U is a more elegant measure of amplification.[9] The latter considers the use of passive neutralizing circuits to eliminate feedback; the possibility that a frequency converter would have to be included in the neutralizing circuit makes U a less practical figure-of-merit for frequency converters than for single-frequency networks. Down-converters with MAG approaching unity have vanishing U, yet are highly desirable. The MAG and U are essentially equal for the up-converter structures that will be described, which shows that up-converter amplification does not rely on positive feedback.

The opposite criticism is that possible gains obtainable through positive feedback are overlooked in the use of MAG as a figure-of-merit. Since regenerative

[8] A. W. Lo, R. O. Endres, U. Zawels, F. D. Waldhauer, and C. C. Cheng, "Transistor Electronics," Prentice-Hall, Inc., Englewood Cliffs, N. J., pp. 100–102; 1955.

[9] S. J. Mason, "Power Gain in Feedback Amplifiers," *M.I.T. Res. Lab. of Elec.*, Tech. Rep. No. 257; 1953.

gains may be useful, particularly in narrow-band applications, this limitation on the conclusions should be kept in mind.

At a level of abstraction corresponding to the rest of the present discussion, all the noise in the symbolic *p-n* junction of Fig. 2 is shot noise. The author has developed the analytical machinery for calculating shot noise in superheterodyne conversion by *p-n* junctions. This work is in too preliminary a stage to be reported at present. However, it appears unlikely that the conclusions based on MAG will be greatly modified by noise considerations.

Exemplary Structures

As far as device structure is concerned, the maximum available gain depends upon the relative values of the small-signal admittances at the various frequencies. This frequency dependence is determined by generalized diffusion, which involves three processes: recombination, diffusion, and drift, in the order in which the corresponding terms appear on the right-hand side of (2). In the idealized examples to be discussed, recombination and drift will be neglected.

Fig. 3 shows the frequency dependence of the small-signal admittance when an absorbing surface is placed a distance w from an $(P+)N$ junction (equations for this and subsequent figures are given in Appendix I).

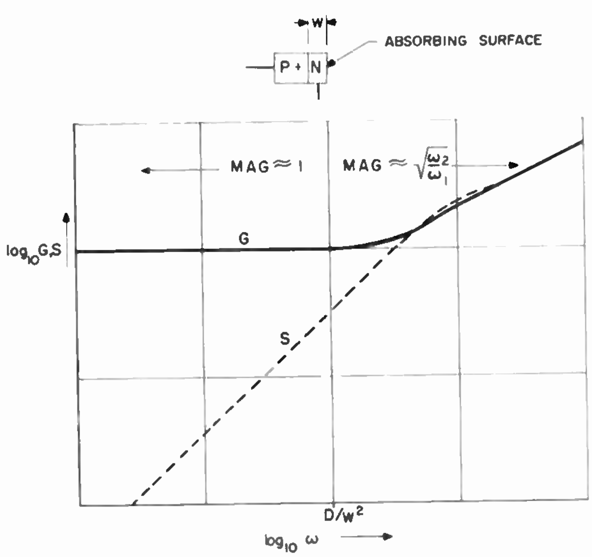

Fig. 3—Frequency dependence of the small-signal admittance of a planar "variable-resistor" structure.

This situation can be realized in a transistor-like structure in which there is an ac short-circuit from collector to base. The admittance is for the emitter-to-base diode and is determined by diffusion of holes in the *n*-region. In converting between two frequencies, both of which lie far to the left or low-frequency side of this plot, the MAG approaches unity. In this low-frequency range, the diode could be called a "variable resistor," because its admittance is predominantly real and constant with frequency; the admittance varies exponentially with dc bias voltage.

At higher frequencies, the MAG is the square root of the ratio of the output frequency to the input frequency. Thus, a moderate amount of gain could be obtained in an up-converter, where the output frequency is higher than the input frequency. On the other hand, use as a down-converter in this region would incur a wholely undesirable loss of signal power.

The division between low and high frequency is determined by the width w. In the absence of the absorbing barrier, the ac diffusion current at angular frequency ω would fall off exponentially in a distance $\sqrt{D/\omega}$. If w is much larger than this distance, the absorbing barrier will not be effective in producing variable-resistor behavior. To put the center of Fig. 3 at 10 kmc would require that w be about 2×10^{-5} cm (for holes in germanium or electrons in silicon).

Simple diffusion, together with an absorbing barrier, is involved in the variable-resistor structure of Fig. 3. Drift in an electric field also can lead to variable-resistor behavior, provided the electric field removes minority carriers from the junction. Electric fields accompany gradients in the concentration of fixed charge in semiconductors. Thus, variable resistor action can be obtained in a $(p+)n$ junction in which the donor concentration in the n region decreases with distance from the junction. Recombination (short minority-carrier lifetime) also favors variable-resistor action, but is not as controllable as the concentrations of donors and acceptors. Moreover, the recombination process can fail when it is most needed—injected carriers in sufficient numbers "swamp" the recombination centers.

An opposite type of structure is illustrated in Fig. 4.

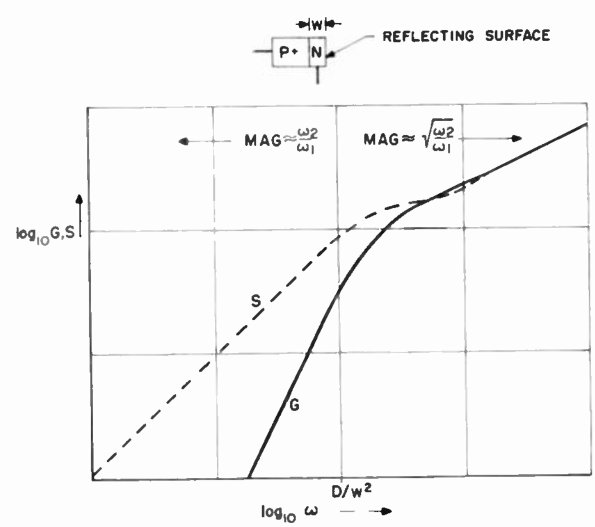

Fig. 4—Frequency dependence of the small-signal admittance of a planar "variable-capacitor" structure.

A reflecting surface is placed a distance w from a $(p+)n$ junction. If the emitter and collector of a *p-n-p* transistor of base width $2w$ were connected together, the plane of symmetry between the emitter and collector would be equivalent to a reflecting surface. At high frequencies the reflecting surface has no effect and the maximum available gain is the square root of the frequency ratio, as before. At low frequencies, the admittance is predominantly susceptive. The susceptance varies exponentially with dc voltage and for a given voltage is proportional to frequency. In this frequency range it seems appropriate to call the device a "variable capacitor."[10] The variable capacitor has a gain equal to the ratio of the output frequency to the input frequency. Large gains are possible in up-conversion. Furthermore, the variable capacitor has the well-known advantage of reactance control of ac power: it does not dissipate in itself much of the power that it modulates.

At very low frequencies, even this variable-capacitor structure becomes a variable resistor, because of recombination. Hence, large but not infinite gains are possible in going from dc to a high frequency.

A graded junction (one in which the fixed charge density increases with increasing distance from the junction) is a variable-capacitor structure in which drift, rather than diffusion, predominates.[11] Still another variable-capacitor structure is a *p-i-n* diode (in which the *i* region thickness should be less than $\sqrt{D/\omega}$) operated at biases such that most of the *i* region is essentially free of space charge—that is, for forward biases or not very high reverse biases.

Some features of common diodes can be explained with the help of Fig. 5. It is the small-signal admittance

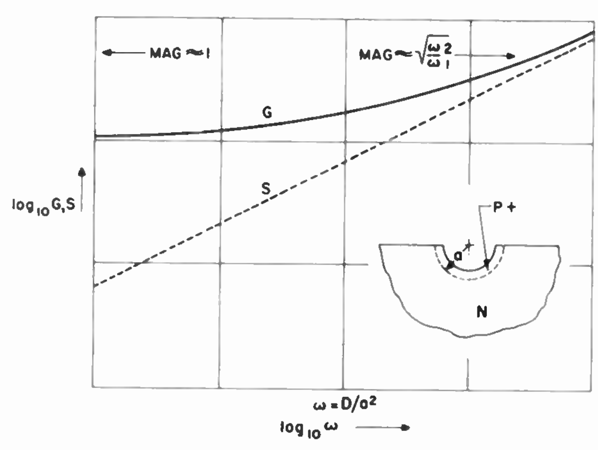

Fig. 5—Frequency dependence of the small-signal admittance of a hemispherical "variable-resistor" structure.

of a hemispherical $(p+)n$ junction. At low frequencies, variable-resistor action is obtained even though recombination is neglected. The reason for this is that carriers that diffuse a considerable distance from the junction have very little chance of finding their way back. They can go to infinity, if necessary, to recombine. A high-frequency hole current, however, does not get far enough from the junction to behave any differently than for a planar junction.

The results of this section can be summarized as follows. Variable-resistor behavior minimizes the loss in down-conversion and can be obtained by reducing minority carrier storage. Variable-capacitor behavior maximizes gain in up-conversion and requires that minority carriers be stored near the junction.

These two kinds of behavior do not exhaust the possibilities of two-terminal semiconductor devices. The use of a *p-n-p* structure as a two-terminal negative resistance has been proposed.[12] Among other things, such a structure could provide local-oscillator power for the frequency conversion diodes that have been discussed here.

Applications to Practical Frequency Converters

For frequencies below 500 mc, extensive single-frequency experiments on transistors have confirmed *p-n* junction theory, so that there can be little doubt of the applicability of the frequency conversion analysis given in this paper. However, some of the most important applications of two-terminal devices involve higher frequencies, so that experimental evidence for *p-n* junction behavior at microwave frequencies is of particular interest.

Whatever the structure (cf. Figs. 3, 4, and 5), the theoretical MAG tends to $\sqrt{\omega_2/\omega_1}$ if both the input and output frequencies ω_1 and ω_2 are sufficiently high. With *p-n* junction up-converters of this character, conversion gains as high as 5.7 db have been obtained with a 75 mc input and a 6175 mc output.[13] This result compares favorably with the theoretical limit of $\sqrt{6175/75}$, or 9.6 db, for this type of junction. A variable-capacitor junction could give up to 19.2 db gain in this situation.

Another probable manifestation of minority carrier storage is "reciprocity failure" in germanium diodes. "Full reciprocity" and "weak reciprocity" have been defined as properties of the conversion matrix.[14] According to the *p-n* junction conversion matrix (18), full reciprocity holds if the device is a variable resistor at all the signal frequencies, whereas variable-resistor action at the intermediate frequency *s* is sufficient for weak reciprocity.

Particularly marked reciprocity failures have been observed for welded-contact germanium rectifiers made by H. Q. North.[15] These diodes probably correspond quite closely to the hemispherical junction shown in Fig. 5 and discussed under Exemplary Structures. A typical value of the radius a is 2.5×10^{-4} cm. For frequencies less than $D/2\pi a^2 \approx 100$ mc, these diodes should

[10] A *PN*-junction variable capacitor of another sort, based on the variation in C_T, was described by J. O'Connell and L. J. Giacoletto at the 1st Annual Technical Meeting of the IRE-PGED at Washington, D. C., October, 1955.

[11] Shockley, *op. cit.*, p. 476. A variable-capacitor structure based on drift in an electric field is analyzed.

[12] W. Shockley, "Negative resistance arising from transit time in semiconductor diodes," *Bell Sys. Tech. J.*, vol. 33, pp. 799–826; July, 1954. Further information has been submitted to the *J. Appl. Phys.* by G. Weinreich.

[13] N. Bronstein and E. R. Showers (private communication).

[14] Torrey and Whitmer, *op. cit.*, sec. 5.5.

[15] *Ibid.*, ch. 13.

behave as variable resistors. Thus weak reciprocity (within the usual errors of such measurements) is to be expected for ordinary intermediate frequencies. But the 10 kmc conductance will be about 10 times the low-frequency conductance. The nonreciprocal conversion matrix obtained [from (18)] for such a frequency-dependent admittance can be shown to exhibit negative impedance to the low-frequency or high-frequency signals, for suitable passive terminations. The low-frequency negative impedance was observed directly.

The welded-contact diodes were also observed to give conversion gain as down-converters. This result is not in contradiction to the result that the best MAG is unity in down-conversion by a *p-n* junction; it just shows that negative impedance gain has been excluded in the calculation of MAG. The bandwidth of down-converters with gain is doubtless quite limited. The theoretical conversion matrix provides a basis for relating gain and bandwidth. Experimentally, increased noise seems to be another penalty of this type of operation. It will be very interesting to see if a shot-noise theory shows this effect to be of a fundamental nature.

In empirically optimized point-contact down-converter diodes, the higher the frequency, the smaller the point size. This result has previously been attributed to the spreading resistance R_s and transition region capacity C_T. The product $\omega R_s C_T$ should be less than unity to avoid serious losses. For a 10 kmc diode, this rule suggests a contact radius of 2×10^{-3} cm or less.[16] The apparent contact radius of such diodes is about 3×10^{-4} cm. A simple argument for the smaller point size is that an improved $\omega R_s C_T$ product makes tuning less critical. However, it is interesting to see how this point size fits into the *p-n* junction picture.

The first question in setting up a model of a surface *p-n* junction is whether the forward current consists predominantly of carrier injection from the bulk into the surface layer or *vice versa*. In the first case, the generalized diffusion of minority carriers will occur in the surface layer and will depend upon the little-understood properties of the metal-semiconductor contact which is a boundary of the surface layer. However, the contrary is true in one well-known example: transistor action in an *n*-type germanium point-contact transistor depends upon injection of holes from the surface into the bulk.

If injection into the bulk continues to be of importance as the impurity density is increased from the values (10^{14}–10^{16} per cm^3) used in transistor fabrication to the values (10^{17}–10^{18} per cm^3) typical of microwave diodes, the effect of point size can be discussed with reference to Fig. 5. The figure pertains to a hemispherical junction, while the point contact is more nearly a circular flat area, but the principle is the same in either case. Variable-resistor action can be approached by making the contact small enough so that minority carriers that have been stored for an appreciable time, compared to the period of the frequency involved, have little chance of diffusing back to the contact. According to this argument, it would be advantageous in detecting 10 kmc to have a contact radius less than $\sqrt{D/\omega}\approx2\times10^{-5}$ cm for holes diffusing in germanium or electrons in silicon. Thus, the typical contact radius is too large to insure variable-resistor action. One would expect a conversion loss of about 10 db to any intermediate frequency below 100 mc, if minority carrier injection into the bulk were fully effective, in addition to losses from R_s and C_T. Since conversion losses of 5 db are not uncommon, some mechanism must operate to reduce minority carrier storage.

A postulate that has been used to explain the dc characteristics is that current flow is not uniform over the contact but flows most readily at certain spots. Such spots could very well be small enough to give variable-resistor behavior.

The field associated with the flow of rectified current through the spreading resistance may aid in removing minority carriers from the junction. Another possibility is that injected carriers recombine rapidly; little is known of minority carrier lifetimes (except that they are short) in the low resistivity materials used in microwave diodes. Low resistivity favors injection from the bulk into the surface; then variable-resistor action implies that the surface layer has the capability of annihilating minority carriers.

The discussion of the point-contact rectifier is inconclusive, largely because the structure itself is a matter of speculation. But it is clear from the experiments on up-converters that minority carrier storage is possible at microwave frequencies. Hence, a down-converter made by techniques for producing known impurity distributions should be designed to minimize minority carrier storage.

Pulse Circuit Applications

The variable resistor desired for down-conversion has little tendency to store minority carriers near the junction. The variable resistor therefore has good turn-off properties in logic circuits. A hemispherical junction of about 1 mil diameter should be a good enough variable resistor for megacycle computing rates. Hence, it is not surprising that gold-bonded diodes have found favor. However, the planar variable-resistor structure of Fig. 3 has a "snappier" turnoff than the hemispherical junction (neglecting recombination in both cases). Fig. 6 shows the transient current for these two cases, relative to the difference between the initial and final currents, for a sudden change in voltage. The behavior of actual diodes is, of course, considerably modified by spreading resistance and transition-region capacity.[17]

[16] *Ibid.*, p. 98.

[17] B. Lax and S. F. Neustadter, "Transient response of a *p-n* junction," *J. Appl. Phys.*, vol. 25, pp. 1148–1154; September, 1954.

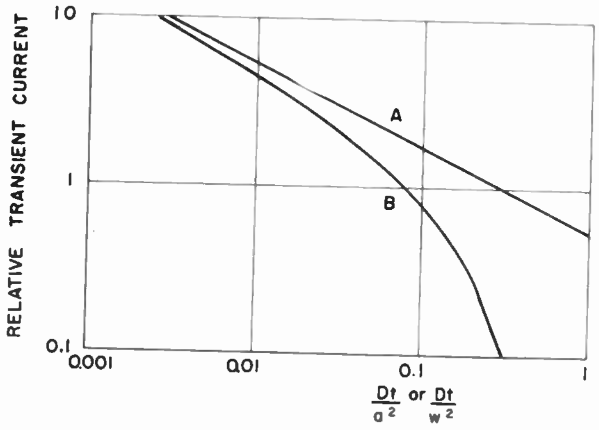

Fig. 6—Transient current in variable-resistor structures (relative to difference between initial and final currents), for a voltage step. For infinite minority carrier lifetime. (a) Hemispherical structure of Fig. 5; (b) planar structure of Fig. 3.

Workers at the National Bureau of Standards have found that carrier-storage effects in diodes can be used to amplify pulses.[18] In this type of operation the diode acts as a time-division transistor—the same junction serves alternately as emitter and collector. Thus, the time-domain point of view makes it easier to understand how a diode that emphasizes carrier storage can give amplification. Variable-capacitor structures should be used in pulse amplifiers. An equation for the transient current of the structure shown in Fig. 4 is given in Appendix I. This transient current is infinite relative to the ultimate change in current, which is zero (since the initial and final currents for a variable capacitor without recombination are both zero).

Conclusion

Up-converters with gain and good power-handling capabilities can be made with *p-n* junction diodes that store minority carriers within a distance $\sqrt{D/\omega}$ from the junction. The same or similar types of diodes can be used as pulse amplifiers.

Minority-carrier storage should be minimized in diodes for down-conversion, just as in logic-circuit diodes for computers.

Time-domain concepts (*e.g.*, "minority-carrier storage") are useful in describing and understanding the types of devices that are desirable in these applications. On the other hand, the frequency dependence of the small-signal admittance characterizes computer diodes no less than frequency-conversion diodes.

Appendix I

Equations for Graphs

The frequency dependence of Y_ω is contained in L_ω, which can be calculated from (2) and (3) for various boundary conditions. In the cases considered here, the current density is uniform over the area A; otherwise the calculation of L_ω requires a suitable averaging over the area. The graphs give $1/L_\omega$ for zero bulk recombination; that is, for $\tau = \infty$.

To calculate L_ω, one first seeks a solution p_ω of (5), satisfying the boundary conditions. Then L_ω is gotten by taking the logarithmic gradient of p_ω at the edge of the transition region:

$$\frac{1}{L_\omega} = -\vec{n}\cdot\left(\frac{\vec{\nabla} p_\omega}{p_\omega}\right)_T \qquad (19)$$

where $\vec{n}$ is a unit vector normal to the junction and pointing into the *n*-region.

For the planar variable resistor shown in Fig. 3,

$$\frac{1}{L_\omega} = \sqrt{\frac{j\omega + 1/\tau}{D}} \coth w\sqrt{\frac{j\omega + 1/\tau}{D}}. \qquad (20)$$

The variable capacitor graph in Fig. 4 is based on

$$\frac{1}{L_\omega} = \sqrt{\frac{j\omega + 1/\tau}{D}} \tanh w\sqrt{\frac{j\omega + 1/\tau}{D}}. \qquad (21)$$

For the hemispherical case of Fig. 5,

$$\frac{1}{L_\omega} = \frac{1}{a} + \sqrt{\frac{j\omega + 1/\tau}{D}}. \qquad (22)$$

The transient current can be obtained from the admittance, or *vice versa*, with the aid of a table of Laplace transforms.[19] The transient current can also be calculated directly from (2) and (3). For the planar variable resistor,

$$i(t)/[i_f - i_i] = \theta_3\left(0, \frac{\pi D t}{w^2}\right) - 1 \qquad (23)$$

where $i(t)$ is the transient current, i_f the final current, and i_i the initial current; t is the time after the sudden change in junction voltage and θ_3 is a theta function.[20] The corresponding expression for the hemispherical case is

$$i(t)/[i_f - i_i] = \frac{a}{\sqrt{\pi D t}}. \qquad (24)$$

Eqs. (23) and (24) are for $\tau = \infty$ and are plotted in Fig. 6.

The transient current for the planar variable capacitor with $\tau = \infty$ is

$$i(t) = \frac{AqDp_N}{w}\left(e^{qv_f/kT} - e^{qv_i/kT}\right)\theta_2\left(0, \frac{\pi D t}{w^2}\right) \qquad (25)$$

where v_i and v_f are the initial and final voltages.

[18] "Diode Amplifiers," *Electronic Design*, pp. 24–25; October, 1954.

[19] B. van der Pol and H. Bremmer, "Operational Calculus," Cambridge Univ. Press, Cambridge, Mass., pp. 383 and 405; 1950. The results necessary to set up a correspondence between the frequency domain and time domain in the cases treated here are given.

[20] E. L. Steele, "Charge storage in junction diodes," *J. Appl. Phys.*, vol. 25, pp. 916–918; July, 1954. Series expansions and graphs for the case of finite lifetime are given.

Appendix II

List of Symbols

a—radius of spherically-symmetric junction
A—area of junction
b—local-oscillator angular frequency
C_T—transition-region capacity
D—diffusion constant for holes
E—electric field
G—conductance
i—current
i_0—dc current
i_ω—zero-to-peak complex ac current at angular frequency ω
I_s—absolute value of the reverse saturation current
I_0, I_1, I_2—modified Bessel functions of the first kind
j—$\sqrt{-1}$
J_p—hole current density at junction
k—Boltzmann constant
L_ω—effective diffusion length for holes at frequency ω (in a given structure)
MAG—maximum available gain of a 2-terminal-pair linear network
n—type of semiconductor in which negative electronic charge carriers (electrons) predominate
$\vec{n}$—unit vector normal to junction
p—concentration of holes (number density)
p_N—equilibrium value of p in an n-type semiconductor
p_{T_N}—value of p at the boundary between the neutral N region and the space-charge (transition) region
p—type of semiconductor in which positive electronic charge carriers (holes) predominate
$p+$—p-type material in which the hole concentration is much larger than the carrier concentration in some of the other regions of a given structure.
q—absolute value of the charge of an electron
R_s—series resistance of a diode
s—angular frequency of the low-frequency ("intermediate frequency") signal in a superheterodyne frequency converter.
S—susceptance
t—time
T—absolute temperature
T_N—subscript denoting that a quantity is to be evaluated at the boundary between the neutral n-region and the space-change (transition) region
U—unilateral power gain.
v—deviation from equilibrium of the voltage across the transition region
v_0—dc value of v
v_ω—zero-to-peak ac value of v.
w—width of n-region in planar $np+$ structures
Y_ω—small-signal admittance (for specified v_0) at angular frequency ω
$\mathcal{Y}_\omega$—large-signal admittance at angular frequency ω
β—$q/kT = 39$ volt^{-1} at 298°K
θ_2, θ_3—theta functions[21]
μ—mobility of holes
ξ—$\frac{1}{2}\beta v_b$, a parameter for the case of open-circuited harmonics
τ—lifetime of holes in n-region
ω—angular frequency

Acknowledgment

Discussions of these subjects with A. E. Bakanowski have been very helpful. The procedure for calculating maximum available gain was provided by R. M. Ryder before "Transistor Electronics,"[8] was published. The terms "up-converter" and "down-converter" were suggested by M. C. Waltz.

[21] *e.g.* B. van der Pol and H. Bremmer, *op. cit.*, p. 236.

Correspondence

Radar Echoes from Meteor Trails Under Conditions of Severe Diffusion*

A simple expression is deduced for the power of the radio echo from a meteor trail when diffusion of the trail is predominant. It is shown that the echo is proportional to the sixth power of the wavelength and inversely proportional to the fourth power of the range. For a given meteor velocity there is a critical height above which the effects of diffusion become serious. This diffusion ceiling is given for various wavelengths between 0.5m and 16m.

Herlofson[1] has shown that the amplitude reflection coefficient, r, per unit length of a cylindrical meteor trail is given by the expression

$$r = r_0 \exp\left(-\frac{16\pi^2 Dt}{\lambda^2}\right) \tag{1}$$

where D is the diffusion coefficient of the trail, t is the time which has elapsed since the passage of the meteoroid and λ is the wavelength of the radar. This equation applies to trails with a line density of $<10^{14}$ electrons m^{-1}. The reflection coefficient has therefore fallen to $1/e$ of its initial value at a distance l behind the meteoroid, moving at velocity v, when

$$l = \frac{v\lambda^2}{16\pi^2 D}. \tag{2}$$

If the line density of electrons in the trail is q, and the scattering cross section of an electron is σ_e, then the equivalent target area of l meters of trail is $(lq)^2\sigma_e$, since the electrons are scattering coherently.

The power, p, returned to a receiver may be found from the radar formula

$$p = \frac{G^2P\lambda^2}{64\pi^3R^4}\cdot\frac{\lambda^4 v^2 q^2\sigma_e}{16^2\pi^4D^2}\ \text{watts} \tag{3}$$

where P is the peak power of the transmitter and G is the power gain of the antenna over an isotropic radiator. The scattered power is therefore proportional to λ^6 and inversely proportional to R^4. To compare this signal with the signal predicted by the Lovell-Clegg formula we may rewrite (3) as

$$p = 2.54 \times 10^{-32}\,\frac{PG^2\lambda^3 q^2}{R^3}\cdot\frac{1}{\eta^2}. \tag{4}$$

The first term is the original expression of Lovell and Clegg[2] and the term $1/\eta^2$ is an attenuation factor where

$$\eta = \frac{16\pi^2 DR^{1/2}}{\sqrt{2v}\lambda^{3/2}}. \tag{5}$$

Eq. (5) agrees with direct numerical integrations that have been made using the complex Fresnel integral[3] and may be deduced from an expression given by Eshelman[4] if his integration limits are changed to cover one-half of the first Fresnel zone. It is valid for $\eta > 1$ and a particular value of η defines a particular echo shape as shown in Fig. 1. It can be seen that the Fresnel pattern becomes severely distorted when $\eta > 2$. Evans[5] in a radio echo experiment at a wavelength of 4m has measured the height and velocity of meteors. Very few echoes were recorded with velocity >45 km and height >103 km which is the region where $\eta > 2$. Greenhow and Neufeld[6] have shown that the diffusion coefficient, D, is related to height h by the relation

$$\log_{10} D = 0.0679 \times 10^{-3}h - 5.663. \tag{6}$$

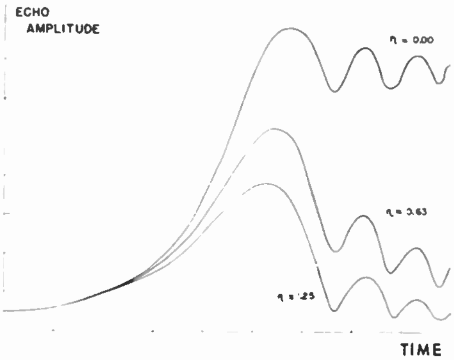

Fig. 1—Echo shape for various values of attenuation.

For a given wavelength, λ, range of trail, R, and meteor velocity, v, there is a critical height h_c meters above which Fresnel zones will tend to disappear. This diffusion ceiling may be found by putting $\eta = 2$, and substituting for D in (5):

$$h_c = 14.7 \times 10^3\left(\log_{10}\frac{v\lambda^{3/2}}{R^{1/2}} + 3.916\right). \tag{7}$$

The diffusion ceiling is plotted for various wavelengths and a range of 1.5×10^5 meters in Fig. 2. It can be seen that the ceiling is lower for low velocity meteors.

When $\eta \gg 2$ the trail begins to lose its property of specular reflection. That is to say the echo is no longer stationary in range but will present a moving target over a small angle on either side of the minimum range position.

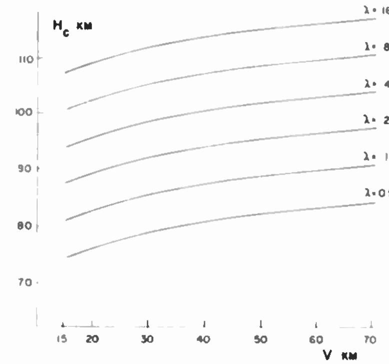

Fig. 2—The diffusion ceiling for $\eta = 2$.

In the interest of experiments that might be conducted in the region where diffusion of the trail is serious, values of the amplitude attenuation coefficient η are tabulated in Table I as a function of range and wavelength for meteors with a velocity of 40 km. The diffusion coefficient D has been taken as 10 m^2 sec^{-1}, which corresponds to a height of 98 km. For meteors with a velocity of 70 km the values of η will be greater by a factor of approximately 2: for meteors of velocity 20 km η will be smaller by a factor of 2.

TABLE I

AMPLITUDE ATTENUATION COEFFICIENT η

R \ λ	4m	3m	2m	1m	0.5m
100 km	1.10	1.70	3.12	8.83	25.0
500 km	2.47	3.80	6.98	19.7	55.8
1000 km	3.49	5.37	9.87	27.9	78.9
2000 km	4.93	7.60	13.9	39.5	112

The foregoing treatment has neglected the initial diameter of the trail on formation and the effect of fragmentation[7] of the meteor body. These factors are at present not known with certainty. The initial diameter will probably be of the order of the mean free path at the height of the trail, and will increase the value of η at the shorter radar wavelengths. Fragmentation of the meteor body will separate the ionizing particles along the trail, tending to reduce the values of η. It would be of considerable interest to obtain backscatter echoes at wavelengths of less than 1 meter to obtain information on the fragmentation processes and initial diameter of meteor trails.

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* Received by the IRE, April 20, 1956. This work was carried out partly while the author was consulting for Sylvania Electronic Systems Div., Waltham, Mass.

[1] N. Herlofson, "Plasma resonance in ionospheric irregularities," *Arkiv Mat. Astr. Fysik*, vol. 3, pp. 247–297; 1951.

[2] A. C. B. Lovell and J. A. Clegg, "Characteristics of radio echoes from meteor trails," *Proc. Phys. Soc. A*, vol. 60, pp. 491–498; 1948.

[3] T. R. Kaiser, "Radio echo studies of meteor ionization," *Phil. Mag. Suppl. 2*, pp. 495–544; October, 1953.

[4] V. R. Eshleman, "The effect of radar wavelength on meteor echo rate," IRE TRANS., vol. AP-1, pp. 37–42; October, 1953.

[5] S. Evans, "Scale heights and pressures in the upper atmosphere from radio echo observations of meteors," *Monthly Notices Roy. Astron. Soc.*, vol., 114 pp. 63–73; 1954.

[6] J. S. Greenhow and E. L. Neufeld, "The diffusion of ionized meteor trails in the upper atmosphere," *J. Atm. and Terrest. Phys.*, vol. 6, pp. 133–140; March, 1955.

[7] L. G. Jacchia, "The physical theory of meteors VIII, fragmentation as cause of the faint-meteor anomaly," *Astrophys. J.*, vol. 121, pp. 521–527; March, 1955.

Contributors

Guy F. Barnett (A'47–M'47) was born May 28, 1910 in Pittsburgh, Pa., and studied at M.I.T. and Rutgers University, receiving the B.S. degree in physics from the latter in 1944. He was employed at the Johns-Manville Research Laboratories from 1940 to 1945. Later that year, he joined the Philco Research Division where he engaged in microwave antenna and microwave tube research. In 1950 he joined the staff of the National Bureau of Standards where his work was concerned with microwave tube development. He returned to Philco's Lansdale Tube Division Development Laboratories in 1951 and since that time has been concerned with cathode ray tube development.

G. F. Barnett

Mr. Barnett is a member of the American Physical Society and holds two Philco patents in traveling-wave tubes.

❖

R. L. Bell was born on September 5, 1924, in Alnwick, Eng. He received the B.Sc. and Ph.D. degrees at Durham University in electrical engineering in 1945 and 1948, respectively. He was associated with the Research Laboratories, General Electric Company, Ltd., Wembley, Eng., engaged in research on receiving-tube development, during the period from 1948 to 1950.

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Dr. Bell is presently in the Services Electronics Research Laboratory, Baldock, Eng., where he is devoting his time to work on microwave gas discharge and noise problems.

❖

Frank J. Bingley (A'34–M'36–SM'43–F'50) was born in Bedford, Eng., on November 13, 1906. He received degrees in mathematics and physics in 1926 and 1927 respectively from the University of London. Thereafter, he was employed by the Baird Television Company of London and New York, in October, 1927.

F. J. Bingley

Mr. Bingley joined Philco Corporation in 1931 where he has been associated with the development of transmitting and receiving equipment, as well as with television systems engineering. Presently, he is engaged in an extensive program of research on color television. Color television and colorimetry have been the subjects of numerous articles published by Mr. Bingley.

He is a member of the Franklin Institute and has served on many industry committees concerned with the development of television standards including the original RMA Committee on Television, the Radio Technical Planning Board, and the National Television System Committee.

❖

Ralph A. Bloomsburgh, Jr. (M'46) was born December 8, 1918 in New York, N. Y. He received the B.S. degree in physics from Wesleyan in 1939. Following graduate work at M.I.T. he was employed by Kollmorgen Optical Corporation. In 1943 he was commissioned in the Naval Reserve and appointed as a lecturer in the Radar School at M.I.T.

R. A. Bloomsburgh

Since joining Philco in 1946 he has been largely concerned with optics, electron optics, and cathode-ray tube displays for monochrome and color television receivers. He holds a Philco patent on cathode ray tubes. Mr. Bloomsburgh is a member of the Optical Society of America.

❖

Wilson P. Boothroyd (SM'47) was born February 14, 1916 in Lawrence, Mass. He received the B.S. from Rhode Island State College in 1937, and joined Philco Corporation the same year. He did engineering work in the patent department and in 1939 was admitted to practice before the U. S. Patent Office. He transferred to the Research Division in 1941 and was active in projects covering airborne radar, radar beacons, range finding equipment, communication multiplex equipment, microwave radio relays and color television. In 1950 he organized the Advanced Development Laboratory in the Radio and Television Division, becoming Chief Engineer of Advanced Development in 1953.

W. P. Boothroyd

Mr. Boothroyd holds seven Philco patents in radar, eight in communications, one in television, and one in refrigeration. He is a member of Panel 16 of the National Television System Committee.

❖

James S. Bryan (A'52) was born February 11, 1926 in Louisville, Ky. He received the B.S.E.E. and the M.S.E.E. degrees in 1952 from M.I.T. He joined the Research Division of Philco Corporation in 1950, where he has engaged in color television display research, particularly in electron optics, and has designed a superior electron-trajectory tracer. This device aids in the study of the design of electron guns for all types of cathode ray tubes used in radar systems and television tubes. Mr. Bryan has been instrumental in substantially improving cathode ray tube spot size, and is one of the inventors of the Apple system. He holds two Philco patents in television.

J. S. Bryan

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❖

For a photograph and biography of Richard G. Clapp, see p. 380 of the March, 1956 issue of Proceedings of the IRE.

❖

Edgar M. Creamer, Jr. (A'45) was born August 26, 1921 in Chester, Pa. He received the B.S.E.E. from Drexel Institute of Technology in 1943. Since joining Philco Corporation the same year, he has been active in the development of uhf airborne television systems, microwave relays, time multiplexed terminal equipment, and television system transient studies. From 1950 to 1953 he was a project engineer directly responsible for color receiver circuit research, with special emphasis on Apple displays and receivers. He is one of the inventors of the Apple system. Later, he was manager of the Engineering Services Division laboratory facilities dealing with licensing activities in Apple color receivers. At present he is a group engineer in the Advanced Development Laboratory of the Television Division, and is concerned with monochrome television problems.

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George A. Fedde (A'52) was born May 29, 1927 in Washington, D. C. He received the B.S.E.E. and the M.S.E.E. degrees in 1951 from M.I.T. The same year he was employed by Philco Corporation's Research Division where he specialized in the development of color television systems and receivers, and the index-system development of the Apple system. After transferring to the Television Division Advanced Laboratory in 1953 he worked on color IF circuits and synchronizing circuits for the Chromatic receiver, Aperture Mask receiver, and the Apple receiver.

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James M. Goldey (M'56) was born on July 3, 1926 in Wilmington, Del. After two years of military service he entered the University of Delaware and received the B.S. degree in physics in 1950. He received the Ph.D. degree from M.I.T. in 1955 for an investigation of electron and hole effective masses in germanium. Dr. Goldey joined Bell Telephone Laboratories in October, 1954 and since that time has worked on transistor development. He is a member of the American Physical Society, Sigma Xi, and Phi Kappa Phi.

J. M. Goldey

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Mervyn Hillier was born in Bristol, Eng., on January 16, 1924. In 1942 he joined the Admiralty Signal Establishment, then located in Bristol University, to work on klystron tube research.

M. Hillier

In 1945 he moved to Baldock, Eng., as one of the original members of the present Services Electronics Research Laboratory. There he commenced development work on gas discharge devices and studied for the telecommunications examination of the City and Guilds Institute. During this period he also took an interest in shock and vibration phenomena in electronic devices and acted as secretary to a tube development committee dealing with this subject.

Recently Mr. Hillier has been concerned with the development and preproduction engineering of millimeter-wave klystrons.

Nick Holonyak, Jr., (S'51–A'55) was born in Zeigler, Ill. on November 3, 1928. He received the B.S. in E.E. degree in 1950, the M.S. degree in 1951, and the Ph.D. degree in 1954 from the University of Illinois. While a graduate student at the University of Illinois, he was a teaching assistant, a research assistant in microwave tubes, a research assistant in semiconductors and transistors, and held the Texas Instruments Fellowship in transistor physics. He joined the transistor development department of Bell Telephone Laboratories in 1954 and was inducted into the Army in October, 1955. He is a private in the Army, currently serving in the Far East.

N. Holonyak, Jr.

He is a member of the American Physical Society, American Association for the Advancement of Science, Sigma Xi, Eta Kappa Nu, Tau Beta Pi, Pi Mu Epsilon, and Phi Kappa Phi.

❖

Mohamed A. W. Ismail was born in Mansoura, Egypt on April 27, 1927. He received the Bachelor's degree in electrical engineering from Fouad University at Giza in 1949. For one year thereafter he was an instructor in the Electrical Engineering Department of the same university. In 1951, he joined the High Frequency Institute of the Swiss Federal Institute of Technology in Zurich, Switzerland where he worked on fm radar systems and received the degree of D.Sc.Techn. in 1954. He worked at the Hasler Company in Bern designing and developing a new system of fm radar during 1954–1955. Since September, 1955, he has been a lecturer in Radio Engineering at Ein-Shams University in Cairo, Egypt.

M. A. W. Ismail

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For a photograph and biography of John L. Moll, see page 1978 of the December, 1955 issue of Proceedings of the IRE.

❖

Robert C. Moore (S'32–A'34–VA'39–SM'51–F'55) was born January 3, 1912 in Pontiac, Ill. He received the B.S.E.E. from the University of North Dakota in 1933, the M.S.E.E. from M.I.T. in 1934, and the B.A. from Oxford University in 1936. He has been with Philco Corporation since 1937, principally in television development and research. At present he is a section engineer in the Advanced Development Laboratory of the Television Division. He is a supervisor in the Apple system project, and has helped develop deflection circuit and Apple tube screen geometry specifications.

R. C. Moore

Mr. Moore holds five Philco patents in radar, fourteen in television, and three in the electronics field. He is a member of the Franklin Institute.

❖

Stephen W. Moulton (S'47–A'48) was born March 16, 1925 in Boston, Mass. He received the B.S.E.E. degree from M.I.T. and the M.S.E.E. degree from the same institution in 1947. Since joining Philco Corporation in 1947, Mr. Moulton has engaged in uhf and vhf research, pioneering in the circuit theory of uhf tuners. At present he is engaged in color television development and research, and is one of the inventors of the Apple system having contributed especially to integration of components, work on the system as a whole, and design of those features of the tube which are closely related to circuit performance.

S. W. Moulton

Mr. Moulton holds two Philco patents in radar and one in television.

❖

Stuart L. Parsons was born September 6, 1912 in Gables, Mich. He received the B.S. from the University of Michigan and the M.S. in physics in 1938 from the same institution. His industrial experience has included studies in mechanism, spectroscopy, physical research, and process development in various electronic companies. A major part of this time has been spent in the design and development of equipment for production of electronic components, namely, cathode ray tubes, receiving tubes, resistors, and semiconductors.

S. L. Parsons

He joined Philco Corporation in 1952, and at present is Chief Equipment Design Engineer for both the research and production engineering groups in the Lansdale Tube Division.

Mr. Parsons is a member of Sigma Phi.

Melvin E. Partin (S'49–A'49) was born February 18, 1925 in Kissimmee, Fla. He received the B.S.E.E. in 1949 from the University of Florida, and he joined Philco Corporation the same year. Since that time he has engaged in circuitry and color television research. One of the inventors of the Apple system, he has made several contributions in integration of components. Mr. Partin holds one Philco patent in television. He is a member of Sigma Tau and Phi Kappa Phi.

M. E. PARTIN

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George W. Pratt (A'43–M'51) was born January 1, 1918 in New York, N. Y. He received the Bachelor's degree in chemical engineering from Cooper University in 1941. From 1942 to 1948 he was employed by RCA where he was in charge of cathode ray tube production engineering. He joined Philco Corporation in 1949 as head of cathode ray tube design in the tube development laboratory. Mr. Pratt made significant contributions to the production of the Apple system.

G. W. PRATT

He is a member of Sigma Xi.

❖

Haraden Pratt, (A'14–M'17–F'29) Secretary and Past President of the IRE, was born in San Francisco, Calif., on July 18, 1891. He began his radio career as an amateur in 1906 and from 1910 to 1914 was a wireless telegraph operator and installer of equipment for the United Wireless Telegraph Company and Marconi Wireless Telegraph Company of America.

H. PRATT

In 1914 he received the B.S. degree in electrical engineering from the University of California and thereafter became a construction and operating engineer for the Marconi Company's 300-kilowatt spark-type trans-Pacific radio stations in California.

As an Expert Radio Aide for the Navy Department from 1915 to 1920, he was concerned with the construction and maintenance of its high-powered radio stations. In 1920, he began the establishment of the public service radiotelegraph system of the Federal Telegraph Company on the West Coast. In 1925 he constructed and operated a radiotelegraph system between Salt Lake City and Los Angeles for the Western Air Express. Two years later he was in charge of development work on radio aids for air navigation at the National Bureau of Standards. In 1928 he became Chief Engineer, and later Vice-President, of Mackay Radio and Telegraph Company. He constructed its world-wide communication system.

For his work during World War II as Chief of the National Defense Research Committee's Division 13 on Communications, Mr. Pratt was awarded a Presidential Certificate of Merit. Immediately after the war, he became Vice-President and Chief Engineer of the Commercial Cable Company, All America Cables and Radio, Inc., and the American Cable and Radio Corporation. For many years he held offices in other companies of the International Telephone and Telegraph Corporation, but retired from these activities in 1951. In October of that year, he received a Presidential appointment to the newly-created post of Telecommunications Advisor to the President. Mr. Pratt has since retired from government service.

He is a member of Sigma Xi and the Veteran Wireless Operators Association, a Fellow of the American Institute of Electrical Engineers and the Radio Club of America and an Associate Fellow of the Institute of the Aeronautical Sciences. In 1944, he received the IRE's Medal of Honor.

❖

Donald Richman (S'42–A'45–SM'52) received the degree of B.E.E. from the College of the City of New York in 1943. He has attended evening classes at Polytechnic Institute of Brooklyn since that time, receiving the M.E.E. degree in 1948 and continuing thereafter with doctoral studies.

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Since 1943 he has been engaged in a number of research and development activities at Hazeltine Corporation where he is now Consulting Engineer for the Research and Licensee Divisions.

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Meier Sadowsky was born May 16, 1915 in San Antonio, Tex. He received the B.S. in 1936 and the M.S. in 1939, both from the College of the City of New York. After teaching chemistry and physics at Essex Junior College, he joined RCA in 1940 as a development and research engineer. In 1949 he was employed by Philco Corporation and at present is Executive Engineer in charge of the Chemical Laboratories, and Chief Chemical Engineer of Philco's Lansdale Tube Division. He holds three Philco patents in cathode ray tubes.

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PETER SWERLING

Dr. Swerling worked on Project Rand at Douglas Aircraft Co. in 1947 and 1948. From September, 1949 to January, 1950 he was a teaching assistant in Mathematics at U.C.L.A. Since 1949 he has been employed by Rand Corporation, working full-time there since 1952.

Theory of random noise, especially as applied to radar performance, has been Dr. Swerling's major field of research.

He is a member of Phi Beta Kappa, Sigma Xi, Pi Mu Epsilon, American Mathematical Society, and the Society for Industrial and Applied Mathematics.

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Morris Tanenbaum was born in Huntington, W. Va., on November 10, 1928. He received the A.B. degree in chemistry from the Johns Hopkins University in 1949 and the M.A. and Ph.D. degrees in chemistry from Princeton University in 1950 and 1952 respectively.

M. TANENBAUM

Since 1952 Dr. Tanenbaum has been with the Bell Telephone Laboratories where he has studied the chemical and physical properties of semiconductors and has also been engaged in exploratory research on semiconductor devices.

Dr. Tanenbaum is a member of the American Chemical Society, the American Physical Society, Phi Lambda Upsilon, Phi Beta Kappa, and Sigma Xi.

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Arthur Uhlir, Jr., (A'53) was born in Chicago, Ill. on February 2, 1926. He received the B.S. and M.S. degrees in chemical engineering from Illinois Institute of Technology in 1945 and 1948, and the S.M. and Ph.D. degrees in physics in 1950 and 1952 from the University of Chicago where he was an AEC pre-doctoral fellow from 1949 to 1951. He was a process analyst at Douglas Aircraft Company in Chicago during 1945, and from 1945 to 1948 did fluid mechanics research at Armour Research Foundation. Since 1951 he has been in the Transistor Development Department of Bell Telephone Laboratories.

A. UHLIR, JR.

There he has worked on point-contact transistor theory, semiconductor surface protection, and electrochemical properties of semiconductors, and has developed an electrolytic micromachining technique for metals and semiconductors. He is now working on microwave semiconductor devices.

Dr. Uhlir is a member of the American Physical Society, Sigma Xi, Phi Lambda Upsilon, and the American Association for the Advancement of Science.

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Armen H. Zemanian (A'51) was born on April 16, 1925, in Bridgewater, Mass. He received the B.E.E. degree from the College of the City of New York in 1947 and the M.E.E. and Eng.Sc.D. degrees from New York University in 1949 and 1953, respectively. He taught in the electrical engineering department of C.C.N.Y. for the academic year 1947–48, and then joined the Maintenance Company, New York, N. Y. Since 1952 he has been teaching electrical engineering at New York University, where he is an assistant professor.

A. H. ZEMANIAN

Dr. Zemanian is a member of the National Society of Professional Engineers, the American Institute of Electrical Engineers, Tau Beta Pi, Eta Kappa Nu, and Sigma Phi Omega.

IRE News and Radio Notes

Kansas City Plans Annual Conference November 8–9

The Kansas City Section of the IRE will hold its annual technical conference at the Town House Hotel, Kansas City, Kansas, November 8–9, 1956. Technical sessions have been scheduled on topics of management, industrial electronics, transistors, magnetic amplifiers, components, and systems.

The annual banquet on the evening of November 8 will feature an address on earth satellites by S. F. Singer, Univ. of Md.

The general committee for the conference, headed by C. V. Miller, Chairman, consists of Allan Shontz, Vice-Chairman and Registrations; A. C. Cotts, Technical Program; C. O. Files, Advertising and Exhibits; Ted Anderson, Facilities; M. R. Jones, Hospitality; F. K. Hyer, Publications; V. M. Mathews, Jr., Publicity; W. H. Ashley, Jr., Treasurer.

The Deadline is November 2 for Papers

IRE National Convention, March 18–21, 1957

Prospective authors are requested to submit all of the following information:
(1) 100-word abstract in triplicate with title, name and address of author.
(2) 500-word summary in triplicate with title, name and address of author.
(3) Indicate the technical field in which your paper falls:

- Aeronautical & Navigational Electronics
- Antennas & Propagation
- Audio
- Automatic Control
- Broadcast & Television Receivers
- Broadcast Transmission Systems
- Circuit Theory
- Communications Systems
- Component Parts
- Electron Devices
- Electronic Computers
- Engineering Management
- Industrial Electronics
- Information Theory
- Instrumentation
- Medical Electronics
- Microwave Theory & Techniques
- Military Electronics
- Nuclear Science
- Production Techniques
- Reliability and Quality Control
- Telemetry & Remote Control
- Ultrasonics Engineering
- Vehicular Communications

Deadline for acceptance of papers: November 2, 1956

Address all material to: Ben Warriner
1957 Technical Program Committee
Institute of Radio Engineers, Inc.
1 East 79 Street, New York 21, N. Y.

Everitt Heads American Engineering Education Group

W. L. Everitt, former president of the IRE and dean of the College of Engineering, University of Illinois, was recently elected president of the American Society for Engineering Education.

W. L. EVERITT

A graduate of Cornell University, Dean Everitt holds advanced degrees from the University of Michigan and Ohio State University. He joined the electrical engineering faculty at Ohio State in 1926 and left there in 1944 to become head of the Department of Electrical Engineering at the University of Illinois. He became dean of the Illinois College of Engineering and director of its Engineering Experiment Station in 1949.

Dean Everitt's professional activities have included service on various committees and panels of the Department of Defense, the National Science Foundation, the Fulbright Award Committee of the National Academy of Science, the National Bureau of Standards, and Engineers' Council for Professional Development.

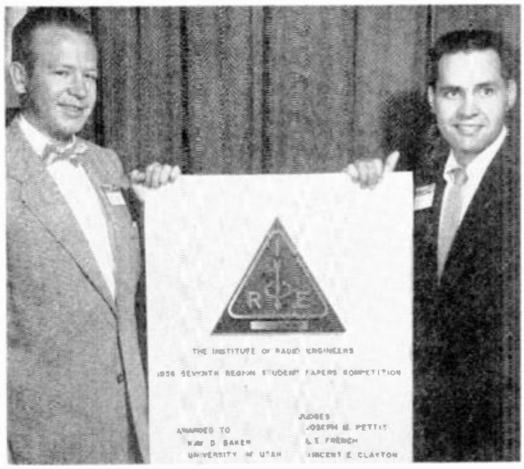
He was 1945 IRE President and 1954 Medal of Honor winner. He is also a fellow of the AIEE. An IRE Fellow, Dr. Everitt is the author and editor of several texts on electrical engineering and consultant to several broadcast stations and radio manufacturing companies.

ACTIVITIES OF IRE SECTIONS AND PROFESSIONAL GROUPS

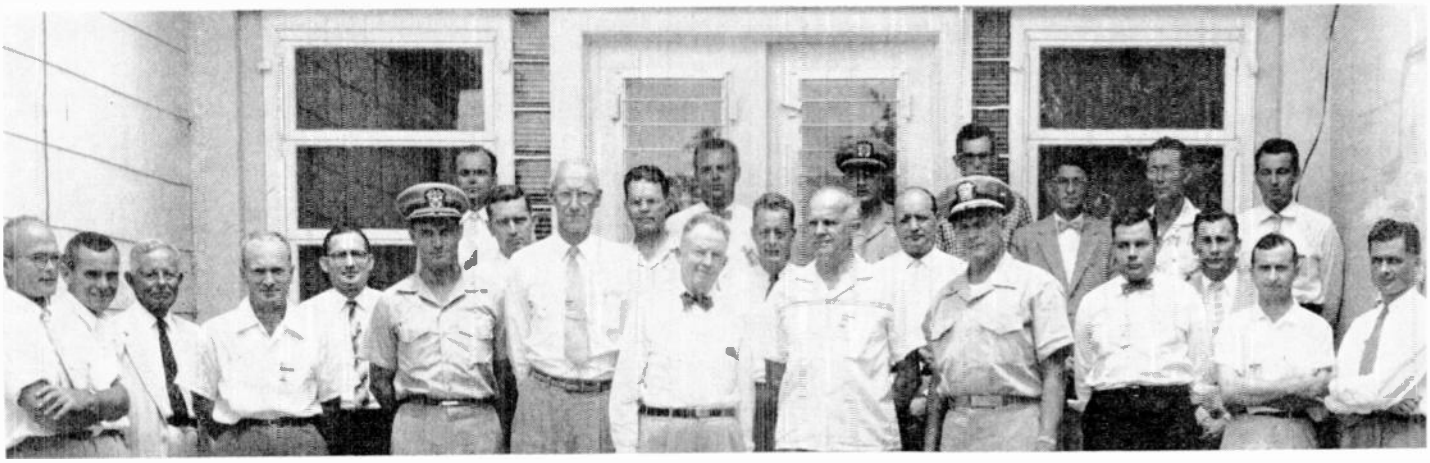
R. I. Cole (*left*) assumes the chairmanship of the Washington, D. C. Section as Henry Metz (*right*) congratulates him. Other officers elected were A. H. Schooley, Vice-Chairman, R. M. Page, Treasurer, and John Durkovic, Secretary. At this meeting J. B. Oakes of the Applied Physics Laboratory, Johns Hopkins University, was awarded the PGAU annual papers award.

Recently-elected 1956 officers of the Dallas Section posed for the camera at the annual dinner meeting. Left to right—J. A. Albano, Secretary-Treasurer, C. F. Seay, Vice-Chairman, and G. K. Teal, Chairman.

Pictured (*left to right*) are members of the planning committee for the Industrial Electronics Symposium: C. F. Schunemann, Publicity; Reuben Kazarian, Co-Chairman; B. M. Jones, Co-Chairman; F. A. Furfari, Local Arrangements; R. C. Rodgers, Technical Program; and H. C. Martin, Finance.



Warren Shelton (*left*), University of Nevada, runner-up, and K. D. Baker (*right*), University of Utah, winner, pose with a plaque presented to Mr. Baker in a student papers competition. The competition was inaugurated at the recent Seventh Regional Technical Conference at Salt Lake City.



On a recent visit to the Northwest Florida Section, 1956 IRE President A. V. Loughren toured the electronic instrumentation facilities of the U. S. Navy Mine Defense Laboratory at Panama City, Florida. Shown are some of the laboratory's personnel with Mr. Loughren. Front row, left to right—Commander P. B. Smith, U. S. Navy Services Department; A. L. Bennett, Associate Superintending Scientist; A. V. Loughren; F. A. Rohrman, Superintending Scientist; and Captain J. C. Myers, U. S. Navy Commanding Officer. Back row, left to right—C. J. Barry, F. J. Murphree, G. C. de Coutouly, L. F. Jones, C. B. Koesy, R. C. Lowry, R. C. Aucremann, R. R. Herold, G. Walker, M. H. Naeseth, C. A. Good, C. A. Haulman, R. L. Nasoni, H. H. Penton, P. K. White, G. P. Mathis, S. B. Marley, R. T. Galloway, B. H. Lloyd, and R. H. Wheeler.

Calendar of Coming Events

Symposium on Information Theory, Cambridge, Mass., Sept. 10–12
Second RETMA Conference on Reliable Electrical Connections, U. of Pa., Philadelphia, Pa., Sept. 11–12
PGBTS Sixth Annual Fall Symposium, Pittsburgh, Pa., Sept. 14–15
Conference on Communications, Roosevelt Hotel, Cedar Rapids, Iowa, Sept. 14–15
Transistor Reliability Symposium, New York City, Sept. 17–18
Instrument-Automation Conference & Exhibit, Coliseum, New York City, Sept. 17–21
Symposium on Radio-Wave Propagation, Paris, France, Sept. 17–22
PGNS Third Annual Meeting, Mellon Institute Auditorium, Pittsburgh, Pa., Sept. 20–22
Industrial Electronics Symposium, Manger Hotel, Cleveland, Ohio, Sept. 24–25
National Electronics Conference, Hotel Sherman, Chicago, Ill., Oct. 1–3
Canadian IRE Convention & Exposition, Automotive Bldg., Exhibition Park, Toronto, Can., Oct. 1–3
Second Annual Symposium on Aeronautical Communications, Hotel Utica, Utica, N. Y., Oct. 8–9
Computer Applications Symposium, Morrison Hotel, Chicago, Ill., Oct. 9–10
URSI Fall Meeting, Univ. of Calif., Berkeley, Calif., Oct. 11–12
IRE-RETMA Radio Fall Meeting, Hotel Syracuse, Syracuse, N. Y., Oct. 15–17
Conference on Magnetism & Magnetic Materials, Hotel Statler, Boston, Mass., Oct. 16–18
PGED Annual Technical Meeting, Shoreham Hotel, Washington, D. C., Oct. 25–26
East Coast Conference on Aeronautical & Navigational Electronics, Fifth Regiment Armory, Baltimore, Md., Oct. 29–30
Convention on Ferrites, Institute of Electrical Engineers, London, England, Oct. 29–Nov. 2
Conference on Electrical Techniques in Medicine and Biology, Governor Clinton Hotel, N. Y., Nov. 7–9
Kansas City IRE Technical Conference, Town House Hotel, Kansas City, Kan., Nov. 8–9
Symposium on Applications of Optical Principles to Microwaves, Washington, D. C., Nov. 14–16
New England Radio Engineering Meeting, Bradford Hotel, Boston, Mass., Nov. 15–16
PGVC Eighth National Meeting, Fort Shelby Hotel, Detroit, Mich., Nov. 29–30
Second Instrumentation Conference & Exhibit, Biltmore Hotel, Atlanta, Ga., Dec. 5–7
IRE-AIEE-ACM Eastern Joint Computer Conference, Hotel New Yorker, New York City, Dec. 10–12
Symposium on Reliability & Quality Control in Elec., Statler Hotel, Wash., D. C., Jan. 14–15, 1957
Symposium on Propagation and Radiation of VLF Electromagnetic Waves, Boulder Labs., Boulder, Colo., Jan. 23–25

TRANSACTIONS OF THE IRE PROFESSIONAL GROUPS

The following issues of TRANSACTIONS are available from The Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, New York, at the prices listed below:

Sponsoring Group	Publications	Group Members	IRE Members	Non-Members*
Aeronautical and Navigational Electronics	PGAE-5: A Dynamic Aircraft Simulator for Study of Human Response Characteristics (6 pages)	$ .30	$ .45	$ .90
	PGAE-6: Ground-to-Air Cochannel Interference at 2900 MC (10 pages)	.30	.45	.90
	PGAE-8: June 1953 (23 pages)	.65	.95	1.95
	PGAE-9: September 1953 (27 pages)	.70	1.05	2.10
	Vol. ANE-1, No. 2, June 1954 (22 pages)	.95	1.40	2.85
	Vol. ANE-1, No. 3, September 1954 (27 pages)	1.00	1.50	3.00
	Vol. ANE-1, No. 4, December 1954 (27 pages)	1.00	1.50	3.00
	Vol. ANE-2, No. 1, March 1955 (41 pages)	1.40	2.10	4.20
	Vol. ANE-2, No. 2, June 1955 (49 pages)	1.55	2.30	4.65
	Vol. ANE-2, No. 3, September 1955 (27 pages)	.95	1.45	2.85
	Vol. ANE-2, No. 4, December 1955 (47 pages)	1.40	2.10	4.20
	Vol. ANE-3, No. 1, March 1956 (42 pages)	1.30	1.95	3.90
	Vol. ANE-3, No. 2, June 1956 (54 pages)	1.40	2.10	4.20
Antennas and Propagation	PGAP-4; IRE Western Convention, August 1952 (136 pages)	2.20	3.30	6.60
	Vol. AP-1, No. 1, July 1953 (30 pages)	1.20	1.80	3.60
	Vol. AP-1, No. 2, October 1953 (31 pages)	1.20	1.80	3.60
	Vol. AP-2, No. 1, January 1954 (39 pages)	1.35	2.00	4.05
	Vol. AP-2, No. 2, April 1954 (41 pages)	2.00	3.00	6.00
	Vol. AP-2, No. 3, July 1954 (36 pages)	1.50	2.25	4.50
	Vol. AP-3, No. 4, October 1954 (36 pages)	1.50	2.25	4.50
	Vol. AP-3, No. 1, January 1955 (43 pages)	1.60	2.40	4.80
	Vol. AP-3, No. 2, April 1955 (47 pages)	1.60	2.40	4.80
	Vol. AP-3, No. 3, July 1955 (66 pages)	2.05	3.10	6.15
	Vol. AP-4, No. 1, January 1956 (100 pages)	2.65	3.95	7.95
	Vol. AP-4, No. 2, April 1956 (83 pages)	2.20	3.30	6.60
Audio	PGA-7: Editorials, Technical Papers & News, May 1952 (47 pages)	.90	1.35	2.70
	PGA-10: November–December 1952 (27 pages)	.70	1.05	2.10
	Vol. AU-1, No. 2, March–April 1953 (34 pages)	.80	1.20	2.40
	Vol. AU-1, No. 5, September–October 1953 (11 pages)	.50	.75	1.50
	Vol. AU-1, No. 6, November–December 1953 (27 pages)	.90	1.35	2.70
	Vol. AU-2, No. 1, January–February 1954 (38 pages)	1.20	1.80	3.60
	Vol. AU-2, No. 2, March–April 1954 (31 pages)	.95	1.40	2.85
	Vol. AU-2, No. 3, May–June 1954 (27 pages)	.95	1.40	2.85
	Vol. AU-2, No. 4, July–August 1954 (27 pages)	.95	1.40	2.85
	Vol. AU-2, No. 5, September–October 1954 (22 pages)	.95	1.40	2.85
	Vol. AU-2, No. 6, November–December 1954 (24 pages)	.80	1.20	2.40
	Vol. AU-3, No. 1, January–February 1955 (20 pages)	.60	.90	1.80
	Vol. AU-3, No. 2, March–April 1955 (32 pages)	.95	1.40	2.85
	Vol. AU-3, No. 3, May–June 1955 (30 pages)	.85	1.25	2.55
	Vol. AU-3, No. 4, July–August 1955 (46 pages)	1.15	1.75	3.45
	Vol. AU-3, No. 5, September–October 1955 (33 pages)	.90	1.35	2.70
	Vol. AU-3, No. 6, November–December 1955 (36 pages)	.95	1.40	2.85
	Vol. AU-4, No. 1, January–February 1956 (27 pages)	.75	1.10	2.25
	Vol. AU-4, No. 2, March–April (17 pages)	.55	.80	1.65
	Vol. AU-4, No. 3, May–June 1956 (34 pages)	.80	1.20	2.40
Automatic Control	PGAC-1: May 1956 (97 pages)	1.95	2.90	5.85
Broadcast Transmission Systems	PGBTS-2: December 1955 (54 pages)	1.20	1.80	3.60
	PGBTS-4: March 1956 (21 pages)	.75	1.10	2.25
Broadcast and Television Receivers	PGBTR-1: Round Table Discussion on UHF TV Receiver Considerations, 1952 IRE National Convention (12 pages)	.50	.75	1.50
	PGBTR-5: January 1954 (96 pages)	1.80	2.70	5.40
	PGBTR-7: July 1954 (58 pages)	1.15	1.70	3.45
	PGBTR-8: October 1954 (20 pages)	.90	1.35	2.70

* Public libraries, colleges and subscription agencies may purchase at IRE member rate.

(Continued on page 1199)

TRANSACTIONS OF THE IRE PROFESSIONAL GROUPS (Continued)

Sponsoring Group	Publications	Group Members	IRE Members	Non-Members*
	Vol. BTR-1, No. 1, January 1955—Papers Presented at the Radio Fall Meeting, 1954 (68 pages)	$1.25	$1.85	$3.75
	Vol. BTR-1, No. 2, April 1955 (40 pages)	.95	1.45	2.85
	Vol. BTR-1, No. 3, July 1955 (51 pages)	.95	1.45	2.85
	Vol. BTR-1, No. 4, October 1955 (19 pages)	.95	1.40	2.85
Circuit Theory	Vol. CT-1, No. 4, December 1954 (42 pages)	1.00	1.50	3.00
	Vol. CT-2, No. 1, March 1955 (106 pages)	2.70	4.05	8.10
	Vol. CT-2, No. 3, September 1955 (62 pages)	1.40	2.10	4.20
	Vol. CT-2, No. 4, December 1955 (88 pages)	1.85	2.75	5.55
Communications Systems	Vol. CS-2, No. 1, January 1954 (83 pages)	1.65	2.50	4.95
	Vol. CS-2, No. 2, July 1954 (132 page—)	2.25	3.35	6.75
	Vol. CS-2, No. 3, November 1954—IRE Symposium on Global Communications, June 23-25, 1954, Washington, D. C. and IRE-AIEE Symposium on Military Communications, April 28, 1954, New York, N. Y. (181 pages)	3.00	4.50	9.00
	Vol. CS-3, No. 1, March 1955—Papers Presented at the Symposium on Marine Communications & Navigation, October 13-15, 1954, Boston, Mass. (72 pages)	1.00	1.50	3.00
	Vol. CS-4, No. 1, March 1956—Symposium on Communications by Scatter Techniques, November 14-15, 1955, Washington D. C. (122 pages)	2.15	3.20	6.45
	Vol. CS-4, No. 2, May, 1956—Symposium on Aeronautical Communications, November 21-22, 1955, Utica, New York (182 pages)	2.90	4.35	8.70
Component Parts	PGCP-1: March 1954 (46 pages)	1.20	1.80	3.60
	PGCP-2: September 1954—Papers Presented at the Component Parts Sessions at the 1954 Western Electronic Show & Convention, Los Angeles, Calif. (119 pages)	2.25	3.35	6.75
	PGCP-3: April 1955 (44 pages)	1.00	1.50	3.00
	PGCP-4: November 1955 (92 pages)	2.00	3.00	6.00
	Vol. CP-3, No. 1, March 1956 (35 pages)	1.70	2.55	5.10
Electronic Computers	Vol. EC-3, No. 3, September 1954 (54 pages)	1.80	2.70	5.40
	Vol. EC-4, No. 2, June 1955 (36 pages)	.90	1.35	2.70
	Vol. EC-4, No. 3, September 1955 (45 pages)	1.00	1.50	3.00
	Vol. EC-4, No. 4, December 1955 (40 pages)	.90	1.35	2.70
	Vol. EC-5, No. 2, June 1956 (46 pages)	.90	1.35	2.70
Electron Devices	PGED-4: December 1953 (62 pages)	1.30	1.95	3.90
	Vol. ED-1, No. 2, April 1954 (75 pages)	1.40	2.10	4.20
	Vol. ED-1, No. 3, August 1954 (77 pages)	1.40	2.10	4.20
	Vol. ED-1, No. 4, December 1954 (280 pages)	3.20	4.80	9.60
	Vol. ED-2, No. 2, April 1955 (53 pages)	2.10	3.15	6.30
	Vol. ED-2, No. 3, July 1955 (27 pages)	1.10	1.65	3.30
	Vol. ED-2, No. 4, October 1955 (42 pages)	1.50	2.25	4.50
	Vol. ED-3, No. 1, January 1956 (74 pages)	2.10	3.15	6.30
	Vol. ED-3, No. 2, April 1956 (40 pages)	1.10	1.65	3.30
Engineering Management	PGEM-1: February 1954 (55 pages)	1.15	1.70	3.45
	Vol. EM-3, No. 1, January 1956 (29 pages)	.95	1.40	2.85
	Vol. EM-3, No. 2, April 1956 (15 pages)	.55	.80	1.65
Industrial Electronics	PGIE-1: August 1953 (40 pages)	1.00	1.50	3.00
	PGIE-2: March 1955 (81 pages)	1.90	2.85	5.70
	PGIE-3: March 1956 (110 pages)	1.70	2.55	5.10
Information Theory	PGIT-3: March 1954 (159 pages)	2.60	3.90	7.80
	PGIT-4: September 1954 (234 pages)	3.35	5.00	10.00
	Vol. IT-1, No. 1, March 1955 (76 pages)	2.40	3.60	7.20
	Vol. IT-1, No. 2, September 1955 (50 pages)	1.90	2.85	5.70
	Vol. IT-1, No. 3, December 1955 (44 pages)	1.55	2.30	4.65
	Vol. IT-2, No. 1, March 1956 (45 pages)	1.60	2.40	4.80
Instrumentation	PGI-3: April 1954 (55 pages)	1.05	1.55	3.15
	PGI-4: October 1955 (182 pages)	2.70	4.05	8.10
	PGI-5: June 1956 (224 pages)	3.20	4.80	9.60
Medical Electronics	PGME-2: October 1955 (39 pages)	.85	1.25	2.55
	PGME-3: November 1955 (55 pages)	1.10	1.65	3.30
	PGME-4: February 1956 (51 pages)	1.95	2.90	5.85

* Public libraries, colleges and subscription agencies may purchase at IRE member rate.

(Continued on page 1200)

Tentative Program Is Set for Oct. Magnetism Conference

The second Conference and Exhibit on Magnetism and Magnetic Materials will be held at the Hotel Statler, Boston, Mass., October 16-18, 1956. The conference is sponsored by the American Institute of Electrical Engineers in cooperation with the American Physical Society, the American Institute of Mining, Metallurgical and Petroleum Engineers, and the IRE. In addition to the program of technical papers, the meeting will feature exhibits by manufacturers of magnetic materials and associated equipment. A dinner will be held on October 17 at the nearby Museum of Science. R. M. Bozorth will speak about his trip to Russia where he attended a recent conference on magnetism.

Eight sessions will include approximately 75 technical papers on magnetic anisotropy, permanent magnets and fine particles, magnetism and physical metallurgy, losses in soft magnetic materials, ferrites, high frequency phenomena, switching devices and magnetic amplifiers, and apparatus and design.

The tentative program includes the following speakers and their papers: J. H. Van Vleck, *A Survey of the Theory of Magnetic Anisotropy*; R. M. Bozorth, *Magnetic Annealing;* T. O. Paine, *Fine Particle Magnets;* C. Zener, *Magnetism and the Constitution of Metals;* J. Goodenough, *The Origin of Losses in Magnetic Materials;* N. Bloemberger, *Fundamentals of Ferromagnetic Resonance;* C. L. Hogan, *Microwave Applications of Magnetic Materials;* R. L. Conger, *High Frequency Effects in Magnetic Films;* T. Bonn, *The Ferractor.*

The complete program will be published in the October issue of *Electrical Engineering.* Abstracts of the technical papers will be available before the meeting.

The Conference Chairman is R. M. Bozorth, Bell Telephone Laboratories. Other conference officers are: Local Conference Chairman, T. O. Paine, General Electric Company; Program Chairman, C. P. Bean, General Electric Research Laboratory; Manager of Exhibits, R. Rimbach, Richard Rimbach Associates, Inc.

URSI Meets October 11-12

The URSI Fall Meeting is scheduled for October 11-12, 1956, at the University of California, Berkeley, Calif. A combined technical session of interest to all participants is scheduled for the morning of October 11. It will be followed by one or more sessions in each of the following fields: Commission 2—Radio and Troposphere (J. B. Smyth, Smyth Research Associates, Chairman); Commission 3—Ionospheric Radio (M. G. Morgan, Dartmouth College, Chairman); and Commission 4—Radio Noise of Terrestrial Origin (A. W. Sullivan, University of Florida, Chairman).

For further information, write to J. P. Hagen, Secretary of the U. S. National Committee of URSI, Code 4100 Naval Research Laboratory, Washington 25, D. C.

The Twelfth General Assembly of the URSI will be held at Boulder, Colo., Aug. 22-Sept. 5, 1957.

Ionospheric Research Group Wins Commerce Dept. Award

A research group consisting of D. K. Bailey, Ross Bateman (A'42), Falls Church, Va., and R. C. Kirby (A'44–SM'54), Chief of the Ionospheric Research Section of the Boulder Laboratories, National Bureau of Standards has been awarded the Department of Commerce Gold Medal for exceptional service. The award was made "for major contributions to the advancement of the science of radio wave propagation and long distance radio communications during the extensive elucidation of the defining features of a new kind of propagation."

The research group took part in the discovery of ionospheric forward scatter and studied the physics of the phenomena involved. In addition, they directed the application of scientific studies to construction of practical communication circuits, which has resulted in supervising Air Force contracts for installation of several such circuits in the Arctic.

A comprehensive report of their work was published in the October, 1955 issue of the Proceedings of the IRE under the title *Radio Transmission at VHF by Scattering and Other Processes in the Lower Ionosphere.*

Optics-Microwave Symposium Scheduled for Nov. 14–16

A Symposium on Optics and Microwaves will be held at Lisner Auditorium, George Washington University, Washington, D. C., on November 14–16, 1956. The meeting is jointly sponsored by the IRE Professional Group on Antennas and Propagation, the George Washington University, and the Optical Society of America. The technical program will consist of six sessions, each embracing a subject of general interest to all persons who deal with optical phenomena in research or application in the fields of engineering, medicine or the related physical sciences. Survey and tutorial type papers will be presented to encourage understanding of the basic physics underlying fundamental characteristics which relate optics and microwaves as the two concepts exist.

It is the purpose of the symposium to promote interest in the primary common problems associated with optics and microwaves, and to demonstrate that these lie within the scope of modern theoretical and practical optics which relate such diverse fields as human vision and radio astronomy.

Advance registration for the meeting is $2.50 and may be made by mailing a check for that amount to "Symposium on Optics and Microwaves," P.O. Box 355 Falls Church, Va. Registration is $3.50 at the door.

TRANSACTIONS OF THE IRE PROFESSIONAL GROUPS (Continued)

Sponsoring Group	Publications	Group Members	IRE Members	Non-Members*
Microwave Theory and Techniques	Vol. MTT-1, No. 2, November 1953 (44 pages)	$ .90	$1.35	$2.70
	Vol. MTT-2, No. 3, September 1954 (54 pages)	1.10	1.65	3.30
	Vol. MTT-3, No. 1, January 1955 (47 pages)	1.50	2.25	4.50
	Vol. MTT-3, No. 4, July 1955 (54 pages)	1.60	2.40	4.80
	Vol. MTT-3, No. 5, October 1955 (59 pages)	1.70	2.55	5.10
	Vol. MTT-3, No. 6, December 1955 (64 pages)	1.75	2.60	5.25
	Vol. MTT-4, No. 1, January 1956 (63 pages)	1.65	2.45	4.95
	Vol. MTT-4, No. 2, April 1956 (69 pages)	1.70	2.55	5.10
Nuclear Science	Vol. NS-1, No. 1, September 1954 (42 pages)	.70	1.00	2.00
	Vol. NS-2, No. 1, June 1955 (15 pages)	.55	.85	1.65
	Vol. NS-3, No. 1, February 1956 (40 pages)	.90	1.35	2.70
	Vol. NS-3, No. 2, March 1956 (31 pages)	1.40	2.10	4.20
	Vol. NS-3, No. 3, June 1956 (24 pages)	1.00	1.50	3.00
Reliability and Quality Control	PGQC-2: March 1953 (51 pages)	1.30	1.95	3.90
	PGQC-3: February 1954 (39 pages)	1.15	1.70	3.45
	PGQC-4: December 1954 (56 pages)	1.20	1.80	3.60
	PGRQC-5: April 1955 (56 pages)	1.15	1.75	3.45
	PGRQC-6: February 1956 (66 pages)	1.50	2.25	4.50
	PGRQC-7: April 1956 (52 pages)	1.10	1.65	3.30
Telemetry and Remote Control	PGRTRC-1: August 1954 (16 pages)	.85	1.25	2.55
	PGRTRC-2: November 1954 (24 pages)	.95	1.40	2.85
	Vol. TRC-1, No. 1, February 1955 (24 pages)	.95	1.40	2.85
	Vol. TRC-1, No. 2, May 1955 (24 pages)	.95	1.40	2.85
	Vol. TRC-1, No. 3, August 1955 (12 pages)	.70	1.05	2.10
	Vol. TRC-2, No. 1, March 1956 (22 pages)	1.00	1.50	3.00
Ultrasonics Engineering	PGUE-1: June 1954 (62 pages)	1.55	2.30	4.65
	PGUE-3: May 1955 (70 pages)	1.45	2.20	4.35
Vehicular Communications	PGVC-4: June 1954 (98 pages)	2.40	3.60	7.20
	PGVC-5: June 1955 (76 pages)	1.50	2.25	4.50

* Public libraries, colleges and subscription agencies may purchase at IRE member rate.

Professional Group News

PGAP Sponsors VLF Symposium

The Denver-Boulder Chapter of the Professional Group on Antennas and Propagation, and the National Bureau of Standards will co-sponsor a symposium on the theoretical and experimental results in the propagation and radiation of very-low-frequency electromagnetic waves (less than 100 kc). The symposium will be held at the Boulder Laboratories of the National Bureau of Standards, January 23–25, 1957.

Solicitation of papers has begun, and brief summaries should be sent to J. R. Wait, Chairman, Denver-Boulder PGAP Chapter, National Bureau of Standards, Boulder, Colorado.

Round table discussions will be a feature of the symposium. Participants will include Owen Storey, Ottawa, K. G. Budden, Cavendish Laboratory, R. A. Helliwell, Stanford University, J. M. Watts and W. Q. Crichlow, National Bureau of Standards, and M. Newman, Lightning and Transients Institute.

D. L. Arenberg Gets First Annual Papers Award of PGUE at Boston Section Meeting

The Professional Group on Ultrasonic Engineering has presented its first annual best paper award to D. L. Arenberg for his paper on ultrasonic delay lines, which was presented at the 1954 IRE National Convention. The presentation of the award was made by J. E. May, Jr., PGUE Secretary, at the recent meeting of the Boston Section.

J. F. Herrick of the Mayo Foundation was elected national chairman and C. M. Harris was chosen vice-chairman. Julius Bernstein will continue as treasurer. Lawrence Batchelder, J. E. May, Jr., and O. H. Schmitt were also elected to serve on the administrative committee.

PGVC Holds Seventh National Annual Conference at Detroit

The Professional Group on Vehicular Communication holds its seventh annual national conference at Hotel Fort Shelby, Detroit, Michigan, November 29–30, 1956. Included in the two-day program of technical papers will be exhibits of mobile radio communications equipment.

A. B. Buchanan of the Detroit Edison Company is Conference Chairman. Members of the steering committee include: E. C. Denstaedt, Vice-Chairman; R. C. Stinson, Secretary-Treasurer; W. J. Norris, Exhibits; T. P. Rykala, Program; Neal Jackson, Arrangements; W. B. Williams, Publicity; Zoltan Kato, Hospitality; H. A. Penhollow, Registration.

Obituary

George P. Dixon (SM'46), a former vice-president of the International Telephone and Telegraph Corporation, died recently.

G. P. Dixon

At his death, Mr. Dixon was executive vice-president of the Armed Forces Communications and Electronics Association, a nonprofit organization that seeks to cement relationships between the communications and electronics industries and the armed forces. Its headquarters are at Washington, D. C.

He was born in Worcester, Mass., graduated from Worcester Polytechnic Institute and in 1912 began his career as a student engineer with the Pacific Telephone and Telegraph Company in San Francisco. He saw Mexican border service with the National Guard in 1916.

In World War I he rose from lieutenant to captain in the Army Signal Corps and served overseas with the Ninety-first Division and the Services of Supplies.

After the war he was an engineer for the Western Electric Company in New York, a New York Telephone Company traffic supervisor and communications engineer for the National City Bank of New York and associated companies from 1929 to 1940.

In 1940, Mr. Dixon became a lieutenant colonel in the Signal Corps and later was a colonel in the Army Air Forces. After the United States' entry into World War II, he went overseas as a signal communications officer with the Eighth Air Force.

He rose to chief signal officer of the Eighth Air Force and then was director of communications of the United States Strategic Air Forces in Europe.

Mr. Dixon became a vice-president of the International Telephone and Telegraph Corporation in 1945. In 1946–48, he was regional vice-president in Brazil, and later his office was in New York until he retired in 1950.

Mr. Dixon was a founder of the Armed Forces Communications and Electronics Association, first called the Armed Forces Communications Association. Since 1953 he had been editor of *Signal*, the organization's journal, as well as executive vice-president.

He won the Silver Star, Legion of Merit and Bronze Star Medal for his wartime services. His foreign decorations include Order of the British Empire, Legion of Honor (France), Belgium Croix de Guerre (Gold Star), and French Croix de Guerre (two palms). He was an honorary member of the French National Academy.

Technical Committee Notes

The **Antennas and Waveguides** Committee met on June 13 at IRE Headquarters with Chairman Henry Jasik presiding. The entire meeting was devoted to the review of the preliminary draft of the Proposed Standards on Antennas and Waveguides: Methods of Measurement of Waveguide and Waveguide Components.

Chairman R. M. Showers presided at a meeting of the **Radio Frequency Interference** Committee on June 20 at IRE Headquarters. The chairman reported that the supplement to IRE Standards on Receivers: Methods of Measurement of Interference Output of Television Receivers in the Range of 300 to 10,000 KC, 1954 will be published in the August issue of the PROCEEDINGS. The committee reviewed the activities of the American Standards Association, the Radio Electronic Television Manufacturers' Association, and the International Electrotechnical Commission in regard to their interference work. The remainder of the meeting was devoted to discussion of plans for future activity.

The **Radio Transmitters** Committee met on June 15 at IRE Headquarters with Chairman H. Goldberg presiding. R. N. Harmon, Chairman of Subcommittee 15.6 on Television Broadcast Transmitters, reported that the subcommittee has just about completed work on the Proposed Standards on Television: Methods of Testing Television Broadcast Transmitters. The subcommittee hopes to submit the proposed standard for approval at the next meeting of the committee. The committee reviewed definitions for the following terms: *spurious transmitter output; spurious transmitter output, radiated; spurious transmitter output, conducted; inband spurious transmitter output, extraband spurious transmitter output.*

Chairman M. W. Baldwin presided at a meeting of the **Standards** Committee on June 14 at IRE Headquarters. The Proposed Standard on Solid State Devices: Methods of Testing Transistors was discussed, amended and unanimously approved.

The Proposed Standard on Electron Tubes: Cathode Ray Tube Definitions was referred back to the Electron Tubes Committee with a request that they give it further review.

The Proposed Standards on Electronic Computers: Definitions of Terms, 1956, was unanimously approved as an IRE Standard

Books

An Introduction to Stochastic Processes by M. S. Bartlett

Published (1955) by Cambridge University Press, 32 East 57 St., N. Y. 32, N. Y. 294 pages+xiv pages +5 page index+bibliography+glossary. 15 figs. 8¾×5¼. $6.50.

Stochastic processes may be defined as functions or series of events generated by an underlying mechanism controlled by the laws of probability. Such series are not hard to find in the field of communication engineering. Random noise and communication signals are among the most familiar. Professor Bartlett's book presents methods for representing, analyzing, and studying such processes. It is aimed primarily at the applied mathematician and statistician, although those electrical engineers engaged in research and analysis, especially involving systems, will appreciate its contents.

The book begins with a chapter on elementary probability and presents a classification of stochastic processes. Discrete random sequences, both of the "random walk" and Markov variety, are discussed in Chapter 2, while their continuous counterparts occupy Chapter 3. Applications of the previously presented methods are taken up in Chapter 4, including sections on queues (applicable to the interesting aircraft traffic problem at a crowded airport), population growth, and the theory of epidemics. Convergence, linear difference, and differential equations applied to stochastic processes are presented in Chapter 5. Stationary processes are described in Chapter 6 using the Wiener-Khintchine tools of generalized harmonic analysis. Applications to the Wiener prediction and Shannon communication theory fall into Chapter 7. The principles of statistical inference applied to stochastic processes, including use of the likelihood function in connection with estimation problems, are the subject of Chapter 8. The final chapter is devoted to correlation regression, and peri dogram analysis of time series with sections on sampling fluctuations and "goodness of fit" tests. The bibliography is well drawn, selective and up to date.

Professor Bartlett has contributed a scholarly work to a field of great current interest. The material is highly compacted, a not unexpected circumstance since the author endeavors "to survey the whole field" in about three hundred pages. Even so, the book contains a wealth of illustrative examples which help clarify the mathematical formalism. A case in point is the presentation of the Theory of Communication which occupies pages 208-220. It is concise and nicely phrased for the person trained in the relevant mathematics. The average engineer, however, should not expect to achieve an understanding of the physical concepts involved without considerable study of other material. This characteristic conciseness does not detract, however, from the general utility of the book as a reference and text. It provides an opportunity for the studious person to gain a wide familiarity with the field, including the most recent contributions, in a framework of applications and without delving too deeply into the rigorous mathematics. As far as this reviewer knows, the book is unique in these respects.

EDWARD E. DAVID, JR.
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Transistors Handbook by W. D. Bevitt

Published (1956) by Prentice-Hall, Inc. Publishers, 70 Fifth Ave., N. Y. 11, N. Y. 390 pages+7 index pages+10 appendix pages+xiv pages. Illus. 8¼×6. $9.00.

A description of the material covered by this book is found in the author's preface. "The first half of the book deals with fundamental concepts, the different types of transistors, their characteristics and their measurements, and the circuit properties and behavior of transistors. The last half of the work covers practical applications and circuits of transistors." A reading of the text leads one to believe that it is as yet *too early* to prepare a really complete transistor handbook. The most significant comment which might be made about *Transistors Handbook* is that it is written on too practical a level to be of real utility to the transistor apparatus engineer.

This is indeed a strange indictment against a new book, but transistor engineering practice has not as yet become sufficiently stabilized to support the load of a handbook written on a practical basis. For example, in the latter portion of the book liberal reference is made to sample circuits for oscillators, amplifiers, receivers and the like as representative of modern transistor engineering practice. Unfortunately, however, the average age of the source material in the bibliography is approximately three years. This gives rise to a description of a superheterodyne broadcast band receiver using eight point-contact transistors and grounded *base* amplifiers throughout. If today's designs for such receivers are any indication, this is surely not an example of modern design practice.

Adherence to the policy of practicality throughout the text has led to an emphasis on what has been done with transistors rather than how to use them. The apparatus engineer who is confronted with a new application problem (or a relatively complicated old one) will receive limited assistance from the text.

Notwithstanding the foregoing, the book should be of great interest to a large portion of the author's intended audience. It will be helpful to "... experimenters, radio amateurs, radio and television servicemen, engineering students...." Meanwhile it would appear as though the transistor apparatus engineer must wait a while longer before all his source material and notes are gathered within a single volume.

R. P. BURR
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Cold Spring Harbor, N. Y.

Color Television Standards by D. G. Fink

Published (1955) by McGraw-Hill Book Company, 330 W. 42 St., N. Y. 36, N. Y. 491 pages+5 page index+21 page appendix+xii pages. 290 figures. 9¼×6¼. $8.50.

The current standards for monochrome and for color television broadcasting both were subjects of noteworthy engineering studies by National Television System Committees prior to their adoption by the FCC. Each of these television committees was composed of foremost technicians representing leading manufacturers of television equipment. Their deliberations included the most advanced thinking in the field of television broadcasting at the time. Consequently, records of their work may be regarded as historic milestones in the progress of the television art.

The important technical record of the National Television System Committee in its studies of monochrome broadcasting is contained in Donald G. Fink's *Television Standards and Practice*, published by the McGraw-Hill Book Company in 1943. Mr. Fink has done a similar service to the industry in editing the record of the deliberations of the NTSC leading to the present color television standards.

In *Color Television Standards*, Mr. Fink has elected himself to the thankless task of editing the voluminous record of the NTSC. However, he has chosen well, and the material he has selected for publication presents a very clear picture of the goal reached by NTSC and the path by which that goal was reached.

The author has written the first chapter, "The Development of Color Television," as a historic summary of the growth of interest in color in the period 1940 to 1953. He points out here how the second NTSC was formed and describes its organization. In the remainder of the book, each chapter is devoted to a particular part of the study of color television. The second chapter describes the standards finally proposed by the NTSC. These may be compared with the standards adopted by the commission which are quoted in appendices 1 and 2. Each succeeding chapter is devoted to the work of an individual panel. Their content may be gathered from the chapter headings: Subjective Aspects of Color Television; Color Video Signal; The Color Synchronization Signal; Field Tests of Compatibility; Field Tests of Color Performance; Field Tests of Networks and Transmitters; Color Film Processes and Transmission Equipment; and Definitions of Color Television Terms and Symbols.

As might be expected where the editor allowed the NTSC to tell its own story, occasional inconsistencies slipped by him. Mr. Fink has done remarkably well in catching most of them in this book. In fact, the only ones that troubled this reviewer were occasional references to documents by cabalistic series of letters and numbers to which there was no key.

Every engineer interested in color television will find *Color Television Standards* a valuable reference. Much of the tutorial material will be helpful to the beginner in the color field.

W. T. WINTRINGHAM
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Murray Hill, New Jersey

Principles of Nuclear Reactor Engineering by S. Glasstone

Published (1955) by D. Van Nostrand Co., Inc., 250 Fourth Ave., New York 3, N. Y. 838 pages+16 page index+ix pages+6 page appendix. 6¼×9¼. Illus. $7.95.

This book was prepared under the sponsorship of the Atomic Energy Commission to fill the need for a source book for reactor engineering education. It is designed as an unclassified text to prepare engineers for careers in nuclear engineering. As such it provides a broad coverage of a large number of topics in this field, including nuclear reactions and radiations, static and dynamic reactor theory, instrumentation and control, reactor materials, fuel and fuel processing, thermal considerations, radiation protection, and reactor design. It thus brings together in a single volume a tremendous amount of material previously available only in scattered form. The broad scope and a desire for completeness has prevented treating the individual subjects in very great detail. Indeed, in the rapidly advancing state of this art, only the specialist can hope to achieve a thorough mastery of any of these topics and keep abreast of new developments. The book does provide, though, an excellent foundation for those who wish to specialize in some particular aspect of nuclear engineering. The practicing engineer will also find here a practical reference and guide book in a rapidly growing field which requires cooperative effort between engineers and scientists from widely differing disciplines. The text has been contributed to by many authorities, most of them staff members of the Oak Ridge National Laboratory. References to the literature are copious.

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Closed-Circuit and Industrial Television by E. M. Noll

Published (1956) by the Macmillan Co., 60 Fifth Ave., N. Y. 11, N. Y. 227 pages+3 index pages+x pages. 128 figures. 9×6. $4.95.

This book covers three separate phases of the field of closed circuit and industrial television. The first chapter is devoted to a general discussion of the manifold uses of closed-circuit TV and cites a number of specific applications. The following six chapters, as stated by the author, provide "information needed by the electronics engineer and technician engaged in installing, operating, and servicing closed circuit units." The concluding chapter briefly describes the design and construction of a nine-tube composite camera that can be fabricated by one with some experience in electronics.

It is noted that detailed descriptive material on commercial systems is limited to that of three suppliers although today there are at least seven major companies engaged in the manufacture and sale of equipment designed for closed-circuit and industrial use. Also too limited in scope is the discussion of monitoring or viewing equipment, now available in many sizes and forms, which is an essential part of the closed-circuit system. The chapter on transmission emphasizes distribution to standard TV receivers using the modulated rf method but does not discuss the details of video transmission and distribution, either local or intercity. The index appears to be derived from section headings rather than from the text material. Examples of subjects discussed in the book but not appearing in the index are: photoconduction, cascode amplifiers, and hum.

Two chapters are devoted to basic information such as resolution, contrast, brightness, interlace, flicker, etc., while the bulk of the text covers specific circuits, their operation and their adjustment. Although the material will not completely serve the needs of systems planners or design engineers, it will be very helpful to the technician or student in the rapidly growing field of non-broadcast TV.

R. D. Chipp
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E. Paterson, N. J.

Frequency Response ed. by Rufus Oldenburger

Published (1956) by the Macmillan Co., 60 Fifth Ave., N. Y. 11, N. Y. 355 pages+15 index pages+xii pages. Illus. 11¼×8¼. $7.50.

This book contains eighteen articles originally presented at the 1953 A.S.M.E. Frequency Response Symposium and subsequently published in the *Transactions* of the A.S.M.E., Vol. 76, No. 8, 1954, together with ten additional articles. The 31 authors are experts in their respective fields both in this country and abroad. Included are two translations of important contributions by Russian engineers, L. C. Goldfarb and Ya. Z. Tsypkin. The book has been capably assembled and annotated by Dr. Oldenburger, Director of Research, Woodward Governor Company, and is appropriately dedicated to H. Nyquist. The title of the book does not clearly indicate the material coverage. A better idea of the contents can be obtained from the nine section headings as follows: Fundamentals; Frequency-Response Aids; Servo, Airplane and Power System Applications; Process Control; Transient Response; Optimum Controls; Nonlinear Techniques; Sampling Controls; and Statistical Methods.

Of the material covered the following are, in the reviewer's opinion, worthy of special comment: the inclusion of standards for the presentation of frequency-response data as recommended by A.S.M.E.—I.R.D. Dynamic Systems Committee, a bibliography of frequency response methods as applied to automatic-feedback-control systems including 284 entries broken down by subject and year, and an article on pneumatic, mechanical and electrical sine-wave generators for obtaining frequency response data.

This book does not cover some of the more sophisticated analyses on this subject that have appeared in the Proc. and similar publications. However it does contain a well integrated presentation of the more important and practical aspects of the subject. The book is therefore strongly recommended for those who are entering the field and those who want background information. The inclusion of analysis details of typical applications arising from several diversified fields considerably enhances the value of this book.

L. J. Giacoletto
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Ford Motor Company
Dearborn, Michigan

Recent Books

1954 Vacuum Symposium Transactions. Compiled by Committee on Vacuum Techniques, Inc., Box 1282, Boston 9, Mass. $10.00.

Abstracts of the Literature on Semiconduction and Luminescent Materials and Their Applications. Compiled by Battelle Memorial Institute. John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. $5.00.

Andres, P. G., Miser, H. J., and Reingold, Haim, *Basic Mathematics for Science and Engineering.* John Wiley and Sons, Inc., 440 Fourth Ave., New York, N. Y. $6.75.

Baker, C., *Technical Publications: Their Purpose, Preparation and Production.* John Wiley and Sons, Inc., 440 Fourth Ave., New York, N. Y. $6.00.

Forbes, G. F., *Digital Differential Analyzers: Part One, The Elements.* G. F. Forbes, 10117 Bartee Ave., Pacoima, Calif. $7.50.

Peek, R. L., Jr., and Wagar, H. N., *Switching Relay Design.* D. Van Nostrand Co., Inc., 250 Fourth Ave., N. Y. 3, N. Y. $9.50.

Review of Current Research and Directory of Member Institutions, ed. by Renato Contini. Engineering College Research Council of the American Society for Engineering Education, New York University, University Heights, New York 53, N. Y. $2.00.

Rider's Specialized Hi-Fi AM-FM Tuner Manual. Compiled by the John F. Rider Laboratory Staff. John F. Rider Publisher, Inc., 480 Canal St., New York 13, N. Y. $3.50.

Schultz, M. A., *Control of Nuclear Reactors and Power Plants.* McGraw-Hill Book Company, 330 W. 42 St., New York 36, N. Y. $7.50.

Schure, Alexander, *Crystal Oscillators.* John F. Rider Publisher, 480 Canal St., New York 13, N. Y. $1.25.

Smith, K. F., *Molecular Beams.* John Wiley and Sons, Inc., 440 Fourth Ave., New York, N. Y. $2.00.

U.R.S.I. Proceedings of the XI General Assembly: Vol. Ten, Part Three, Commission III on Ionospheric Radio. General Secretariat, 42 Rue des Minimes, Brussels, Belgium. $4.00.

Yarwood, J., *High Vacuum Technique, third ed., revised.* John Wiley and Sons, Inc., 440 Fourth Ave., New York, N. Y. $5.50.

Professional Groups†

Aeronautical & Navigational Electronics—James L. Dennis, General Technical Films, 3005 Shroyer, Dayton, Ohio.

Antennas & Propagation—H. G. Booker, School of Physics and Elec. Engrg., Cornell Univ., Ithaca, N. Y.

Audio—D. W. Martin, The Baldwin Piano Company, 1801 Gilbert Ave., Cincinnati 2, Ohio.

Automatic Control—J. C. Lozier, Bell Tel. Labs., Whippany, N. J.

Broadcast & Television Receivers—L. R. Fink, Research Lab., General Electric Company, Schenectady, N. Y.

Broadcast Transmission Systems—O. W. B. Reed, Jr., Jansky & Bailey, 1735 DeSales St., N.W., Washington, D. C.

Circuit Theory—H. J. Carlin, Microwave Res. Inst., Polytechnic Inst. of Brooklyn, 55 Johnson St., Brooklyn 1, N. Y.

Communications Systems—F. M. Ryan, American Telephone and Telegraph Co., 195 Broadway, New York 7, N. Y.

Component Parts—R. M. Soria, American Phenolic Corp., 1830 S. 54 Ave., Chicago 50, Ill.

Electron Devices—R. R. Law, CBS-Hytron, Danvers, Mass.

Electron Computers—J. D. Noe, Div. of Engineering, Research, Stanford Research Institute, Stanford, Calif.

Engineering Management—Rear Adm. C. F. Horne, Jr., Convair, Pomona, Calif.

Industrial Electronics—C. E. Smith, Consulting Engineer, 4900 Euclid Ave., Cleveland 3, Ohio.

Information Theory—M. J. Di Toro, Polytech. Research & Dev. Corp., 200 Tillary St., Brooklyn, N. Y.

Instrumentation—F. G. Marble, Boonton Radio Corporation, Intervale Road, Boonton, N. J.

(Professional Groups cont'd)

† Names listed are Group Chairmen.

Medical Electronics—V. K. Zworykin, RCA Labs., Princeton, N. J.
Microwave Theory and Techniques—H. F. Englemann, Federal Telecommunication Labs., Nutley, N. J.
Military Electronics—C. L. Engleman, 2480 16 St., N.W., Washington 9, D. C.
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North Carolina-Virginia (3)—M. J. Minor, Route 3, York Rd., Charlotte, N. C.; A. L. Comstock, 1404 Hampton Dr., Newport News, Va.

Northern Alberta (8)—J. E. Sacker, 10235—103rd St., Edmonton, Alberta, Canada; Frank Hollingworth, 9619—85th St., Edmonton, Alberta, Canada.

Northern New Jersey (2)—A. M. Skellett, 10 Midwood Terr., Madison, N. J.; G. D. Hulst, 37 College Ave., Upper Montclair, N. J.

Northwest Florida (3)—F. E. Howard, Jr., 573 E. Gardner Dr., Fort Walton, Fla.; W. W. Gamel, Canoga Corp., P.O. Box 188, Shalimar, Fla.

Oklahoma City (6)—C. M. Easum, 3020 N.W. 14th St., Oklahoma City, Okla.; Nicholas Battenburg, 2004 N.W. 30th St., Oklahoma City 6, Okla.

Omaha-Lincoln (5)—M. L. McGowan, 5544 Mason St., Omaha 6, Neb.; C. W. Rook, Dept. of Electrical Engineering, University of Nebraska, Lincoln 8, Neb.

Ottawa (8)—C. F. Pattenson 3 Braemar, Ottawa 2, Ont., Canada; J. P. Gilmore, 1458 Kilborn Ave., Ottawa, Ont., Canada.

Philadelphia (3)—M. S. Corrington, RCA Victor TV Division, Cherry Hill 204-2, Camden 8, N. J.; I. L. Auerbach, 1243—65th Ave., Philadelphia 26, Pa.

Phoenix (7)—Everett Eberhard, 30 E. Colter St., Phoenix, Ariz.; R. V. Baum, 1718 East Rancho Dr., Phoenix, Ariz.

Pittsburgh (4)—Gary Muffly, 715 Hulton Rd., Oakmont, Pa.; H. R. Kaiser, WHC-WWSW, Sherwyn Hotel, Pittsburgh 22, Pa.

Portland (7)—J. M. Roberts, 4325 N.E. 77, Portland 13, Ore.; D. C. Strain, 7325 S.W. 35 Ave., Portland 19, Ore.

Princeton (2)—J. L. Potter, Rutgers Univ., New Brunswick, N. J.; P. K. Weimer, RCA Laboratories, Princeton, N. J.

Regina (8)—William McKay, 2856 Retallack St., Regina, Saskatchewan, Canada; J. A. Funk, 138 Leopold Crescent, Regina, Saskatchewan, Canada.

Rochester (1)—W. F. Bellor, 186 Dorsey Rd., Rochester 16, N. Y.; R. E. Vosteen, 473 Badkus Rd., Webster, N. Y.

Rome-Utica (1)—M. V. Ratynski, 205 W. Cedar St., Rome, N. Y.; Sidney Rosenberg, 907 Valentine Ave., Rome, N. Y.

Sacramento (7)—E. W. Berger, 3421—5th St., Sacramento 20, Calif.; P. K. Onnigian, 4003 Parkside Ct., Sacramento, Calif.

St. Louis (6)—F. W. Swantz, 16 S. 23rd St., Belleville, Ill.; Gilbert Pauls, 1108 Pembroke Dr., Webster Groves 19, Mo.

Salt Lake City (7)—V. E. Clayton, 1525 Browning Ave., Salt Lake City, Utah; A. L. Gunderson, 3906 Parkview Dr., Salt Lake City 17, Utah.

San Antonio (6)—Paul Tarrodaychik, 215 Christine Dr., San Antonio 10, Texas; J. B. Porter, 647 McIlvaine St., San Antonio 1, Texas.

San Diego (7)—R. A. Kirkman, 3681 El Canto Dr., Spring Valley, Calif.; A. H. Drayner, 4520—62 St., San Diego, Calif.

San Francisco (7)—J. S. McCullough, 1781 Willow St., San Jose 25, Calif.; E. G. Goddard, 2522 Webster St., Palo Alto, Calif.

Schenectady (1)—J. S. Hickey, Jr., General Electric Co., Box 1088, Schenectady, N. Y.; C. V. Jakowatz, 10 Cornelius Ave., Schenectady 9, N. Y.

Seattle (7)—K. R. Willson, 1100—17th Ave 206, Seattle 22, Wash.; W. J. Siddons, 6539—39th N.E., Seattle 15, Wash.

Southern Alberta (8)—W. Partin, 448—22nd Ave. N.W., Calgary, Alberta, Canada; R. W. H. Lamb, Radio Station CFCN, 12th Ave. and Sixth St. E., Calgary, Alberta, Canada.

Syracuse (1)—P. W. Howells, Bldg. 3, Room 235, General Electric Co., Electronics Division, Syracuse, N. Y.; G. M. Glasford, Electrical Engineering Department, Syracuse Univ., Syracuse 10, N. Y.

Tokyo—Hidetsugu Yagi, Musashi Kogyo Daigaku, 2334 Tamagawa Todoroki 1, Setagayaku, Tokyo, Japan; Fumio Minozuma, 16 Ohara-Machi, Meguro-Ku, Tokyo, Japan.

Toledo (4)—M. E. Rosencrantz, 4744 Overland Parkway, Apt. 204, Toledo, Ohio; L. B. Chapman, 2459 Parkview Ave., Toledo 6, Ohio.

Toronto (8)—F. J. Heath, 830 Lansdowne Ave., Toronto 4, Ont., Canada; H. F. Shoemaker, Radio College of Canada, 86 Bathurst St., Toronto, Ont., Canada.

Tucson (7)—R. C. Bundy, Department 15, Hughes Aircraft Co., Tucson, Ariz.; Daniel Hochman, 2917 E. Malvern St., Tucson, Ariz.

Tulsa (6)—J. D. Eisler, Box 591, Tulsa 2, Okla.; J. M. Deming, 5734 E. 25th St., Tulsa, Okla.

Twin Cities (5)—J. L. Hill, 25—17th Ave. N.E., North St. Paul 9, Minn.; W. E. Stewart, 5234 Upton Ave. S., Minneapolis 10, Minn.

Vancouver (8)—J. S. Gray, 4069 W. 13th Ave., Vancouver, B. C., Canada; L. R. Kersey, Department of Electrical Engineering, Univ. of British Columbia, Vancouver 8, B. C., Canada.

Washington (3)—R. I. Cole, 2208 Valley Circle, Alexandria, Va.; R. M. Page, 5400 Branch Ave., Washington 23, D. C.

Williamsport (4)—F. T. Henry, 1345 Pennsylvania Ave., Williamsport, Pa.; W. H. Bresee, 818 Park Ave., Williamsport, Pa.

Winnipeg (8)—H. T. Wormell, 419 Notre Dame Ave., Winnipeg, Manitoba, Canada; T. J. White, 923 Waterford Ave., Fort Garry, Winnipeg 9, Manitoba, Canada.

Subsections

Berkshire (1)—A. H. Forman, Jr., O.P. 1-203, N.O.D., General Electric Co., 100 Plastics Ave., Pittsfield, Mass.; E. L. Pack, 62 Cole Ave., Pittsfield, Mass.

Buenaventura (7)—W. O. Bradford, 301 East Elm St., Oxnard, Calif.; M. H. Fields, 430 Roderick St., Oxnard, Calif.

Centre County (4)—W. L. Baker, 1184 Omeida St., State College, Pa.; W. J. Leiss, 1173 S. Atherton St., State College, Pa.

Charleston (3)—W. L. Schachte, 152 Grove St., Charleston 22, S. C.; Arthur Jonas, 105 Lancaster St., North Charleston, S. C.

East Bay (7)—H. F. Gray, Jr., 2019 Mira Vista Dr., El Cerrito, Calif.; D. I. Cone, 1257 Martin Ave., Palo Alto., Calif.

Erie (1)—R. S. Page, 1224 Idaho Ave., Erie 10. Pa.; R. H. Tuznik, 905 E. 25 St., Erie, Pa.

Fort Huachuca (7)—J. H. Homsy, Box 123, San Jose Branch, Bisbee, Ariz.; R. E. Campbell, Box 553, Benson, Ariz.

Lancaster (3)—W. T. Dyall, 1415 Hillcrest Rd., Lancaster, Pa.; P. W. Kaseman, 405 S. School Lane, Lancaster, Pa.

Memphis (3)—R. N. Clark, Box 227, Memphis State College, Memphis, Tenn. (Chairman)

Mid-Hudson (2)—J. C. Logue, IBM, Research Lab., Poughkeepsie, N. Y.; R. R. Blessing, IBM, Main Plant, Poughkeepsie, N. Y.

Monmouth (2)—G. F. Senn, Orchard Rd., River Plaza, Red Bank, N. J.; C. A. Borgeson, 82 Garden Rd., Little Silver, N. J.

Orange Belt (7)—F. D. Craig, 215 San Rafael, Pomona, Calif.; C. R. Lundquist, 6686 De Anza Ave., Riverside, Calif.

Palo Alto (7)—W. W. Harman, Electronics Research Laboratory, Stanford University, Stanford, Calif.; W. G. Abraham, 611 Hansen Way, c/o Varian Associates, Palo Alto, Calif.

Pasadena (7)—R. M. Ashby, 3600 Fairmeade Rd., Pasadena, Calif.; J. L. Stewart, Department of Electrical Engineering, California Institute of Technology, Pasadena, Calif.

Piedmont (3)—C. W. Palmer, 2429 Fairway Dr., Winston-Salem, N. C.

Quebec (8)—R. E. Collin, 41-B Boulevard des Allies, Quebec, P. Q., Canada; R. M. Vaillancourt, 638 Ave. Mon Repos, Ste. Foy, Quebec, Canada.

Richland (7)—R. G. Clark, 1732 Howell, Richland, Washington; R. E. Connally, 515 Cottonwood Dr., Richland, Wis.

San Fernando (7)—J. C. Van Groos, 14515 Dickens St., Sherman Oaks, Calif. (Chairman).

Tucson (7)—R. C. Eddy, 5211 E. 20 St., Tucson, Ariz.; P. E. Russell, Elect. Eng. Dept., Univ. Ariz., Tucson, Ariz.

USAFIT (5)—L. D. Williams, USAF Institute of Technology, MCLI, Box 3039, Wright-Patterson AFB, Ohio; G. P. Gould, Box 3274, USAFIT, Wright-Patterson AFB, Ohio.

Westchester County (2)—F. S. Preston, Norden Laboratories, 121 Westmoreland Ave., White Plains, N. Y.; R. A. LaPlante Philips Laboratories, Inc., S. Broadway, Irvington, N. Y.

Western North Carolina (3)—Officers to be elected.

Wichita (6)—M. E. Dunlap, 548 S. Lorraine Ave., Wichita 16, Kan.; English Piper, 1838 S. Parkwood Lane, Wichita, Kan.

Third Annual Meeting

SEPTEMBER 20–22, 1956

MELLON INSTITUTE & HOTEL WEBSTER HALL, OAKLAND DISTRICT PITTSBURGH, PENNSYLVANIA

SPONSORED BY THE PROFESSIONAL GROUP ON NUCLEAR SCIENCE

The Professional Group on Nuclear Science will hold its Third Annual Meeting at Mellon Institute and Hotel Webster Hall, Pittsburgh, Pennsylvania, September 20–22.

A tour of the Shippingport Nuclear power station is scheduled for the morning of September 22. The evening before, open house will be held at the new Westinghouse Research Laboratories. Ladies' activities include tours of the H. J. Heinz plant, Phipps Conservatory Flower Show, and Carnegie Institute.

A dinner meeting, at which Clark Goodman, Assistant Director for Technical Operations, Division of Reactor Development, U. S. Atomic Energy Commission, will speak, is planned for the evening of September 20 at Webster Hall Hotel. Tickets at $5.00 per person can be obtained at the registration desk.

Attendees may register at the registration desk in Mellon Institute which will open 8:00 A.M., September 20. The fees will be: PGNS members, $2.00; IRE members, $3.00; non-members, $4.00. Advance registrations and reservations can be obtained from J. B. Callaghan, Westinghouse Bettis Plant W3R-N, P. O. Box 1458, Pittsburgh 30, Pennsylvania.

All papers presented at the meeting will appear in a future issue of the PGNS TRANSACTIONS.

Thursday Morning, September 20

SESSION I

NUCLEAR SCIENCE—GENERAL

Research and Services for the Nuclear Industry, R. A. Brightsen, Nuclear Science & Engineering Corporation.

Electronics at the French Atomic Energy Commission, M. Surdin, French Atomic Energy Commission, Saclay.

Application of Radioactive Tracers to Oceanographic Studies, Ralph Ely, Jr., Nuclear Science & Engineering Corporation.

Comparison of Radiation Sources for Industrial Applications, K. H. Sun, Westinghouse (Materials Engineering Department).

Thursday Afternoon

SESSION II

COMPUTATION AND SIMULATION

Recent Results with a Nuclear Simulator, C. Caillet, French Atomic Energy Commission, Saclay.

Numerical Solution of the Two-Group Diffusion Equation in X-Y Geometry, R. S. Varga, Westinghouse (Bettis).

A Computer Code for the Solution of the Two-Group Diffusion Equations, W. R. Cadwell, Westinghouse (Bettis).

How to Measure Reactivity, G. S. Stubbs, Franklin Institute.

Nuclear Reactor Start-up Simulation, J. P. Franz and N. F. Simcic, Westinghouse (Bettis).

The Application of Digital Computer Techniques to Neutron Velocity Selector Experiments, H. L. Garner, University of Michigan.

Use of a Medium Scale Digital Computer for Processing of Large Scale Computer Results in Reactor Calculations, H. S. Bright, Westinghouse (Bettis).

Simulation of Hot Channel Boiling in Water Cooled Reactors, S. O. Johnson, J. V. Reihing, N. J. Curlee, Westinghouse (Bettis).

A Large Scale Automatic Scintillation Counting Facility with Digital Computer Data Reduction, K. Relf, O. F. Swift, Westinghouse (Bettis).

Friday Morning, September 21

SESSION III

INSTRUMENTATION

Instrumentation for Fast Neutron, Time of Flight Studies, R. V. Smith, Westinghouse (Research Laboratory).

On the Measurement of Transit Time Dispersion in Electron Multipliers, M. H. Greenblatt, RCA (Princeton).

French Reactor Instrumentation, J. Weill, French Atomic Energy Commission, Saclay.

Effects of a Nuclear Explosion on a Semiconductor Device, W. R. Langdon, General Electric (General Engineering Lab.).

Regarding Cosmic Ray Effects on Semiconductors, R. H. Vought, General Electric (Genl. Eng. Lab.).

Super Power Tubes for Particle Accelerator Applications, M. V. Hoover, RCA (Lancaster).

A Highly Reliable Radiation Monitor Set, W. E. Landauer, Airborne Instruments.

Measurement Techniques in Reactor Kinetic Studies—Spert Project, F. L. Bentzen, Phillips Petroleum.

Instrumentation and Reactor Safety, J. C. Simons, Jr. National Research Corporation.

Friday Afternoon

SESSION IV

REACTOR CONTROL

Reactor Plant Instruments and Control, W. H. Hamilton, J. E. Stell, Westinghouse (Bettis).

Two Examples of Automation in Control of Nuclear Reactors, P. Braffort, French Atomic Energy Commission, Saclay.

Frequency Response Measurements of Power Reactor Characteristics, H. Estrada, Westinghouse (Bettis).

Designing Heterogeneous Reactors for Stability, D. Little, M. Schultz, Westinghouse (Commercial Atomic Power).

Stability Analysis of Pressurized Water Reactors, T. E. Fairey, Westinghouse (Bettis).

Stability and Control of a Direct Cycle Boiling Water Reactor, R. P. Rose, J. N. Grace, Westinghouse (Bettis).

Comparative Stability of Pressurized and Boiling Water Reactors, J. MacPhee, American Machine and Foundry.

Canadian IRE Convention

OCTOBER 1—3, 1956

AUTOMOTIVE BUILDING, EXHIBITION PARK, TORONTO, CANADA

SPONSORED BY THE TORONTO SECTION AND REGION EIGHT

The Canadian IRE Convention and Exposition will be held in Toronto, October 1–3, 1956. C. A. Norris, general chairman, has announced that over 120 exhibitors have booked space. The convention marks the thirtieth anniversary of IRE in Canada.

Among the topics on the agenda will be electronic components, antennas, audio equipment, broadcast transmission systems, aeronautical and navigational electronics, tubes, transistors, electronic computers, instrumentation, medical electronics, industrial electronics, radio telemetry and remote control, vehicular communications, ultrasonics engineering, production machines, instruments and equipment, laboratory apparatus, measurement equipment, and packaging education. Altogether over

125 papers have been accepted by George Sinclair, technical program chairman. In addition, there will be exhibits dealing with the industrial applications of nuclear science and a display of electronic equipment as used by the various branches of the Armed Forces of Canada.

The Joint Service Committee of Canada's Department of National Defense is planning to participate in the convention to stress the importance of reliability and component parts standardization. Sessions will begin on Monday afternoon, continue all day Tuesday, with an evening session that day, and end Wednesday afternoon.

A convention banquet will be held in the concert hall of the Royal York Hotel on Monday evening, October 1. A ladies' program is also being arranged. Special entrance privileges are being arranged for university students to attend the technical papers sessions and exhibits.

Advance registrations for hotels can be arranged by writing to Grant Smedmor, IRE convention manager, 745 Mount Pleasant Road, Toronto 12, Canada.

Monday Afternoon, October 1

Broadcast

A New Broadcast Remote Control System, F. Mathers, Canadian General Electric Co. Ltd., Toronto.

A Portable Speech Input System Employing Transistor Amplifiers, W. J. Ives, Northern Electric Co. Ltd., Belleville.

A Graphic Volume Unit Recorder, D. H. McRae, Canadian Broadcasting Corp., Montreal.

A Canadian Transcontinental Microwave System, S. Bonneville, Canadian Broadcasting Corp., Montreal.

The London-Windsor Microwave System, R. D. Pynn, Canadian General Electric Co. Ltd., Toronto.

Semiconductors

Thermal and Field Effects in Point Contact Diodes, R. E. Burgess, University of British Columbia, Vancouver.

Inertia Phenomena in Photoconductors, Z. Szepesi, Canadian Marconi Co., Montreal.

Optical Shuttering with Barium Titanate, H. W. Jaderholm, Canadian Marconi Co., Montreal.

Double Junction Phototransistor Theory, H. J. Goldie, Northern Electric Co. Ltd., Montreal.

Recent Developments in the Diffusion Technique Used to Produce Semiconductor Devices, J. Y. Perron, Northern Electric Co. Ltd., Montreal.

Radio Relay Systems

A Subcarrier Type Microwave Communication System for 36 Voice Channels, C. R. Hill, Canadian General Electric Co. Ltd., Toronto.

An FM Demodulator Linearity Test Set, B. N. Sherman, Canadian Marconi Co., Montreal.

On Distortion in FM Subcarrier—AM Carrier Systems, R. Sandri and A. G. W. Timmers, Canadian Westinghouse Co. Ltd., Hamilton.

Supervision and Control of a Multihop Radio Relay System, R. J. Beddie and J. E. Raftis, Rogers Majestic Electronics Ltd., Leaside.

The Quebec Hydro Microwave Communication System, J. G. Sutherland, J. Leahy and G. F. Baylis, RCA Victor Co. Ltd., Montreal.

Measurements

A Standing Wave Line for Low Radio Frequencies, E. A. Walker, D.R.T.E., Defense Research Board, Ottawa.

High Precision Standard of Frequency, S. N. Kalra, Nation Research Council, Ottawa.

Automation in the Laboratory—Description of a Strain Recording and Plotting System, G. F. Kelk, G. F. Kelk and Co., Toronto.

Improved Calorimeters and Loads For Better Measurement and Absorption of Microwave and Lower Frequency Power, S. Freedman, Chemalloy Electronics Corp., Santee, Calif.

Video Transmission Requirements and Testing Techniques, A. Ste. Marie, Canadian Broadcasting Corp., Montreal.

Electronics for Defense

Electronics for Defense, M. L. Card, Dept. of National Defense, Ottawa.

Automatic Range Radar for the F.86 Aircraft, L. W. deCocq, Canadian General Electric Co. Ltd., Toronto.

Packaging Guided Missile Electronics, J. W. Keenan, Canadian Westinghouse Co. Ltd., Hamilton.

The Control of Guided Missiles, A. Ratz, Canadian Westinghouse Co. Ltd., Hamilton.

Radio Telemetering in Guided Missile Development, A. L. Lortie, C.A.R.D.E., Quebec.

Tuesday Morning, October 2

Antennas I

Development Report on Tunable Microwave Bandpass Filters and Duplexing Filters, B. Vural, Canadian General Electric Co. Ltd., Toronto.

Some Aspects of Band Pass and Band Reject Filters Used for Duplexing, W. V. Tilston, Sinclair Radio Laboratories Ltd., Toronto.

Pattern Range for HF Shipborne Antennas, J. Y. Wong and J. C. Barnes, National Research Council, Ottawa.

Investigation of the Pattern of a Ground Plane Antenna, A. H. Secord, Sinclair Radio Laboratories Ltd., Toronto.

Some Conductivity Characteristics of Canadian Terrain at Medium Radio Frequencies, P. A. Field, D.R.T.E., Defense Research Board, Ottawa.

Nucleonics I

Practical Implications in the Routine Measurement of Low Concentrations of Alpha Emitting Isotopes, J. Nicholls, Canadian Westinghouse Co. Ltd., Hamilton.

Problems in Radiation Instruments for Defense, A. Hendrikson, D.R.T.E., Defense Research Board, Ottawa.

Use of Transistors in Nucleonics, F. S. Goulding, A.E.C.L., Chalk River.

Millimicrosecond Time Measurements, R. E. Bell, McGill University, Montreal.

New Scaling Techniques, W. D. Howell, A.E.C.L., Chalk River.

Color Television

Testing Facilities for Color Television Receivers, R. Anthes, Canadian Westinghouse Co. Ltd., Brantford, Ontario.

Video Transmission Testing Techniques for Monochrome and Color, J. Raymond Popkin-Clurman, Telechrome Mfg. Corp., Amityville, L.I., New York.

Measuring Equipment for Chrominance Channel Characteristics, P. A. Wigley and W. Shurben, Canadian Radio Mfg. Corp. Ltd., Toronto.

Design Considerations for Color Television I.F. Amplifiers, K. R. Van der Keyl, Canadian Radio Mfg. Corp. Ltd., Toronto.

Sub-Carrier Matrixing in Color Television, A. E. Kimmel, Canadian Radio Mfg. Corp. Ltd., Toronto.

Circuits

Design of Oscillators to Temperature Compensate Inductance Type Transducers, A. G. Christensen, Phoenix Engineered Products Ltd., Toronto.

Wide Band Power Amplifiers, P. Gomard, T. S. Farley Limited, Hamilton.

AC Gain Stabilization by Use of DC Degeneration, H. P. Moen and D. A. Anderson, Canadian Marconi Co., Montreal.

A New Approach to Variable Frequency Oscillator Design, F. A. Baily, Canadian Marconi Co., Montreal.

The Application of Phase-Locked Frequency Control Systems, E. H. Hugenholtz, Rogers Majestic Electronics Ltd., Leaside.

Symposium: Reliability and Quality Control

Introduction of Panel Discussion, R. A. Muller, Chairman, Canadian General Electric Co. Ltd., Toronto.

Design Considerations for Reliability, P. E. J. Wilburn, Canadian General Electric Co. Ltd., Toronto.

Production Practices for Reliability, P. D. Balmer, Canadian General Electric Co. Ltd., Toronto.

Organization and Analysis of Field Reports for Reliability, A. S. Best, Canadian General Electric Co. Ltd., Toronto.

Applications of Statistical Quality Control, J. J. Fitzsimmons, Canadian Marconi Co., Montreal.

Fundamentals of Statistical Quality Control, J. B. Pringle, Bell Telephone Co. of Canada, Montreal.

Tuesday Afternoon, October 2

Antennas II

Broadband Centre-Fed Slotted Cylinder Antenna, N. Tomcio, Canadian General Electric Co. Ltd., Toronto.

Two-Dimensional Slotted Arrays, G. C. McCormick, National Research Council, Ottawa.

Designs Procedures for Small Annular Slot Antennas, W. A. Cumming, National Research Council, Ottawa.

A Broadly Tunable Antenna System, S. Presentey, Canadian Marconi Co., Montreal.

Development of Cross-Polarized Antennas, R. Meier, RCA Victor Co. Ltd., Montreal.

Nucleonics II

Electronic Instrumentation for the Location and Assaying of Radio Active Ores, G. G. Eichholz, Dept. of Mines & Technical Surveys, Ottawa.

Nuclear Power Plant Analog for NPD, W. S. Brown, Canadian General Electric Co., Peterborough.

System Reliability in Reactor Control, E. E. Siddal, A.E.C.L., Chalk River.

Magnetic Amplifier Servo Control for NRU, N. F. Wood, Ferranti Electric Ltd., Toronto.

Instrument Reliability in Reactor Instrumentation, A. Pearson, A.E.C.L., Chalk River.

Monochrome Television

A Tuning Indicator for Television Receivers, W. E. Liddell, Canadian Radio Mfg. Corp. Ltd., Toronto.

Closed Circuit Television, W. M. Booth, Rogers Majestic Electronics Ltd., Leaside.

Television Distribution Systems, E. O. Swan, Ernie Swan Television Co., Ltd. Toronto.

VHF Television Relay and Booster System, V. E. Isaac, A. Hodgson, J. E. Pauch, RCA Victor Co., Ltd., Montreal.

Interference Immunity of TV Receivers, E. Luedicke, RCA Victor Co. Ltd., Renfrew

Electronic Tubes

Microphonic Testing of Tubes, S. Love, Radio Valve Company of Canada Ltd., Toronto.

Cold Cathode Tubes, A. F. Knowles, Rogers Majestic Electronics Ltd., Leaside.

The 6CW5 and Its Application in a Biampli High Fidelity System, R. J. A. Turner, Philips Industries Ltd.

Design and Performance of a 2KW CW Klystron Amplifier for C-Band, E. A. Conquest, Varian Associates of Canada Ltd., Georgetown.

An Investigation into Simple Methods of Forecasting the Life of Electron Tubes with Indirectly Heated Cathodes, R. H. Taplin, Canadian Marconi Co., Montreal.

Data Processing

Some Applications of Electronic Data Processing Systems in the Air Transport Industry, L. E. Richardson, Trans-Canada Air Lines, Montreal.

Automatic Recording and Processing of Operating Data at Hydro's Niagara River Plants, J. R. Leslie, H.E.P.C., Toronto.

Direct Simulation of Analog Computers through Signal Flow Graphs, L. P. Robichaud, C.A.R.D.E., Quebec.

Magnetostriction Delay Lines, J. V. Scott, Ferranti Electric Ltd., Mount Dennis, Toronto.

Applications of Symbolic Logic to Electronic Engineering, G. B. Thompson, Northern Electric Co. Ltd., Belleville.

Tuesday Evening, October 2

Electronic Sorting of Mail—A First for Canada—Part I—The Canadian Post Office Electronic Mail Sorter, W. J. Turnbull, Deputy Postmaster General. *Part II—Fundamental Principles of the Canadian System*, M. Levy, Post Office Dept., Ottawa.

The Route Reference Computer, C. G. Helwig, H. B. Brown, L. R. Wood, Ferranti Electric Ltd., Toronto.

Wednesday Morning, October 3

Scatter Propagation

A Reflection Theory for Beyond the Horizon Propagation, H. T. Friis, A. B. Crawford and D. C. Hogg, Bell Telephone Laboratories, Holmdel, N.J.

2700 MC/S Scatter Propagation between Ottawa and Toronto, L. H. Doherty, National Research Council, Ottawa.

Detailed Performance Characteristics of Communication Circuits Employing Tropospheric Scatter Propagation, R. M. Ringoen, Collins Radio Co. of Canada Ltd., Toronto.

System Design Problems Associated with VHF Scatter Circuits, J. W. Smith, Collins Radio Co. of Canada Ltd., Toronto.

Scatter Propagation, P. L. Rice, C.R.P.L. Bureau of Standards, Boulder, Colo.

Transistor Circuitry

High Frequency Transistor Amplifier Design, G. T. Lake, D.R.T.E., Defense Research Board, Ottawa.

Transistor DC Converters, G. M. Kerrigan, Computing Devices of Canada Ltd., Ottawa.

Transistor Circuitry for Wide Ranges of Environmental Temperature, W. Greatbatch, Consultant, Taber Instrument Co., N. Tonawanda, N.Y.

Transistor Parameter Variations, S. V. Soanes, Ferranti Electric Ltd., Toronto.

Logical Use of Transistors in Communications Applications, S. Kagan, Crosley Defense and Electronics Div., Moffats Ltd.

Medical Electronics

Analysis of Heart Murmurs by Electronics, R. S. Richards, National Research Council, Ottawa.

Electronic Applications in Cardiovascular Surgery, J. A. Hopps, National Research Council, Ottawa.

Low-Frequency Analyzers in Electro-Physiology, J. F. Davis, McGill University, Montreal.

The "SCAD"—A Servo Calibrating Auto Densitometer, P. E. P. Smith, Electrodesign, Montreal.

A Demonstration Oscilloscope and Electro-Physiological Unit, P. Sekelj, B. D. Burns and C. Pinsky, McGill University, Montreal.

Electronics in Medicine, W. E. Hodges, Consultant, Toronto.

Manufacturing Techniques

The Quality Control of Printed Circuit Board Manufacturing, F. H. Edwards, United-Carr Fastener Co. of Canada Ltd., Hamilton.

Mechanized Processes for Design and Manufacture of Printed Wiring Units for Data Processing Systems, D. E. Nuttall, Ferranti Electric Ltd., Toronto.

Discussion of Methods of Producing Prototype Printed Circuit Boards, I. Meitlis, Canadian Gen. Elec. Co. Ltd., Toronto.

Epoxy Casting of Electronic Circuitry, D. L. Harvey, Canadian General Electric Co. Ltd., Toronto.

Development and Installation of a 250 KW Low Frequency Transmitter, V. Ziemelis, RCA Victor Co. Ltd., Montreal.

Components and Materials

Magnetic Recording Tape, L. F. Bennett, C.A.M.E.S.A., Ottawa.

A Study of Sintered Oxide Mixtures as Resistive Materials, J. A. Cowan, D.R.T.E., Defence Board, Ottawa and J. H. Simpson, National Research Council, Ottawa.

Microwave Hybrids Using Strip Line, R. McClelland, Canadian Marconi Co., Montreal.

Ferrites in Microwave Work, D. J. Whale, Canadian Westinghouse Co. Ltd., Hamilton.

General Design Factors and Characteristics of Foil Tantalytic Capacitors, F. R. Flood, Canadian General Electric Co. Ltd., Hudson Falls, N.Y.

Propagation

The Cost of Decibels, Frederick Gall, Canadian Marconi Co., Montreal.

Microwave Refractometer Measurements of Atmospheric Refractive Index, A. W. Adey and W. J. Heikkla, R.P.L., Defense Research Board, Ottawa.

An Experimental Investigation of the Diffraction of Electromagnetic Waves by a Dominating Ridge, J. W. B. Day, J. H. Crysdale and W. S. Cook, R.P.L., Defense Research Board, Ottawa.

Electronics in Meterology, W. R. Smith, Department of Transport, Ottawa.

Applications of Transistors to Computers

A P-N-P-N Bistable Element Suitable for Digital Computers, N. F. Moody, D.R.T.E., Defense Research Board, Ottawa.

Computer Circuits Using the P-N-P-N Transistor Element, C. D. Florida, D.R.T.E., Defense Research Board, Ottawa.

Transistorized Logical Building Blocks, D. C. Redpath, Crosley.

The Systems Design of a Transistorized General Purpose High Speed Digital Computer, D. F. Parkhill and G. G. Desloovere, Crosley.

A Design Method for Direct-Coupled Transistor Computer Circuits, G. S. Collins, C. G. Helwig, D. K. Ritchie and R. S. Wedgewood, Ferranti Electric Limited Toronto.

Communications

Public Mobile Telephone Service Problems, S. H. Whitaker, Bell Telephone Co. of Canada, Montreal.

Some Considerations in Radio-Telephone Interconnections, A. Lovas, Canadian General Electric Co. Ltd., Toronto.

A Canadian Designed 150 mc Vehicular Communication Equipment, G. M. Koch and W. Ornstein, Canadian Marconi Company, Montreal.

High Frequency Single Sideband Techniques, W. S. Bruene Collins Radio Co. of Canada Ltd., Toronto.

An Experimental Radio Teleprinter Broadcast Service for North-Atlantic Air Routes, B. G. Doutre, Trans-Canada Air Lines, Montreal.

MANAGEMENT

Cost Reduction and Product Improvement, E. H. Tovee, Canadian Westinghouse Co. Ltd., Hamilton.

Planning, Controlling and Measuring Engineering Projects, J. M. Toye, Canadian General Electric Co., Ltd., Toronto.

Common Errors in Measurement and Design, W. C. Benger, Northern Electric Co. Ltd., Belleville.

The Organization of a Transistor Measurements Laboratory, D. P. Henderson, D.R.T.E., Defense Research Board, Ottawa.

Cascade Co-Operative Education, Richard Scott, Can. Aviation Elec. Ltd., Montreal,

ELECTRONIC APPLICATIONS

Magnetoresistive Amplifiers A. Aharoni and E. H. Frei, Weizmann Institute of Science, Rehovoth, Israel.

Automatic Direction Finder Type CMA-301, E. W. Beasley and T. Janssen, Canadian Marconi Company, Montreal.

A Microwave Position-Fixing System F. R. Park and K. Ayukawa, National Research Council, Ottawa.

Some Considerations Affecting the Design or Servos in Flight Simulator Applications, F. Borlace, Canadian Aviation Electronics Ltd., Montreal.

Electronic Power Supplies, T. C. Gams, NJE Corp., Kenilworth, N.J.

Second National Symposium on Aeronautical Communications

The IRE Professional Group on Communications Systems is sponsoring the Second National Symposium on Aeronautical Communications at Hotel Utica, Utica, New York, October 8–10, 1956.

The 1956 Symposium will stress communications systems in support of present and future aeronautical activities and will be of interest to all those engaged in the communications and aeronautics fields.

On October 10, a symposium will be held at Rome Air Development Center, Griffiss Air Force Base, Rome, New York, which will be classified *confidential*. Participation will be limited to those possessing the required clearances. It is suggested that those desiring to attend arrange for the proper security clearance through the appropriate channels at an early date in order to avoid confusion and embarrassment.

A cocktail party will be held Monday evening followed by the symposium banquet and a luncheon will be held on Tuesday at the Hotel Utica.

Officers of the symposium are as follows: *Chairman*, J. W. Worthington, Jr.; *Advisory Staff*, H. Davis, M. R. Johnson, H. F. Konig, A. A. Kunze, H. F. Mayer, P. J. Schenk, Brig. Gen. A. R. Maxwell, and Maj. Gen. S. P. Wright; *Executive Vice-Chairman*, R. S. Grisetti; *Arrangements*, R. E. McMillan; *Exhibits*, C. B. Appleman; *Finance*, H. J. Crowley; *Guest Speakers*, R. L. Marks; *Publicity*, R. C. Benoit, Jr.; *Registration*, F. R. Priest; *Technical Program*, F. Koskowitz.

Monday, October 8

10:00–12:30

SESSION I

COMMUNICATION SYSTEM CONCEPTS

Session Chairman: General P. Sandretto, Federal Telecommunications Laboratory.

USAF Aeronautical Communications: A Link in the Servo Control Loop, Lt. General Joseph Smith, Hq. MATS.

AF Communications Problems and the Future Air Force Operational System, C. K. Chappuis, Rome Air Development Center.

A New Look at Communications in the Field Army, H. P. Hutchinson.

The Four Systems Tests, Major Walter White, Jr., Office of the Chief Signal Officer, Department of the Army.

Aeronautical Communications Contribution to Public Safety Communications Systems, W. C. Collins, Dept. of Communications of Los Angeles.

2:00–4:30

SESSION II

EQUIPMENTS AND SYSTEM COMPONENTS

Session Chairman: Harry Mayer, General Electric Co.

The Atomichron, An Atomic Frequency Standard, R. T. Daly, National Company.

A One-Kilowatt High Level Modulated UHF Amplifier with Low Distortion, C. R. Ellis, General Electric, K. H. Owen, McIntosh Electronics, G. R. Weatherup, Rome Air Development Center.

A UHF Exciter for AM, FM or SSB, Peter Hume, Westinghouse Electric Corp.

A VHF-UHF Antenna Multicoupler Employing Resonant Cavities in Tandem, M. W. Caquelin, Collins Radio Co.

A 600-Kilowatt High Frequency Amplifier, J. O. Weldon, Continental Electronics.

Tuesday, October 9

9:30–12:00

SESSION III

COMMUNICATION SYSTEM CONSIDERATIONS

Session Chairman: R. I. Cola, Melpar, Inc.

A Clinical Approach to Engineering Development, R. J. Silbiger, Norden-Ketay Corp.

Results of UHF Mutual Environment Test Program at RADC, Joseph Berliner and John Augustine, RADC.

A Method for Studying Data Transmission Requirements for Large Systems, J. E. Barmack, Dunlap & Associates, Inc.

A Compatible Single-Sideband System, L. R. Kahn, Kahn Research Laboratories.

An Integrated High Frequency Single Sideband System, M. I. Jacob, Westinghouse Electric Corp.

2:30–5:00

SESSION IV

COMMUNICATION TECHNIQUES

Session Chairman: Harry Davis, Technical Director, Rome Air Development Center.

Factors Affecting Intelligibility of Single-Sideband Voice Communication, N. H. Young, Federal Telecommunications Laboratory

Synchronous Communications, J. P. Costas, General Electric Co.

The Possibility of Extending the Range of VHF-UHF Aeronautical Communications, T. F. Rogers, L. A. Ames, Lt. E. J. Martin, AF Cambridge Research Center.

An Incredutor Tuned VHF Panoramic Receiver, C. G. Sontheimer, CGS Laboratories, Inc.

Nuclear Radiation Applications to Aeronautical Navigational Aids for Airfield Guidance During Final Approach, M. J. Cohen, R. M. Wilmotte, Inc.

Wednesday, October 10

GRIFFISS AIR FORCE BASE

9:30–12:00

SESSION V

Confidential Symposium on Military Data Links.

2:00–4:30

SESSION VI

Confidential Symposium on Long-Range Communications.

Abstracts of IRE Transactions

The following issues of "Transactions" have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
Aeronautical & Navigational Electronics	Vol. ANE-3, No. 2	$1.40	$2.10	$4.20
Audio	AU-4, No. 3	.80	1.20	2.40
Broadcast & TV Receivers	Vol. BTR-2, No. 1	1.10	1.65	3.30
Instrumentation	PGI-5	3.20	4.80	9.60
Vehicular Communications	PGVC-6	1.55	2.30	4.65

* Public libraries and colleges may purchase copies at IRE Member rates.

Aeronautical & Navigational Electronics

Vol. ANE-3, No. 2, June, 1956

1956 Pioneer Awards in Airborne and Navigational Electronics

Recent Developments in the Simulation of Terminal Area and En Route Area Air Traffic Control Problems—T. K. Vickers and R. S. Miller

This paper briefly describes the air traffic simulation facilities presently installed at the CAA Technical Development and Evaluation Center (TDEC). It then reviews some important developments which have been achieved through the use of these facilities in the study of terminal area and en route area traffic control problems.

A Servomechanism Approach to the Problem of Communication for Aircraft Control—S. J. O'Neil

The traffic-handling capability of a communication system used for aircraft control during airport traffic control, ground-controlled approach, and ground-controlled landing is studied. The system is treated as a multiloop servomechanism which includes both ground and airborne equipment. The factors which affect the location of the communications link in the system are discussed. These include message rate, airborne equipment or ground measurements necessary, flight geometry, and flight safety. It is shown that the minimum message rate is possible if the quantity transmitted is the same as the flight motion desired. Several methods are presented for increasing the traffic-handling capability by reducing the message rate. One of these introduces controlled backlash into the outermost feedback loop and automatically determines the minimum message rate while navigating to a point. A dynamic analysis is used to determine the limitations of the method. The methods presented are illustrated by showing their application to automatic flight control during ground-controlled approach and ground-controlled landing.

An Improved Simultaneous Phase Comparison Guidance Radar—H. H. Sommer

Simplification of Airborne Navigation by Use of the Vortex Thermometer—R. E. Ruskin

Means are described for reducing the complexity of the electronics usually required in airborne navigational computing systems to determine true free-air temperature and true air speed. The vortex thermometer, when equipped with a linear resistance-temperature element, permits use of a linear potentiometer in a simple digital null-balancing servo. The vortex thermometer eliminates corrections which are otherwise needed for dynamic heating due to the speed of the plane.

If true free-air temperature is measured along with the temperature of a probe having a known higher recovery factor, the difference may be combined directly using linear resistance thermometers to give an indication of true air speed without use of either pitot or static pressure measurements. Mach number and altitude may also be computed from these temperature data, together with pitot pressure, but without measurement of static pressure which is usually subject to considerably more error than pitot pressure.

Wide Band Instrumentation—R. L. Strazzulla

Wide-bandwidth measurement equipment is used where both long and short time accuracy is required. Present sensing devices can be broken into two classes, namely those that function accurately over long time measurements and conversely those that indicate changes but are lacking long time stability. Pressure pickoffs and radars are examples that fall into the former class of devices while velocity pickoffs and accelerometers possess only short time accuracy.

This paper presents a technique whereby both types of inputs are injected into a mixer whose output retains the good properties of each input while rejecting the inaccurate portions of the input data. The resultant output of the mixer is superior to either input, since it possesses both high and low frequency accuracy.

A general method for solution of the mixer characteristic is given which indicates that complementary linear filters may be used to perform the mixing function. The system may be made dynamically exact, that is, the mixer has a unity transfer function and contributes no error in the absence of noise.

The mixer parameters are chosen on the basis of minimizing the root mean square error in the output considering the noise inputs to the system.

An illustrative example is given based on a Baro-Inertial Mixer using both pressure and inertial data to give accurate wide band vertical velocity as its output.

Random Time-Modulation of the Main Bang for Increased Accuracy in Digital Range Measurement—L. B. Harris

Because of its conceptual simplicity, random modulation of the main bang is an attractive solution to the resolution problem in digital ranging. The accuracy of such a scheme in the presence of gate jitter has been analyzed in this paper and the results show that it can effectively reduce the resolution error.

PGANE News

Contributors

Roster of Members—Professional Group on Aeronautical and Navigation Electronics

Audio

Vol. AU-4, No. 3, May–June, 1956

Message from the New National Chairman

PGA News

Magnetic Recording—1882–1952—C. F. Wilson

Thirty-eight of the more important patents, from more than five hundred that were investigated, have been listed along with a complete bibliography of magnetic recording covering the years 1888–1952.

Broadcast & TV Receivers

Vol. BTR-2, No. 1, April, 1956

Design Considerations in the Reduction of Sweep Interference from Television Receivers—A. M. Intrator

A Discussion of the Design Problems Encountered in the Development of a Transistorized Radio Receiver—J. A. Worcester

Tentative Methods of Measurement of Color Television Receiver Performance—S. P. Ronzheimer and R. J. Farber

Instrumentation

PGI-5, June, 1956

(*IRE Instrumentation Conference and Exhibit—November 28–30, 1955—Atlanta, Georgia*)

The Impact of Automation on Data Processing—W. S. Buckingham, Jr.

Control and Power Supply Problems of Unmanned Satellites—Ernst Stuhlinger

Schemes for an attitude control system and for electric power supplies for an unmanned satellite of a few hundred pounds total weight are presented. The attitude of the satellite with respect to the earth's center is controlled within ±10 degrees by utilizing the shadowing effect of the earth on the isotropic cosmic radiation. A number of Geiger counters are arranged so that they sense the location of the shadow cone of the earth. Signals resulting from the counting rates of the counters control fly wheels that cause the satellite to rotate around its center of gravity.

Three sources for electric power are described, each of which delivers an average of about 100 watts. The first converts the sun's radiating energy with a silicon junction photoelectric generator. A sun-seeking device keeps the generator oriented toward the sun during daytime. In the second system, the sun's radiation is directed toward a pile of thermocouples made of ZnSb and constantan. Thermocouples have been built of these materials which convert solar energy into electric energy with an efficiency of 5.6 per cent. The third method uses a radioactive isotope, strontium 90, and its daughter product, yttrium 90, as heating element for a pile of thermocouples. The half life of strontium 90 is 20 years. Each of these three sources has a specific power production of the order of 0.4 to 0.7 watts per pound of weight. The attitude control system and the methods of power supply described are applicable also to larger satellites. If the total power to be provided is of the order of 20 kilowatts or more, a steam-electric generator with the sun, a radioactive isotope, or a nuclear reactor as heat source becomes more efficient than the systems described here. For a total operating time of only 2 or 3 days, dry cell batteries are preferable to any other system at power levels up to a few hundred watts.

A Twenty-Four Channel Cathode Ray Oscilloscope for Monitoring Magnetic Tape Records—F. C. Smith, Jr., and R. R. Pittman

Magnetic tape is not directly susceptible to visual examination, a technique which still has some value in this age of automation. Oscillographic recording is time-consuming and expensive, and where a permanent visible record is not required, the cathode ray oscilloscope suggests itself as the ideal tool.

A twenty-four channel oscilloscope has been developed to meet this need. A twenty-one inch single-gun television tube, standard except for the aluminized P-7 phosphor, is used as the viewing unit. Magnetic deflection is used to provide the raster and sweep, while the signals appear as intensity modulation. A comparison circuit eliminates the need for electronic switching, and the amplitude, linearity, and frequency response are not affected by the number of channels displayed.

Simplified Automatic Data Plotter for Telemetering Systems (SADAP)—H. B. Riblet

This data plotting system utilizes magnetic tape storage of telemetering information; an electronic system for the scaling of time signals; and an optical-mechanical system which simultaneously records on photographic film a horizontal function scale grid, vertical timing lines, and the galvanometer trace of the telemetered function. The function scale grid can be produced by a photographic process with errors less than 0.25 per cent from the calibration curve of the telemetering system by means of a specially designed optical-mechanical pantograph.

This system produces a continuous graphic plot of telemetered information on a linear or nonlinear scale (as may be necessary) with errors less than 1 per cent. The system essentially produces its own graph paper during the plotting process without the use of complex electronic circuity. Two functions may be plotted on one film and are automatically synchronized in time. The evaluation of this system has proven a savings of 20 or 30 to 1 in time and manpower over standard plotting techniques.

New Airborne Recorder Design Techniques—A. L. Klein

A new magnetic tape recorder designed specifically for airborne and mobile use is described. This recorder combines performance and versatility usually associated with laboratory recorders. It operates within specifications in the environmental conditions specified in MIL-E-5400. Using miniaturized components and modular construction, the Ampex Model 800 Series recorders accept data in the three electrical forms for which laboratory equipment is presently available.

A High-Speed Reader for Perforated Tape—R. J. Bianco

A perforated tape reader which is simple in concept, is troublefree and easy to operate, and is very fast (1000 characters per second) is described. The reading head consists of six photoelectric cells to read standard five-channel tape. The sprocket hole is used for timing purposes. The design could easily be expanded for six, seven, or eight-channel tapes and for higher speeds. However, no attempt is made to stop the tape on an individual character or between characters. Schematic diagrams and typical observed waveshapes are included.

Permanent Digital Function Storage Using Neon Tubes—M. S. Raphael and A. S. Robinson

For some purposes, such as storage of tables needed in computations, there is no need to ever change the contents of the memory. The information may therefore be permanently wired in at the time of construction or be changed only by some manual operations. A memory of this type utilizing the presence or absence of a small neon tube as the memory element is presented here. The layout of the memory is in effect rows and columns of wires with the neon tubes connected to the appropriate intersections. The neon tube detects a "coincident voltage" at the intersection and the resultant light output is converted back to voltage by means of a photo tube. Information can be obtained from a new location in the memory in less than 100 μsec and since no regeneration is necessary, over 10,000 bits per second may be obtained from each memory matrix.

A Survey of Navigational Measurements Methods for Missile Guidance Systems—S. L. Johnston

The basic methods of obtaining navigational data for guided missiles are reviewed. These are grouped in methods of measurement of distance, velocity, and angle. A generalized approach is used to indicate the similarity between certain optical and radar measurement schemes.

A Central Facility for Processing Engineering Test Data—E. C. Allmon

At the Air Force Armament Center at Eglin Air Force Base, Florida, very comprehensive facilities are utilized in the engineering testing of air armament items. The large volumes of data generated must be suitably recorded, then processed, mathematically manipulated, then delivered to project personnel. To accomplish this operation, which is considered to constitute "data processing," there must be adequate devices for sensing analog and digital data parameters, transmitting data to appropriate points, recording data, and then processing through a central computing facility. At the Air Force Armament Center, a large electronic digital computer was developed and installed to serve as the "nerve center" of such a central data processing facility. This computer is augmented by special input and output devices which enable it to communicate with its outside environment at very high speeds.

Requirements of Data Processing Facilities—I. R. Heimlich

A survey of flight test requirements reveals that the bulk of the data can be processed with relatively small quantities of equipment and manpower. The paper briefly reviews some of the economic considerations of analog and digital processing techniques.

A Digital Data-Gathering System—C. Fanwick, J. S. Lanza and J. Ottobre

The equipment to be described provides facility for monitoring, recording and storing angular position data. It was developed under a Bureau of Ordnance contract for use in a synchro data transmission system but is not limited to this application. Two separate equipments comprise the system. A Data Recorder which accepts the electrical outputs of nine sets of dual-speed synchro generators (or less); converts the sine wave outputs to pulses; counts these to determine a time interval between the pulses representing the order and a reference pulse representing zero order; records the count on magnetic tape. The tape is applied to a laboratory equipment—the Data Reproducer—which, from the pulse code on the tape, makes trigger signals available to a card punching device which perforates a card in such manner as to completely express the position of the synchro generator shaft at the time of sample.

Digital Solutions to Instrumentation and Automatic Control Problems—Benjamin Kessel and R. W. Brooks

Research and industry are constantly demanding more accurate, reliable, and flexible means of instrumentation. The purpose of this paper is to discuss the application of digital techniques to problems which may recently have been considered more in the realm of analog or other non-digital techniques. Invariably digital techniques produce a system that is extremely flexible, swift, accurate, and reliable.

The Analog Computer as an Operational Test Instrument for Jet Engine Testing—L. F. Burns and W. K. McGregor

One of the most cumbersome features of wind tunnel testing has been data taking and data reduction. This is mainly because of instrumentation methods which require step by step procedures from raw data to reduced data. In many tests it would greatly enhance the test value if immediate indication of reduced parameters such as combustor efficiency or calculated thrust were available. One method being attempted at the Engine Test Facility, Arnold Engineering Development Center for providing these indications is through use of electronic analog computer equipment. Some of the equations encountered and the method of computing the reduced values are presented in this paper. The instrumentation for transducing the fundamental measurements is investigated. Finally, some conclusions are drawn about the accuracy and usefulness of such an instrument.

An Analog Data Handling System—J. M. Googe

The fm magnetic tape storage system and associated equipment at the Engine Test Facility of the Arnold Engineering Development Center are discussed in this paper. The requirements of an analog data storage system for wind tunnel testing of propulsion systems are established and the methods by which these requirements are met by the present system discussed. The input instrumentation and recording techniques are described and the reduction of analog data from the storage system with oscillograph, spectrum analyzer and plotting board is described. Samples of data from tests are presented. The relative merits of the fm tape data storage system at the Engine Test Facility are discussed.

A General Purpose Electronic Multiplier—R. A. Meyers

An electronic multiplier for the multiplication of two analog voltages has been designed. Either or both analog voltages may vary arbitrarily with time, and may contain components in the frequency range of from 0 to 150 sinusoidal cycles per second. An instantaneous algebraic summation of the analog voltages, of 200 volts peak amplitude in either plus or minus polarity is accommodated at the input. As presently designed the device has a dynamic range of 500 to 1 for an accuracy of about 1% absolute magnitude. Products as low as 5mv rms can be detected. The sensitivity of the instrument is denoted by the over-all attenuation of 3. The drift of the unit is less than 20.0 mv at the output over an 8 hour period. A pilot model has been built and tested, and for the past year has been in use at the Bureau of Ships, Navy Department, Washington, D. C.

The Application of Analogue Techniques to a Continuously Rotating Magnetic Drum—J. L. Douce and J. C. West

The paper describes how a rapidly revolving magnetic drum can be employed as a versatile device for data handling and analogue computing. In particular it shows how long time, slow varying functions may be stored as they occur and portrayed instantaneously on a C.R.O. with a fast time sweep.

Information is stored on the drum, and can easily be modified, by pulse width modulation of a 10 kc/s pulse train, rendering the accuracy largely independent of drum eccentricity and magnetic properties. The output of any track is repetitive, the repetition rate being the rotational frequency of the drum (approx. 50 c.p.s.). The x deflection of the display tube is a linear time base of the same frequency and gives a repetitive trace of the data recorded on any one track.

Rapid Automatic Digitization and Sorting of Random Graphical Data—V. S. Carson

This paper discloses a means of automatic data abstraction which is rapid and accurate. The method yields many more samples than can be obtained manually. Linear photoelectric spot-scanning is used. Impulses are obtained from both the inked record and the coordinate lines usually printed on such charts. A partial-coincidence method is used to reject signals from the coordinates. The desired signal from the stylus line is injected into its proper count-register by a high speed electronic commutator having a resolving power of one-fiftieth the width of the chart.

A Phase Filter Applied to Spectral Phonocardiography—F. H. Middleton, G. B. Gilbert, W. H. Huggins and G. N. Webb

Spectral phonocardiography is the name applied to the process of analyzing heart sounds to provide a frequency-amplitude-time display. A segment of heart sound is recorded on a magnetic medium and then repeatedly analyzed by a heterodyne filter system. Because the bandwidth of a conventional filter determines both the frequency and the time resolution, a difficulty arises in that the selection of a wideband filter to obtain adequate time resolution leads to a poor frequency resolution and vice-versa.

By using the fact that the phase-frequency characteristic of a filter is changing at resonance whereas its amplitude response is not, it is possible to obtain both good time and frequency resolution at the expense of amplitude information. This paper describes the application of this principle to the design of a phase filter and shows that it produces an analysis having some of the features of the sense of hearing.

The ORDRAT—Ordnance Dial Reader and Translator—P. M. Kintner, R. E. Howard, S. B. Peterson and R. C. Webb

The ordnance Dial Reader and Translator is an electronic data reduction machine designed specifically for the purpose of reading the azimuth and elevation dial images on cinetheodolite films as produced in ballistic measurements. The data are applied to the machine in the form of 35-millimeter film records which may be either the double-frame type produced in the Askania theodolites or single-frame records as produced in the BRL/NGF theodolite. The machine reads double-frame records at the rate of ten dial images per second and single-frame films at the rate of twenty images per second.

The identifying time code, the degree values, and the vernier readings of the dials are processed through the machine, translated into digital form and recorded on six channels of a standard eight-channel digital computer tape. The data are recorded on the tape in proper form for immediate application to a digital computer of the ERA 1103 type. Other types of computers can be readily served.

The Doppler Data Translator—P. M. Kintner and E. J. Armata

Doppler data may be reduced by (1) determining the number of cycles occurring in a fixed time interval, or (2) determining the time interval for a fixed and, in general, an integer number of Doppler cycles. The former is usually preferred by the data user; however, for automatic reduction, method (2) is preferable because of the inherent requirement in (1) that fractional parts of cycles be determined. The method used in the Doppler Data Translator is a compromise between the two methods in that a nominal fixed time interval is established as a reference, with the true measurement interval determined by adding to the reference interval until an integer number of cycles is encompassed. The data user thus is given data based on fairly constant timing intervals, yet the difficult problem of the measurement of fractional parts of cycles is avoided.

High-Speed Analog-Digital Convertors—M. L. Klein

Electronic analog digital conversion techniques resolve themselves into three general classes: feedback trial encoding, time coding, space coding. The latter technique is the fastest and involves a monoscope tube or tube of special design in which the code is designed onto a target or grid configuration. The sweeping of a beam in analog with the input signal produces a pulse code modulation of the signal. This technique requires a quantizer to eliminate ambiguous readings which can introduce an error of half the resolution of the system. Feedback encoding resolves itself into two distinct types. Clock cycled feedback encoding uses, say, binary trial voltages compared according to a fixed program with the input voltage. By suitable logical circuitry, the pulse code is emitted when the trial voltage is less than the residual signal voltage. This technique is presently the most widely used. Self-compensating feedback techniques essentially correct errors as the signal varies beyond the resolution of the system and makes use of an on-line digital-analog technique producing a quantized analog representation of the digital code which is continuously compared with the signal voltage. Time coding is based on pulse width of time gated comparison whereby a disabling gate time is made proportional to the signal voltage. During the enabling gate, a clock count is accumulated which is disabled and finally read.

An Unusual Electronic Analog-Digital Conversion Method—B. D. Smith, Jr.

A method of electronic analog to digital conversion is presented which has not been given extensive treatment in previous literature. The method employs d-c amplifiers and passive elements only in contrast to other methods employing flip-flops, shaft digitizers, relays, etc. It also provides binary digital outputs in parallel form continuously in response to an input analog voltage; that is, no sampling or timing processes are involved. Two arrangements are presented: one which provides a parallel digital representation of an analog voltage in the normal binary code, and the other which provides the parallel digital representation in the reflected binary (Gray) code having certain advantages. The method consists of cascading identical stages, each of which produces one binary digit output and a difference signal to the following stage. An experimental 5-digit coder was constructed which employed only two dual-triodes per stage.

Sine-Cosine Angular Position Encoders—C. P. Spaulding

Sine-Cosine Angular Position Encoders are useful in coupling an analog device to a digital computer. The desired result is increased accuracy of the analog equipment and simplification of the digital equipment. The design of sine-cosine encoders involves compromising a conflict between mechanical simplicity in the encoder and circuit simplicity in the digital computer.

Precision Direct-Reading, Binary-Position Encoders—W. I. Frank and A. B. White

A 16-digit binary, digital, shaft-position encoder and analog-to-digital converter has been developed which reads shaft position to 1 part in 65,000. Cyclical binary code is employed which limits maximum ambiguity to $\pm\frac{1}{2}$ count. In the present model angles can be read as often as 100 times/sec at angular rates from zero to 20,000 counts/sec. Other desirable features of this system are: direct reading with no coarse and fine readers, gears, or discriminators; practically unlimited slewing rates; no brushes; unique readout without counters.

Digitization of Carrier Excited Transducers Using a Programmed Attenuator—J. R. Zweizig

A precision analog-to-digital converter has been developed for use in measuring parameters which appear as carrier output voltages from transducers. A common ac carrier supply source is used for both the transducer and an attenuator so that any change in the excitation voltage does not affect the measurement accuracy. This also eliminates the necessity for a standard voltage source for the attenuator. The time required for digitization is 30 milliseconds, accuracy is one part in 3,000 and the sensitivity is such that a 5 microvolt change in the unknown voltage representing the parameter being measured is easily detected.

Quantization of a Signal Plus Random Noise—G. H. Myers

This paper discusses the problem of quantizing analog information which has been disturbed by random noise. The following system is considered.

Given a signal "S" confined to a definite range, with all values of "S" equally likely over this range. Let "S" be disturbed by Gaussian noise, and then quantize the transmission over a data link. In the absence of noise, no errors are greater than one-half of the width of a quantum step. When noise is present, a certain percentage of the errors are greater than one-half the quantum step width.

The problem is: How large should the quantum step be so that the probability of an error greater than one-half the step width is less than a pre-assigned value?

The error distribution function for this case is derived; the distribution function is graphically integrated to give the desired probability of large error. The final curve is in normalized form and may be used as a design aid.

Mathematical Definitions for Transducer Measure Criteria—L. J. Fogel

This paper suggests some logical definitions, and their mathematical equivalents, for the most common transduction qualities so as to allow the formation of applicable sets of measure criteria. These are not only useful for numeric computation but also provide a means for the comparative evaluation of actual or proposed systems. Some basic properties of all transducers are considered with respect to the statistical aspects of input-output relations. With these presumed or measured, and the purpose of transduction specified, it should be possible to determine a mutually compatible set of criteria. It may include such measures as accuracy, reliability, tolerance, precision, readability, sensitivity, and others. The development of an appropriate weighting function for the specified set of applicable criteria can be used to form a meaningful figure of merit which will provide a single dimensional measure for the value of the transducer in relation to the function it is intended to perform.

New Multi-Purpose Industrial Transducers—W. F. Newbold and J. V. Werme

This paper deals with a line of compatible transducers whose electric output is the analogue of the important industrial measurements of flow, temperature, pressure, and liquid level. The emphasis in these designs has been on accuracy, long life, and ruggedness.

Their use with electric control systems and data handling systems is described. Operating principles of the transducers are given along with some applications illustrating their flexibility.

A Digital Sine-Cosine Transducer—W. Henn and A. S. Robinson

This paper describes a system for deriving the sine and cosine of an angle as parallel digital numbers from either a shaft input or from a pulse rate corresponding to rate of change of angle. The shaft input can be either an analog rotation or a parallel digital number. A specific system intended for the real-time processing of radar data is described in which the sine and cosine are constant during the sweep period and change to their new value during sweep dead time.

The system described in detail accepts an analog shaft rotation and derives the sine and cosine from a number of disks on which appropriate information is encoded. While only 1024 total bits of information are stored, the sine and cosine are both derived with an accuracy of 1 part in 8191. Maximum antenna speed for this unit is 10 rpm. New values of sine and cosine appear in the output registers 20 sec after the start of sweep dead time is indicated.

A New DC Voltage Discriminator with Independent Control of Threshold Voltage and Voltage Differential—N. P. Stucky

The design and operation of a new d-c voltage discriminator is explained. This discriminator was developed for use in a house heating analog computer to stimulate the thermostat, in that application it was designed to trigger at any pre-set threshold voltage between 55 volts and 85 volts. The voltage differential between the absolute (lower) and terminal (upper) threshold is independently variable between 0.1 volts and 7 volts. Both of these parameters can be modified to satisfy specific circuit requirements.

Output of the discriminator is a square wave with a rise time of approximately 0.3 microsecond. A circuit diagram is shown and photographs of oscillographic traces show its characteristics.

A Binary Coded Decimal Converter—Martin Ziserman

The fields of digital data handling and data processing often require that a shaft position be converted to a set of electrical signals denoting a digital quantity. When computations are involved, a weighted number code is desirable. Where automatic printout with visual indication is to be used, a code requiring a minimum amount of translation to a familiar number system (i.e., decimal) is advantageous. A code effecting a simple solution to both problems is found in a weighted 4-bit decimal code.

In particular, this paper shows how a shaft position encoder, utilizing the straightforward 8-4-2-1 decimal code, has been designed. The ambiguities associated with non-progressive codes have been eliminated by a unique configuration of disc pattern and brush position, avoiding the necessity of complex external electronic circuitry.

200-Channel Sequential SADIC System—D. E. Jorgensen

This abstract describes a 200 channel Sequential Data Processing SADIC System, wherein the signal from each strain gage is presented as a visual display, is tabulated on an electric typewriter, and is punched into tape suitable for subsequent introduction into an IBM summary punch. Any or all of the 200 channels can be sampled in any sequence the operator may desire. The signal from each strain gage bridge is individually accomplished without affecting the condition existing in other channels. A multipoint commutator, or input switch, is utilized to provide switching, such that the voltage from each of the large number of strain gages is sampled in turn.

The output from each strain gage, sensed as a low level d-c vòltage, will be amplified and converted to decimal digits by means of the Consolidated Engineering Corporation's SADIC type analog-to-digital converter. Provisions for automatic balance and automatic sensitivity control for each of the gages are provided. The resolution of the input signal will be within one microvolt.

Vehicular Communications

PGVC-6, July, 1956

(Papers Presented at the Sixth Annual Meeting, Portland, Ore.)

The Bonneville Power Administration Land-Mobile Communications System—Max Peckhart and Donald Johannson

Unusual Applications of Mobile Radio Peculiar to the Forest Industries—R. W. Olin

Public benefit of the use of mobile radio in industrial forest work is great. Each year many lives of workmen and visiting public are saved, and huge property damage averted through improved communications by radio. Industrial tree farming programs are accelerated and made practical by mobile radio. These programs will provide wood resources and water reserves to be enjoyed as public benefits for many future generations.

Most timbered areas have no roads or public communications and are frequently in mountainous regions. Mobile radio easily adapted to the circumstances proves to be the only adequate means of communication to administer, protect, and harvest timber crops. Modern machines used in forest work are widely scattered on each logging job but radio coordinates their efforts. It might be compared with military use of radio to coordinate the fire power of many mobile units on a single target.

Forest products radio service may be small in total radio units, but in effective public use per mobile unit this service has established an outstanding national record. Mobile radio is certainly an indispensable tool of industrial forests.

Frequency Management in the Forest Industries Radio Communications—Myron Savage

VHF Propagation Measurements in the Rocky Mountain Region—R. S. Kirby, H. T. Dougherty, and P. L. McQuate

Mobile measurements of vhf propagation over various irregular terrain paths have been made by the National Bureau of Standards in the Colorado Rocky Mountain region in an effort to evaluate terrain effects upon broadcast and point-to-point communications at very high frequencies. Mobile measurements of the varying path transmission loss were obtained in a continuous manner while driving along selected routes with a mobile field strength recording unit, which consists of a modified house trailer equipped with a telescoping mast and pulled by a pick-up truck. The paths used ranged from relatively smooth to very rough.

The results of the measurements are considered in the light of current irregular terrain theory. The correlation of sector median transmission loss for different frequencies over irregular terrain tends to be high when the paths are nearly the same, becoming significantly less when the paths diverge. This would indicate that the frequency selectivity of an irregular terrain path is small.

450 MC Coverage Tests at Chicago—K. V. Glentzer

Extending Mobile Radio Range by VHF Repeaters—C. A. Kemp

The distance from a base station over which a mobile radio may operate is usually limited to near the optical horizon, particularly with frequencies of 150 mc and higher. When it is desired to extend the mobile range beyond this distance remotely controlled base stations involving either telephone lines or microwave relay systems are used quite successfully. However, these systems are expensive, particularly where the cost of the microwave system or telephone line cannot be shared by other functions such as additional telephone channels or telemetering and control channels. By the judicious use of mountain tops, where available, and the use of unattended automatic repeater (relay) stations, it is possible to extend mobile range from a base station to several times that which is obtained with a single base station without incurring the expense or complexity involved in a microwave relay system. This paper will describe some of the methods which have been used by the Bureau of Reclamation to provide this type of long distance base to mobile coverage.

The Occurrence of E_2 and F_2 Skip in the 30–50 MC Mobile Band—E. W. Allen

A High Performance Mobile Unit for 450 Megacycles—M. A. Robbins and George Ayer

A simplified transmitter employs a magnetic reactor-modulator and a frequency tripling power stage.

The receiver's sensitivity is twice the accepted value of 1 microvolt. This is made possible by the use of a planar-grid triode input stage.

D. C. Transformers—J. S. Smith

The introduction of power junction transistors as switch devices has opened new horizons to mobile power supply design. The use of these devices in conventional circuits presently offers equipment capable, within the temperature limits of the transistors, of out-performing present vibrator supplies at low power levels. As power transistors are improved wide application of this principle of *dc* to *dc* conversion is inevitable.

AM Systems for 1955—Ray Morrow

A re-evaluation of vehicular communications in the light of narrowing bandpass restriction points to amplitude modulation equipment. AM equipment provides full intelligence, a favorable signal to noise ratio and excellent squelch action with only a 6 kc bandwidth. It also retains extreme sensitivity and frequency stability. All this is packaged in a compact but easily accessible unit for convenience in servicing. There are excellent systems in existence, using this type of modern AM equipment which attest to the effectiveness of present day AM communications.

The Integration of Municipal Radio Systems—M. E. Kennedy

VHF Marine Mobile Systems in British Columbia—M. E. Green

The purpose of this paper is to give a general description of the vhf maritime mobile facilities of the North-West Telephone Company, who provide this type of service to vessels operating in British Columbia coastal waters.

The present four-channel system, which has been in operation since 1948, will be described. A proposed six-channel system, based on assignments in the 152 to 162 mc/s band in accordance with the 1952 International Agreement, will be outlined.

Vehicular Communications in the Petroleum Industry—R. L. Ransome

A brief history of the development and growth of petroleum industry use of radio is given and general applications of mobile radio described. A number of the more pressing and immediate problems common to most mobile services are outlined and several steps toward possible solutions suggested.

Spectrum Compression and its Problems—Curt Schultz

Abstracts and References

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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.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Index to the Abstracts and References published in the PROC. IRE from February, 1955 through January, 1956 is published by the PROC. IRE, June, 1956, Part II. It is also published by *Wireless Engineer* and included in the March, 1956 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534-16:061.3 2271

Conference on "Sound and Vibrations in Solid Bodies," Gottingen, 19th-22nd April, 1955—(*Akust. Beihefte*, no. 1, pp. 49-227; 1956.) The text is given of 35 papers, the majority in German and the remainder in English; English, French, and German summaries are provided.

534.121.1 2272

On the Fexural Vibrations of Circular and Elliptical Plates—W. R. Callahan. (*Quart. Appl. Math.*, vol. 13, pp. 371-380; January, 1956.) The frequency equations for the normal modes of vibration are studied.

534.2-8 2273

A Simple Method for the Visualization of Ultrasonic Fields—Y. Torikai and K. Negishi. (*J. Phys. Soc. Japan*, vol. 10, p. 1110; December, 1955.) A method using ordinary photographic paper is outlined.

534.2-8-14 2274

The Propagation of Ultrasonics in Suspensions of Particles in a Liquid—J. Busby and E. G. Richardson. (*Proc. Phys. Soc.*, vol. 69, pp. 193-202; February 1, 1956.) Measurements at frequencies between 1 and 10 mc are reported, on suspensions of glass spheres (monodisperse) and silica particles (polydisperse) of radius 8-100 μ.

534.6 2275

An Apparatus for measuring Air-Flow Resistance of Acoustical Materials—H. J. Sabine. (*ASTM Bull.*, no. 211, pp. 29-32; January, 1956.)

621.395.616 2276

Artificial Stabilization of the MR-103 Type Condenser Microphone—T. Hayasaka, K. Masuzawa, and M. Suzuki. (*Rep. Elect. Commun. Lab., Japan*, vol. 3, pp. 59-60; October, 1955.) Titanium is used as diaphragm material on account of its strength. To obtain the same stability as would be provided by aging for a year or more at normal temperature, it is only necessary to heat the microphone for 5 hours at 200°C.

621.395.623.7 2277

Miniature Loudspeakers for Personal Radio Receivers—J. C. Bleazey, J. Preston, and E. G. May. (*RCA Rev.*, vol. 17, pp. 57-67; March, 1956.) Two experimental loudspeakers are described in which the cone housing and the magnet occupy the same space, thus reducing the over-all depth. Directional characteristics, distortion, and frequency response are similar to those of conventionally constructed loudspeakers.

621.395.623.8:621.396.975 2278

Wireless Sound Systems—C. W. Hargens. (*J. Franklin Inst.*, vol. 260, pp. 351-356; November, 1955.) Description of a system installed at the Franklin Institute in a hall with a reverberation time of several seconds. Miniature transistor-equipped receivers are used. The carrier frequency is between 550 and 1600 kc, the particular value being chosen to meet a Federal Communications Commission requirement regarding field strength at distance $\lambda/2$. Any public-address amplifier can serve as modulator. The transmitter antenna is a long conductor making a loop round the hall; the receiver antennas are ferrite-core types.

621.395.625.2:621.396.712.3 2279

Reproducing Equipment for Fine-Groove Records—G. V. Buckley, W. R. Hawkins, H. J. Houlgate, and J. N. B. Percy. (*BBC Engng. Div. Monographs*, no. 5, pp. 1-19; February, 1956.) Description of a reproducing desk designed to facilitate the location of desired excerpts.

621.395.625.3 2280

Mechanical Aspects of Magnetic-Recorder Design—G. P. Bakos. (*Tijdschr. Ned. Radiogenoot.*, vol. 21, pp. 17-37; January, 1956. In English.) A review of modern practice, covering tape, sheet, and disk machines, as well as multichannel equipment.

ANTENNAS AND TRANSMISSION LINES

621.315.212.1.011.3 2281

The Inductance of Two Elliptic Tubes—E. E. Jones. (*Brit. J. Appl. Phys.*, vol. 7, pp. 56-58; February, 1956.) The inductance is calculated of cables comprising two confocal or concentric elliptic tubes of nonmagnetic material.

621.372 2282

An Investigation of the Properties of Radial Cylindrical Surface Waves launched over Flat Reactive Surfaces—W. M. G. Fernando and H. E. M. Barlow. (*Proc. IEE*, part B, vol. 103, pp. 307-318; May, 1956.) Experiments were made using a vertical dipole source arranged at various heights above the center of a horizontal metal disk inductively loaded either by means of a thin dielectric coating or by forming a series of concentric corrugations. Theory developed by Cullen (22 of 1955) is applied; the observations are in good agreement with the theoretical predictions both as regards field distributions and launching efficiency, and it is confirmed that a launching efficiency approaching 80 per cent is attained for a particular height of the dipole.

621.372.2:621.317.34:621.317.729 2283

An Investigation into some Fundamental Properties of Strip Transmission Lines with the Aid of an Electrolytic Tank—J. M. C. Dukes. (*Proc. IEE*, part B, vol. 103, pp. 319-333; May, 1956.) A technique was developed by means of which the line parameters could be rapidly evaluated with a useful degree of accuracy for a range of dimensions for which rigorous solutions by direct analysis are not readily available, this range including lines with characteristic impedance between 20 and 150 Ω. The investigations covered the balanced-parallel-plate line, the strip-above-ground line, and the sandwich, or triplate, line. The validity of formulas derived by other workers is discussed and new methods for calculating the line parameters are developed. The results indicate that the dominant mode in a microstrip line is closer to the TEM mode than has been supposed hitherto. The sandwich line has some theoretical advantages over the strip-above-ground line, but these may be offset by practical disadvantages.

621.372.2:621.385.029.6 2284

Interpretation of Wavelength Measurements on Tape Helices—C. P. Allen and G. M. Clarke. (*Proc. IEE*, part C, vol. 103, pp. 171-176; March, 1956.) An unexpected region of dispersion observed by Sensiper (1247 of 1955) is found to be due to the finite length of the helix. The effect may be important in relation to the design of traveling-wave devices.

621.372.8 2285

A New Treatment of Lossy Periodic Waveguides—P. N. Butcher. (*Proc. IEE*, Part B, vol. 103, pp. 301-306; May, 1956.) The new treatment is based on introduction of a "complex Q factor," Q_c; the propagation coefficient of a mode in a lossy guide at the frequency ω is equal to that of the corresponding mode in a loss-free guide at the frequency $\omega(1-j/2Q_c)$. An explicit formula is given for Q_c for the case of small losses.

621.372.8 2286

Junction Admittance between Waveguides

of **Arbitrary Cross-Sections**—E. D. Farmer. (*Proc. IEE*, part C, vol. 103, pp. 145–152; March, 1956.) If the dominant modes of the adjoining waveguides have similar patterns over the coupling aperture, the junction can be represented approximately by a two-terminal network. A general definition of characteristic impedance is introduced enabling the junction to be regarded as an "impedance mismatch" together with a shunt susceptance "junction effect." The limits of applicability of the theory are assessed by making calculations for some special cases, including a junction between a rectangular and a hexagonal guide.

621.372.8 2287
Microwave Propagation in Anisotropic Waveguides—A. E. Karbowiak (*Proc. IEE*, part C, vol. 103, pp. 139–144; March, 1956.) Analysis is presented based on the surface-impedance approach developed previously (3484 of 1955). The corrugated surface and the conducting helix are two particular cases of the anisotropic surface considered. All E and H_0 modes are shown to be stable whatever the orientation of the principal axes of the surfaces; higher-order H modes are unstable unless the principal axes coincide with the coordinate axes of the surface, but a certain combination of H waves ("spinning H wave") can be propagated.

621.372.8:538.221 2288
Attenuation and Permeability of Ferromagnetic Waveguides between 9000 and 9675 Mc/s—J. Allison and F. A. Benson. (*Proc. IEE*, part C, vol. 103, pp. 205–211; March, 1956.) "Measurements of the attenuations produced by air-filled rectangular waveguides of nickel, mild steel, mumetal, radiometal, and rhometal have been made in the frequency range 9000–9675 mc. The permeabilities of the materials have been determined from these measurements and a knowledge of the roughness and resistivity of each waveguide internal surface. The effects of temperature on the hf permeabilities have also been studied, and some qualitative results are included on the effect of superimposing a steady magnetic field on the hf one."

621.372.8:538.221:538.6 2289
Equation of Circularly Polarized Waves in a Gyroparamagnetic Medium—J. Soutif-Guicherd. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1418–1421; March, 1956.) Analysis is outlined for propagation in a waveguide containing a medium whose permeability is represented by a tensor. The propagation coefficient is expressed in a form susceptible to limited development for the case of a paramagnetic medium; the field equations for a circularly polarized wave are derived.

621.372.8:621.3.012.8 2290
The Calculation of the Equivalent Circuit of an Axially Unsymmetrical Waveguide Junction—R. E. Collin and J. Brown. (*Proc. IEE*, part C, vol. 103, pp. 121–128; March, 1956.)

621.372.8:621.318.134 2291
Temperature Behavior of Ferrimagnetic Resonance in Ferrites located in Waveguide—B. J. Duncan and L. Swern. (*J. Appl. Phys.*, vol. 27, pp. 209–215; March, 1956.) The investigation described is an extension of that reported previously (2192 of 1955). Measurements were made on MgMn and NiZn ferrites over the temperature range from 25°C. to the Curie point in each case. Temperature variation of the resonance-line width and the apparent gyromagnetic ratio was observed. The effect on the microwave transmission properties is discussed.

621.372.8:621.372.2 2292
The Excitation and Propagation of E_{0n} Modes in a Circular Waveguide with Coaxial Lines at Input and Output—A. Sander. (*Arch. Elekt. Übertragung*, vol. 10, pp. 77–85; March, 1956.) Analysis is presented in which the concepts of "field" and "hybrid" quadripoles and the corresponding matrices are introduced.

621.372.8+621.396.677.85]:621.372.43 2293
The Design of Quarter-Wave Matching Layers for Dielectric Surfaces—R. E. Collin and J. Brown. (*Proc. IEE*, part C, vol. 103, pp. 153–158; March, 1956.) "A quarter-wave transformer to match the junction between an empty waveguide and one completely filled with a dielectric may be made from a waveguide partially filled with dielectric. A method of designing such a transformer, when all the waveguides have the same cross section, is described, and experimental results are given to show that this design is satisfactory. A similar arrangement can be used to match the surfaces of a dielectric lens: slots are cut on the surface and design information is given for slots parallel or perpendicular to the electric field of the wave incident on the surface. Measured reflection coefficients for a surface matched in this way are in good agreement with calculated values."

621.396.67:001.4 2294
Russian Antenna Terminology—G. F. Schultz. (PROC. IRE, vol. 44, pp. 692; May, 1956.) A short list of representative terms is given with the English equivalents.

621.396.674.3:621.397 2295
Wide-Band Television Aerials—M. G. O'Leary. (*Wireless World*, vol. 62, pp. 288–291; June, 1956.) An illustrated review of current North American practice in the design of combined antenna systems for reception in bands I and III.

621.396.676.012.12 2296
The Radiation Pattern of an Antenna mounted on a Surface of Large Radius of Curvature—J. R. Wait. (PROC. IRE, vol. 44, p. 694; May, 1956.) Calculations are made of the radiation pattern of a dipole or a slot on a conducting sphere of large radius, by applying van der Pol-Bremmer theory.

621.396.677.71:621.397.61 2297
The Omniguide Antenna—an Omnidirectional Waveguide Array for U.H.F.-Television Broadcasting—O. M. Woodward, Jr. and J. Gibson. (*RCA Rev.*, vol. 17, pp. 13–36; March, 1956.) Detailed description of an antenna comprising an octagonal-section inner waveguide surrounded by eight ridged waveguides with offset longitudinal slots. The picture and sound signals are diplexed, using either a combining filter and single feed line or a separate coaxial-line sound input with a special diplexer. The construction is of aluminum with a thin covering of fiber glass for weather protection.

621.396.677.85 2298
Successive Approximation and Expansion Methods in the Numerical Design of Microwave Dielectric Lenses—R. L. Sternberg. (*J. Math. Phys.*, vol. 34, pp. 209–235; January, 1956.)

AUTOMATIC COMPUTERS

681.142 2299
The Short Electronic Analogue Computer—R. J. A. Paul. (*Overseas Engr.*, vol. 29, pp. 205–208; January, and pp. 251–252; February, 1956.) Description of the design and operation of a general-purpose computer designed for quantity production and capable of single-shot and repetitive operation.

681.142 2300
Tridac, a Large Analogue Computing Machine—F. R. J. Spearman, J. J. Gait, A. V. Hemingway, and R. W. Hynes. (*Proc. IEE*, part B, vol. 103, pp. 375–390; May, 1956. Discussion, pp. 390–395.) A detailed description; a shorter account was abstracted previously (943 of 1955).

681.142 2301
Function Generators based on Linear Interpolation with Applications to Analogue Computing—E. G. C. Burt and O. H. Lange. (*Proc. IEE*, part C, vol. 103, pp. 51–58; March, 1956.) By using suitable combinations of diode circuits and high-gain feedback amplifiers it is possible to generate functions without restriction to monotonic characteristics. Experimental results are presented for a sin x generator in which the error is about about ½ per cent of the maximum output.

681.142 2302
Construction and Method of Operation of Modern Integrating Equipment [differential analysers]—H. Hoffmann. (*Elektrotech. Z., Edn A*, vol. 77, pp. 41–52; January 11, and pp. 77–83; February, 1956.) General principles are discussed and a detailed description is given of an installation in Germany. The results are presented to three or four significant figures by counter mechanisms, and in the form of curves on function benches. Curves of empirical functions are dealt with by photoelectric scanning. Errors do not exceed 0.1 per cent–1 per cent.

681.142:[621.314.63+621.314.7 2303
Engineering Multistage Diode Logic Circuits—B. J. Yokelson and W. Ulrich. (*Elect. Engng., N. Y.*, vol. 74, p. 1079; December, 1955.) Design of computers using crystal diodes and transistors is discussed. Transistor input circuits directly coupled to multistage logic circuits may avoid the need for intermediate amplifier stages.

681.142:621.314.7 2304
Transistor Circuits for Analog and Digital Systems—F. H. Blecher. (*Bell Syst. Tech. J.*, vol. 35, pp. 295–332; March, 1956.) A summing amplifier, an integrator, and a voltage comparator using junction transistors are described, together with a voltage encoder made up from them, for translating voltages into equivalent time intervals for analog-to-digital conversion.

681.142:621.314.7 2305
An Experimental Transistorized Calculator—G. D. Bruce and J. C. Logue. (*Elect. Engng., N. Y.*, vol. 74, pp. 1044–1048; December, 1955.) The machine is functionally identical with the IBM Type-604 calculating punch, but the tubes are completely replaced by transistors and Ge diodes.

681.142:621.37 2306
An A.M.-A.M. Multiplier—L. Lukaszewicz. (*Bull. Acad. Polon. Sci., Classe* 4, vol. 3, pp. 145–148; 1955. In English.) A relatively simple purely electronic multiplier circuit for differential analyzers is described. Working with an upper frequency limit of 10 kc for both factors, accuracy is within about 0.3 per cent of full scale.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011 2307
Abac of the Function $2J_1(z)/zJ_0(z)$ for studying the Initial Complex Permeability of Circular-Cross-Section Conductors at High Frequency—J. Benoit and E. Naschke. (*J. Phys. Radium*, vol. 17, pp. 77–78; January, 1956.) The impedance of the conductor is measured and the parameter z is then found from the abac, using a known formula; the complex permeability μ is then derived using a given relation between μ and z. This abac supplements that presented by Prache (1875 of 1950), which was not applicable to good conductors.

621.3.011:621.396.822 2308

Physical Sources of Noise—J. R. Pierce. (PROC. IRE, vol. 44, pp. 601–608; May, 1956.) The principal types of electrical noise encountered in circuits and tubes are identified and the mechanisms giving rise to them are discussed. Equations used to represent noise phenomena are derived.

621.3.011.21:621.375.13 2309

The Impedance Concept—G. C. Mayo and J. W. Head. (*Wireless Engr.*, vol. 33, pp. 96–102; April, and pp. 121–128; May, 1956.) Impedances and associated transfer functions are expressed in terms of a variable $p=\alpha+j\omega$ which is closely associated with time differentiation. The fundamental properties of the "p world" are discussed. A general condition for an algebraic equation to be free from roots with positive real parts is obtained; a network whose characteristic equation has this property is stable. Conditions are deduced for a system to have a damping rate at least equal to a specified value. The effect of adding terms in p^4 and p^5 on the roots of a given cubic in p is examined. The effect on the gain and maximum obtainable feedback of the addition of a "step circuit" to a three-stage RC amplifier is considered in detail.

621.3.066.6:621.318.5 2310

Properties and Comparative Tests on Relay Contacts—T. Gerber. (*Tech. Mitt. Schweiz. Telegr. TelephVerw.*, vol. 34, pp. 1–26; January 1, 1956. In French.) See 2545 of 1955.

621.314.222:621.397.6 2311

Toroidal Transformers pass Video Bandwidths—G. W. Gray. (*Electronics*, vol. 29, pp. 150–153; May, 1956.) Television video transformers are wound on supermalloy tape-wound toroids having a low-frequency permeability of 70,000 diminishing with increasing frequency; a bandwidth of 6 mc is obtainable. These transformers may be used for matching a 50-Ω coaxial cable and for interstage coupling in transistor video amplifiers.

621.314.222.012.3 2312

[Power-] Transformer Design Chart—R. Lee and N. E. Mullinix. (IRE TRANS., vol. CP-3, pp. 10–14; April, 1955. *Electronics*, vol. 29, pp. 184–186; April, 1956.) The chart gives data for designing two-winding 60-cps lv transformers.

621.316.8.029.6:621.315.212:621.372.22 2313

The Theory and Design of Coaxial Resistor Mounts for the Frequency Band 9–4000 Mcs—I. A. Harris. (PROC. *IEE*, part C, vol. 103, pp. 1–10; March, 1956.) A design is described in which the resistive inner conductor has uniform diameter while the outer conductor has a tractrix profile. Lead-in cones are designed to avoid discontinuity at the connections. Experimental results indicate that the impedance is within 1 per cent of the dc resistance, with an extremely small phase angle, at all frequencies up to the highest measured, namely 3.45 kmc.

621.316.86:546.281.26 2314

The Operating Mechanism of Voltage-Dependent Silicon-Carbide Resistors—K. Zückler. (*Z. Angew. Phys.*, vol. 8, pp. 34–40; January, 1956.) Measurements on aggregates of SiC particles, such as pressed powder, show how thermal effects depend on grain size and the duration of the current pulse. Thermal and field effects can be separated by reference to resistance/voltage characteristics at different temperatures for single contacts between crystals or between a crystal and a metal knife-edge.

621.316.86:621.3.012.8 2315

The Specification of the Properties of the Thermistor as a Circuit Element in Very-Low-Frequency Systems—C. J. N. Candy. (*Proc. IEE*, part B, vol. 103, pp. 398–409; May, 1956.) "An analysis based on sinusoidal applied voltages shows that a conductor whose resistance is a function of temperature may be represented by an equivalent circuit having a semicircular impedance locus. An expression for the distortion of the waveform is also obtained, and this is found to be small provided that the alternating current is less than a quarter of the steady polarizing current which flows in the conductor. The impedance loci of a bead-type thermistor are plotted by means of a null technique. A typical impedance varied from a negative resistance of 3000 ohms at very low frequencies to a pure inductance of 2000 H at 0.3 cps and then to a positive resistance of 5000 ohms at high frequencies. The use of the equivalent circuit is illustrated by designing phase-shift networks suitable for use in the stabilizing of very-low-frequency control systems. These circuits may be used in systems where either ac or dc data transmission is employed."

621.316.86.002.2 2316

Problems Encountered and Procedures for obtaining Short-Term Life Ratings on Resistors—W. T. Sackett, Jr., (IRE TRANS., vol. CP-3, pp. 15–29; April, 1955.) Account of investigations made at the Battelle Memorial Institute on the extent to which length of life must be sacrificed when composition resistors are operated at high temperature. A machine system for handling the data is described.

621.318.435.3:621.316.722 2317

A.C. Controlled Transductors—A. G. Milnes and T. S. Law. (*Proc. IEE*, part C, vol. 103, pp. 81–94; March, 1956.) Analysis of the behavior of the single-core auto-self-excited transductor is made for the condition when the control circuit has finite resistance. AC control of fullwave transductors and some push-pull circuits with half-wave and full-wave outputs are referred to.

621.318.435.3.011.6 2318

The Residual Time-Constant of Self-Saturating (Auto-Excited) Transductors—U. Krabbe. (*Proc. IEE*, part C, vol. 103, pp. 71–80; March, 1956.) Theoretical and experimental evidence is presented indicating that the main winding of a self-saturating transductor influences the time constant for small signals to an extent dependent on the blocking intervals of the rectifiers and on the main-winding resistance. The blocking interval is in turn dependent on the output amplitude; this is consistent with the normal experience that transductor response becomes faster as the output increases.

621.318.57:621.374.32:621.387 2319

A Digital Differential—W. H. P. Leslie. (*Electronic Engng.*, vol. 28, pp. 190–193; May, 1956.) A simultaneous bidirectional counter circuit is described in which a dekatron tube is used to indicate the running difference in count between two independent pulse trains; the circuit can also be used in frequency- and speed-control applications.

621.318.57:621.397.61 2320

Electronic Switches for Television—Spooner. (See 2551.)

621.319.4 2321

The Effective Leakage Resistance of Several Types of Capacitors—R. W. Tucker and S. D. Breskend. (IRE TRANS., vol. CP-3 pp. 3–9; April, 1955.) A rapid method of measuring leakage resistance based on rate of charge is described. Variation of leakage resistance with time was measured for commercial capacitors of values ranging from 0.001 to 0.033 μF, with various dielectrics, at temperatures ranging from 73° to 212° F in most cases. The best dc properties were exhibited by a capacitor having a polytetrafluoroethylene dielectric.

621.319.4:621.314.63 2322

A Variable-Capacitance Germanium Junction Diode for U.H.F.—Giacoletto and O'Connell. (See 2559.)

621.319.4:621.372.542.2:621.318.134 2323

Cascaded Feedthrough Capacitors—H. M. Schlicke. (PROC. IRE, vol. 44, pp. 686–691; May, 1956.) By interspersing lossy ferrite washers between the stacked ceramic disks of capacitors of the type described previously (1279 of 1955), the filtering properties for vhf and uhf are greatly improved; such constructions are thus useful for miniaturized low-pass filters. Properties of suitable ferrites are discussed.

621.37:39(083.74) 2324

National Bureau of Standards Preferred Circuits Program—J. H. Muncy. (*Elect. Engng.*, *N. Y.*, vol. 74, pp. 1088–1090; December, 1955.) See 342 of 1956.

621.372 2325

Invariance and Mutual Relations of Electrical Network Determinants—I. Cederbaum. (*J. Math. Phys.*, vol. 34, pp. 236–244; January, 1956.) Work by earlier authors, *e.g.* Tsang (48 of 1955), is generalized. The basic values of the impedance and admittance determinants are connected by a simple relation involving the determinant of the branch parameter matrix.

621.372 2326

Nonlinear Network Problems—G. Birkhoff and J. B. Diaz. (*Quart. Appl. Math.*, vol. 13, pp. 431–443; January, 1956.) General analysis is presented for flow problems; a number of theorems are proved. Relaxation methods are used.

621.372:[621.385.3+621.314.7 2327

Transformation of the Matrices of Generalized Admittances (Impedances) for Various Triode Connections—E. I. Adirovich. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1436–1443; August, 1955.) The usual three sets of connections for triode tubes and for transistors are examined.

621.372.4/.5:621.3.011 2328

The Correlation between Decay Time and Amplitude Response—S. Demczynski. (*Proc. IEE*, part C, vol. 103, pp. 64–70; March, 1956.) An investigation is made of the relation between a) the decay time and the delay time of the indicial response, and b) the bandwidth and peak values of the steady-state amplitude response, for various minimum-phase lumped-parameter networks. Formulas are derived expressing the functional relation between decay time and the ratio f_3/f_6, where f_3 is the bandwidth at -3db and f_6 that at -6db. A formula for delay time is derived which is valid for multistage circuits.

621.372.41:621.3.015.3 2329

Energy Considerations for Growth and Decay Transients in a Simple Resonator Circuit—G. Čremošnik. (*Arch. Elekt. Übertragung*, vol. 10, pp. 65–72; February, 1956.) Analysis is presented for a series RLC circuit, values of current, voltage, etc., benig determined in terms of the characteristic resistance $K=\sqrt{L/C}$. The solution of the differential equations is simplified by taking the energy of the circuit as the basic time variable, since this is fixed by the initial conditions.

621.372.412 2330

40–50-Mc/s Overtone Quartz Crystal Units—K. Takahara, M. Kobayashi, I. Ida, and Y. Arai. (*Rep. Elect. Commun., Lab., Japan*, vol. 3, pp. 46–50; October, 1955.) The experimental crystal described is a circular plate carrying an evaporated metal film, held at diametrally opposite points by springs attached to the lead terminals. Details are given of the lapping process and the frequency adjustment.

621.372.5 2331

The Most Elementary Geometrical Representation of Loss-Free Linear Quadripoles—J. de Buhr. (*Nachrichtentech. Z.*, vol. 9, pp. 80–84; February, 1956.) "It is possible to describe impedance transformations in lossless and linear quadripoles by geometrical quantities and by cascade parameters. An unambiguous geometrical representation in the form of a system of two transformation lines equivalent to the one transformation line of an impedance transformation is also possible and this leads to a new and unified method of representing the three different transformations by linear quadripoles and by elliptical and hyperbolic quadripoles as well as by the parabolic reactance quadripole. This gives for the linear quadripoles an intuitive and elementary form of treatment which has proved to be very useful for the solutions of many quadripole problems. An impedance transformation using the images with respect to two transformation lines is given for the example of a parabolic reactance quadripole such as a series capacity."

621.372.542.4:621.396.41 2332

Interference Spectra and Aerial Filters A Note on the Problem of the Simultaneous [two-way] Operation of Directional Radio Systems with Pulse Modulation—A. Käch. (*Nachrichtentech. Z.*, vol. 9, pp. 63–69; February, 1956.) The design of continuously tunable multicircuit diplexing filters for uhf communication systems is discussed. A frequency separation of 75 mc is assumed between the associated transmitter and receiver, and the asymmetrical nature of the transmitter spectrum is taken into account.

621.373:621.316.729 2333

Theory of Synchronization of Self-Oscillations of Arbitrary Form—I. I. Minakova and K. F. Teodorchik. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 658–660; February 1, 1956. In Russian.) Analysis is given for an oscillation containing only a few harmonic components. The effect of applying a sinusoidal force, of a frequency near the third harmonic, is investigated by a method involving the use of Fourier-series coefficients.

621.373.42:621.316.729 2334

Discrimination of a Synchronized Oscillator against Interfering Tones and Noise—D. G. Tucker and G. G. Jamieson. (*Proc. IEE*, part C, vol. 103, pp. 129–138; March, 1956.) The discrimination exercised by a nonlinear regenerative tuned circuit against unwanted signals accompanying a synchronizing tone is due partly to the frequency response of the system and partly to the nonlinearity, provided that the synchronizing signal has a greater amplitude than the unwanted signals after allowing for frequency response. The phenomenon is analyzed, and measurements of discrimination against noise are reported; the degree of discrimination can be very great when the synchronizing frequency is very close to the natural frequency of the circuit. When the synchronizing tone is absent or is not dominant there is no reduction of the interference intensity or bandwidth.

621.373.421 2335

A Wide-Range RC Phase-Shift Oscillator—W. Fraser. (*Electronic Engng.*, vol. 28, pp. 200–202; May, 1956.) An oscillator with a frequency range from below 1 cps to over 100 kc is described; dc coupling is used. A polyphase version is also described.

621.373.421+621.375.23]:621.385.3.029.6 2336

A Grounded-Grid Valve System with High Stability Characteristics—Exley and Young. (See 2588.)

621.373.421.11 2337

Frequency Stability of LC Oscillators with Large Grid and Anode Capacitances—J. Groszkowski. (*Bull. Acad. Polon. Sci., Classe 4*, vol. 3, pp. 149–155; 1955. In English.) General analysis is given for the Clapp circuit. Various factors affecting the frequency, and the optimum distribution of instability components among the circuit elements, and supply voltages are discussed. See also 3170 of 1954 (Clapp).

621.373.43/.44 2338

The Generation and Application of Rectangular Pulses—R. S. Sidorowicz. (*A.T.E. J.*, vol. 12, pp. 23–42; January, 1956.) A survey of various known techniques, covering both relaxation oscillators and monostable circuits. A delay-line pulse generator, a free-running multivibrator, and a cathode-coupled monostable multivibrator are discussed in detail. The suitability of a tube for switching applications can be estimated on the basis of a figure of merit given by the ratio between the mutual conductance g_m and the total interelectrode and stray capacitance. Transition times are about 30 mμs for circuits using high-g_m double triodes or pentodes and of the order of 100 mμs for circuits with low-g_m tubes. 49 references.

621.373.431 2339

Study of a Flip-Flop with Four Positions of Equilibrium by the Methods of Topological Analysis—L. Sideriades. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1583–1586; March 19, and pp. 1704–1707; March 26, 1956.) Analysis is presented in general terms for a two-stage circuit, assuming square-law tube characteristics. Three possible states identified are a) static, b) dynamic, and c) impulse. The Eccles-Jordan and multivibrator circuits are particular cases. An experimental circuit has been designed; its performance agreed well with the predictions.

621.373.431.1:621.314.7 2340

Multivibrator Circuits using Junction Transistors—A. E. Jackets. (*Electronic Engng.*, vol. 28, pp. 184–189; May, 1956.) Conventional circuits operating predictably at frequencies up to at least 10 kc are described.

621.373.44:621.387 2341

Reduction of the Minimum Striking Voltage of Hydrogen Thyratrons—A. E. Barrington. (*Electronic Engng.*, vol. 28, p. 219; May, 1956.) An auxiliary tripping circuit is described by means of which the output voltage of a line-type thyratron pulse generator is made continuously variable from 0–20 kv.

621.373.52:621.398 2342

A Temperature-Stable Transistor V.C.O. [voltage-controlled oscillator]—F. M. Riddle. (IRE TRANS., vol. RTRC-2, pp. 11–15; November, 1954. Abstract, PROC. IRE, vol. 43, p. 514; April, 1955.) An oscillator for telemetry purposes is described.

621.375.225.029.3 2343

Cascode A. F. Amplifier—L. B. Hedge. (*Wireless World*, vol. 62, pp. 283–287; June, 1956.) A cathode-coupled phase inverter circuit using cascode-connected twin triodes is described, as part of a high-fidelity of an amplifier which does not require a specially designed output transformer.

621.375.3.012 2344

Analysis of a Differential Magnetic Amplifier with Flux Reset Control—C. A. Belsterling. (*J. Franklin Inst.*, vol. 260, pp. 485–505; December, 1955.)

GENERAL PHYSICS

53.05 2345

The Smoothing of Non-formulated Experimental Laws by an Averaging Operation involving No Spurious Deviations—P. Vernotte. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1697–1699; March 26, 1956.) A section of an experimental curve comprising M points is assimilated to a second-degree polynomial and the required ordinates are hence calculated without introducing spurious undulations.

530.152.15 2346

A Simplified Mathematical Approach to Hysteresis Losses—H. L. Armstrong. (*Elect. Engng., N. Y.*, vol. 74, p. 1060; December, 1955.) The hysteresis loop such as a normal B/H curve is approximated by an ellipse, features resulting from harmonics (*i.e.*, nonlinearities) being neglected; calculations are thus simplified.

535.343.4 2347

The Absorption Spectrum of Nitric Oxide in the Far Ultraviolet—J. Granier and N. Astoin. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1431–1433; March 12, 1956.)

535.42 2348

On Asymptotic Series for Functions occurring in the Theory of Diffraction of Waves by Wedges—F. Oberhettinger. (*J. Math. Phys.*, vol. 34, pp. 245–255; January, 1956.) An asymptotic expansion of the integral expression for diffraction by a wedge, at distances large compared with λ, is obtained as the sum of a Fresnel integral, as leading term, and an asymptotic series of inverse powers of the distance; for small values of the distance the expansion takes the form of a series involving Bessel functions.

537/538 2349

Multipotentials of Multipoles—P. de Belatini. (*Bull. Tech. Univ. Istanbul*, vol. 8, pp. 57–74; 1955. In English.) General theory is presented, applicable equally to magnetic or dielectric multipoles. Every multipole is shown to be associated with a scalar or vector multipotential which has spherical symmetry; from the equation of this multipotential the usual scalar potential can be obtained by successive derivations.

537.21 2350

The Electrostatic Centre of a Conductor—R. Cade and D. O. Vickers. (*Proc. Phys. Soc.*, vol. 69, pp. 175–179; February 1, 1956.)

537.221:537.533 2351

Variation of Volta Potential [work function] with Temperature—G. C. Mönch. (*Z. Phys.*, vol. 144, pp. 263–268; January 17, 1956.) Experimental results show that the measured work functions of Ag_2S, AgI and Cu_2O depend on the experimental conditions rather than on the electron concentration or structural phase changes. This was investigated in greater detail by Böttger (2352 below).

537.221:537.533 2352

Investigation of the Temperature Dependence of the Electron Work Function of Metals and Semiconductors—O. Böttger. (*Z. Phys.*, vol. 144, pp. 269–295; January 17, 1956.) The variation of the work function of copper and nickel sheets was determined from the I/V characteristics of a special diode with a tungsten filament, as a function of time, temperature (100°–300°K) and degree of vacuum. The temperature dependence was also investigated for p-type Cu_2O, p-type NiO and n-type CuO. Results, presented graphically, indicate that for work-function measurements on metals a vacuum better than 10^{-8} Torr is required.

537.228.2 2353

On the Molecular Theory of Electrostriction—B. K. P. Scaife. (*Proc. Phys. Soc.*, vol. 69, pp. 153–160; February 1, 1956.) The case of a sphere of fluid dielectric subjected to a uniform external field is discussed from both macroscopic and microscopic points of view

That part of the spur of the stress tensor $3\Delta P$ which depends quadratically on the applied field is calculated for both nonpolar and dipolar dielectrics; the Lorenz-Lorentz and Debye theories lead to the result that ΔP is zero, but more modern theories lead to a different result.

537.312.62:538.569.4 2354
Millimeter-Wave Absorption in Superconducting Aluminum—M. A. Biondi, M. P. Garfunkel, and A. O. McCoubrey. (*Phys. Rev.*, vol. 101, pp. 1427–1429; February 15, 1956.) Available experimental evidence indicates that for wavelengths above ~1 cm the absorption is different for the normal and the superconducting states. Measurements are reported on a high-purity-Al waveguide; the ratio of surface resistivities for the superconducting and normal states is plotted as a function of temperature for mm values of λ. The results are discussed in relation to alternative energy-gap models.

537.312.62:538.569.4 2355
Very-High-Frequency Absorption in Superconductors—M. J. Buckingham. (*Phys. Rev.*, vol. 101, pp. 1431–1432; February 15, 1956.) Measurements Blevins *et al.* (1356 of 1956) indicate that the temperature at which the absorption departs from that in the normal state differs from the transition temperature T_0 by an amount strongly dependent on the frequency. A brief discussion shows that this dependence follows immediately from the concept of a gap in the electron-energy-level spectrum of a superconductor.

537.52 2356
Glow-to-Arc Transition—W. S. Boyle and F. E. Haworth. (*Phys. Rev.*, vol. 101, pp. 935–938; February 1, 1956.) Conditions for the glow-to-arc transition at moderately high pressures (50–1300 mm Hg) have been studied experimentally. Over this pressure range the transition is certain to occur only when the field reaches a critical value; this result is consistent with a field-emission mechanism for the transition.

537.525 2357
Theory of the High-Frequency Discharge in Gases at Low Pressures Determination of Starting Conditions—J. Salmon. (*J. Phys. Radium*, vol. 17, pp. 33–36; January, 1956.) Continuation of work reported previously (2910 of 1955). The importance of secondary emission from the walls is emphasized. Experimental and theoretical curves of starting voltage as a function of pressure for frequencies of 25, 42.8 and 70.6 mc are compared.

537.535.9.08 2358
Method of determining the Cathode Fall [of potential] in a [cold-cathode] Glow Discharge—K. Rademacher and K. Wojaczek. (*Naturwissenschaften*, vol. 43, p. 78; February, 1956.) A brief account is given of a method involving the use of a hot-cathode discharge in one cross-arm of a cruciform tube with its positive column surrounding the cold cathode in the other arm.

537.533 2359
Experimental Verification of the Wave-Mechanical Theory of Field Electron Emission—R. Haefer. (*Acta Phys. Austriaca*, vol. 10, pp. 149–161; January, 1956.) Methods described by Drechsler and Henkel (522 of 1955) are used to calculate the current density and field strength at a tungsten point and hence to estimate the accuracy of results obtained previously (1680 of 1941). Consideration is extended to the case when the tungsten is coated with foreign atoms.

537.533:621.38.032.212 2360
Electron Emission from Cold Metal Surfaces at Medium Field Strengths (~10^4V/cm)—K. Kerner. (*Z. Angew. Phys.*, vol. 8, pp. 1–8; January, 1956.) Report and discussion of measurements of the field emission of Fe, Al, and Ag cathodes. Ag gave the highest emission; Fe the lowest. See also 3193 of 1954 (Kerner and Raether).

537.56 2361
Dynamics of Ionized Media—S. Gasiorowicz, N. Neuman, and R. J. Riddell, Jr. (*Phys. Rev.*, vol. 101, pp. 922–934; February 1, 1956.) "The behavior of an ionized plasma is discussed in an approximation in which an individual particle is assumed to obey a Fokker-Planck equation, and where its interaction with the environment is incorporated in the coefficients of the partial differential equation."

538.221 2362
The Study of Ferromagnetism in the Institute of Physics at the University of Ferrara—A. Drigo. (*Ricerca Sci.*, vol. 26, pp. 138–143; January, 1956.) A short report outlining some of the more important results achieved. Both thin-film and massive specimens have been studied. Research on internal dissipation is being pursued.

538.3 2363
Electromagnetic Momentum and Electron Inertia in a Current Circuit—E. G. Cullwick. (*Proc. IEE*, part C, vol. 103, pp. 159–170; March, 1956.) The magnetic energy of a current circuit is identified with the kinetic energy of the mass equivalent of the total em energy of the conduction electrons. The concept of em momentum in a current circuit is used to determine the force on the end wire of a long rectangular circuit and to bring the known effects of electron inertia in a circuit within the scope of em theory.

538.561:[537.533.9+537.591.8 2364
Radiation emitted by a Uniformly Moving Electron in Electron Plasma in a Magnetic Field—A. A. Kolomenski. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 982–985; February 21, 1956. In Russian.) Results of the theoretical considerations presented indicate that Čerenkov-type em radiation may be produced by charged particles not necessarily moving at relativistic speeds, *e.g.*, cosmic particles in the ionosphere. The frequencies of the radiated ordinary waves lie in the range $\omega<\omega_0$, where ω_0 is the plasma angular frequency, that of extraordinary waves in the range $\omega_0<\omega<\sqrt{\omega_0^2+\omega_H^2}$, where ω_H is the gyrofrequency. Extraordinary waves only are produced a) in weak magnetic fields and b) by relativistic electrons.

538.566:535.13 2365
Investigation of the Propagation of [optical-type] Signals in Dispersive Media, using an Acoustic Model—T. Ankel. (*Z. Phys.*, vol. 144, pp. 120–131; January 17, 1956.) The theory of propagation of em waves in a dispersive medium, as developed by Sommerfeld and Brillouin, is experimentally verified by measurements on an acoustic model which comprises a long hollow tube with closely spaced Helmholtz resonators along it.

538.566:535.42 2366
An Approximate Theory of the Diffraction of an Electromagnetic Wave by an Aperture in a Plane Screen—R. F. Miller. (*Proc. IEE*, part C, vol. 103, pp. 177–185; March, 1956.) Theory based on the Sommerfeld half-plane solution is developed. For certain regions the field can be conceived as arising from the flow of electric and magnetic currents along the edge of the half-plane. This concept is extended to apertures of arbitrary form, the case of the circular aperture being studied particularly; the theory is supported by results of measurements.

621.3.013.78:538.221 2367
The Magnetic Screening Effect of Iron Tubes—P. Hammond. (*Proc. IEE*, part C, vol. 103, pp. 112–120; March, 1956.) The problem is approached by considering the induced pole strength on the surface of the iron; the distribution of pole strength in general produces a magnetic field which varies from place over the screened region. Observations support the calculated values of screening ratio.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16:551.5 2368
Radio Astronomy and the Fringe of the Atmosphere—A. C. B. Lovell. (*Quart. J.R. Met. Soc.*, vol. 82, pp. 1–14; January, 1956.) A survey presenting results of investigations on scintillations of radio stars in relation to the atmosphere in the 400-km altitude region, on meteor echoes, and on echoes from auroras. The determination of the total electron content in the earth-moon space by the study of lunar echoes is discussed.

523.16:621.396.822:551.510.535 2369
Cosmic Radio-Frequency Radiation near One Megacycle—G. Reber and G. R. Ellis. (*J. Geophys. Res.*, vol. 61, pp. 1–10; March, 1956.) Observations were made of cosmic radiation at 2.13, 1.435, 0.9, and 0.52 mc during the period March–October 1955, at Hobart, Tasmania. Photographs of specimen records are reproduced. For values of the critical frequency near the observing frequency there is strong correlation between critical frequency and received amplitude; for lower values of critical frequency the received amplitude increases to an independent limiting value. The greatest intensity of the radiation arriving from the zenith when the plane of the galaxy was overhead was 10^{-19} W/m^2 per cps per steradian at 2.13 mc. (The name "jansky" is suggested for the corresponding unit.) Only ordinary ionospheric propagation was important in the observations recorded.

523.78:538.56.029.6 2370
Observation of R.F. Emission from the Sun during the Solar Eclipse of 30th June 1954 at Byurakan [Armenian S.S.R.]—V. A. Sanamyan and G. A. Erznkanyan. (*Dokl. A. N. Arm. S.S.R.*, vol. 20, pp. 161–164; 1955. In Russian. *Referativnyi Zh.*, *Fizika*, Abstract 8172; March, 1956.) Results, presented graphically, of intensity measurements during this partial eclipse (97 per cent of total) show a decrease of up to 75 per cent at 1.5 m λ and of up to 35 per cent at 4.2 m λ. The diameter of the sun is 1.2 and 1.7 times the optical diameter at 1.5 m λ and 4.2 m λ, respectively.

550.38:523.165 2371
On Deriving Geomagnetic Dipole-Field Coordinates from Cosmic-Ray Observations—J. A. Simpson, F. Jory, and M. Pyka. (*J. Geophys. Res.*, vol. 61, pp. 11–22; March, 1956.) The coordinates of an equivalent dipole representing the external geomagnetic field can be determined from measurements of the nucleonic component longitude and latitude effects in the region of the geomagnetic equator. The measurement method and the relevant theory are outlined.

550.380.3 2372
Note on the Adjustment of Isomagnetic Charts to Mutual Consistency—A. J. Zmuda. (*J. Geophys. Res.*, vol. 61, pp. 57–58; March, 1956.)

550.385 2373
Variations in Strength of Wind System, in the Dynamo Mechanism for the Magnetic Diurnal Variation, deduced from Solar-Flare Effects at Huancayo. Peru—S. E. Forbush. (*J. Geophys. Res.*, vol. 61, pp. 93–105; March. 1956.)

551.510.41:523.78 2374
Electrophotometric Investigation of Atmospheric Ozone during the Solar Eclipses of 25th February 1952 and 30th June 1954—Sh. A. Bezverkhni, A. L. Osherovich, and S. F. Rodionov. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 651–654; February 1, 1956. In Russian.)

551.510.52:621.396.11.029.62 2375
Abnormal V.H.F. Propagation—Hooper. (See 2517.)

551.510.53 2376
Atmospheric Temperatures and Winds between 30 and 80 km—W. G. Stroud, W. Nordberg, and J. R. Walsh. (*J. Geophys. Res.*, vol. 61, pp. 45–56; March, 1956.) Rocket experiments made in New Mexico between July, 1950 and September. 1953 are described and the results are analyzed. The mean altitude distribution of temperature exhibits a maximum of about 270°K at 50 km, with a lapse rate of about 2.5°/km above the peak. The highest wind speed was observed during winter; its value was 104m/s at 55 km.

551.510.53 2377
Arctic Upper-Atmosphere Pressure and Density Measurements with Rockets—H. E. LaGow and J. Ainsworth. (*J. Geophys. Res.*, vol. 61, pp. 77–92; March, 1956.) Report of measurements made during 1953 and 1954 using rockets launched at an altitude of about 25 km from balloons; the greatest altitude reached was about 80 km. Results deviated in some cases from previous rocket measurements.

551.510.535 2378
Electron Distribution in the Ionosphere—G. A. M. King. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 184–185; March, 1956.) Addendum to an earlier note (117 of 1955).

551.510.535 2379
Irregularity and Regularity of the Sporadic-E Layer—K. Rawer. (*Geofis, Pura Appl.*, vol. 32, pp. 170–224; 1955. In German.) From observations at recording stations much information is obtained about the highest vertical-incidence reflection frequency fE_s, but less about the blanketing frequency f_bE_s. Results must be interpreted in statistical terms. Time- and distance-correlation functions are established. Diurnal, seasonal, and geographic regularities are discussed. A sharp maximum in fE_s occurs at the magnetic equator. No well defined influence of the solar cycle has been found and only a very weak lunar-tide effect. Observations of the variation of reflection coefficient with frequency have been made apart from the routine evaluations. In temperate latitudes in about a third of all cases there is no partial reflection; in other cases local variations of electron concentration are such that the peak-value/mean-value ratio is between 1 and 2; higher ratios are rare. At low latitudes the variation may be more important. Ionograms of different stations have been classified for transparency, scatter, angle of incidence, and layer development. Diffuse echoes exist often near the magnetic equator. In most cases E_s ionization originates as a thin layer of constant altitude. Transitory downward movements are responsible for E_{2s} in daytime. A cumulo-cirrus cloud layer is a good model for E_s ionization. Possible ionization processes are discussed. About 50 references.

551.510.535 2380
Temperature Distribution of the Ionosphere under Control of Thermal Conductivity—F. S. Johnson. (*J. Geophys. Res.*, vol. 61, pp. 71–76; March, 1956.) Bates' theory (988 of 1952) is extended. The energy absorbed in the F region is assumed to be conducted downward into a denser region where it is dissipated by infrared emission. Calculations indicate that the atmosphere is isothermal above about 250 km and that there is a very strong temperature gradient between 100 and 200 km. The temperature in the isothermal region must be assumed to be 1100°K to meet the requirement for the atmosphere near 300 km to support an F_2 region. The low temperature at 80 km is due primarily to the lack of absorbed energy there rather than to the presence of a strongly emitting layer.

551.510.535 2381
A New Method for obtaining Electron-Density Profiles from P′-f Records—J. E. Jackson. (*J. Geophys. Res.*, vol. 61, pp. 107–127; March, 1956.) "A practical and accurate method for reducing P'-f records to electron densities vs true height is described and used to analyze P'-f records taken at White Sands. Direct measurements of electron densities in the ionosphere obtained with the aid of rockets are used to check the method. Results obtained by these two independent techniques are shown to be in excellent agreement. Twenty P'-f records were reduced for the period from 1948 to 1954, all of which reveal a considerable degree of regularity in the height of the daytime E_1 and F_2 regions. Some of the profiles obtained are shown. One of the illustrations shows a one-hour sequence, where the F_2 virtual height varied from 650 km to 410 km, whereas the true height remained essentially unchanged."

551.510.535 2382
Airborne Ionospheric Measurements in the North Pole Area—G. Gassmann. (*J. Geophys. Res.*, vol. 61, pp. 136–138; March, 1956.) A brief preliminary report.

551.510.535:523.78:621.396.11 2383
Ionospheric Observations at Banaras during the Total Solar Eclipse on 20 June 1955—Banerjee, Surange, and Sharma. (See 2514.)

551.510.535:621.3.082.7 2384
Polarization of the Echoes from the Ionosphere—J. K. D. Verma and R. Roy. (*Indian J. Phys.*, vol. 30, pp. 36–46; January, 1956.) Details are given of an improved type of radio polarimeter, similar in principle to that evolved by Eckersley and Farmer (1968 of 1946), for operation in conjunction with high-resolution sounding equipment. With this arrangement it is possible to separate normal echoes from those received from thin E_s layers or cloud-type irregularities. Photographs of some observed polarization patterns are reproduced and briefly discussed.

551.510.535:621.396.812.3 2385
The Fading of Radio Waves of Frequencies between 16 and 2400 kc/s—Bowhill. (See 2519.)

551.510.535:621.396.812.3 2386
The Fading Periods of the E-Region Coupling Echo at 150 kc/s—Parkinson. (See 2520.)

551.510.535:621.396.812.3 2387
The Determination of the Horizontal Velocity of Ionospheric Movements from Fading Records from Spaced Receivers—D. W. G. Chappell and C. L. Henderson. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 163–168; March, 1956.) A method is presented which does not require a long sample of record and which is valid for any distribution of orientation of "lines of maximum amplitude." The derived formula is compared with that of Mitra (96 of 1950). See also 1418 of 1956 (Court).

551.510.535:621.396.812.3 2388
A Determination of Ionospheric Winds for a 24-Hour Period—G. W. G. Court and E. S. Gilfillan. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 169–170; March, 1956.) Results are given of an analysis, using the method proposed by Court (1418 of 1956), of the fading records obtained at spaced receivers during an arbitrarily chosen 24-h period. A diurnal variation of wind direction appears to be indicated.

551.594.1 2389
On the Deviations of the Course of Elements of Atmospheric Electricity on Continents from the Worldwide Course—R. Mühleisen. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 146–157; March, 1956.) Extensive recordings show that continental deviations from the normal oceanic diurnal pattern of potential-gradient and air-earth-current variations are caused mainly by positive space charges produced by urbanization, industry, and traffic and distributed by air movements.

551.594.5:551.510.535 2390
Relationships between Aurora and Sporadic-E Echoes at Barrow, Alaska—R. W. Knecht. (*J. Geophys. Res.*, vol. 61, pp. 59–69; March, 1956.) Report of observations of visual aurora made simultaneously with ionospheric soundings during March, 1951; the observations for three nights are described in detail. Analysis of the results shows that E_s echoes at frequencies >7 mc tend to occur when the aurora is near the zenith, that there is a direct relation between the brightness of inactive auroras and the top frequency of E_s echoes, and that E_s echo ranges correspond with estimated slant ranges of visible auroral forms.

551.594.6 2391
The Annual Variations of the Atmospherics—Existence and Explanation of a Second Maximum in Winter, if Only Strong Impulses are Counted—R. Reiter. (*J. Geophys. Res.*, vol. 61, pp. 23–26; March, 1956.) Records obtained at Munich of the numbers of atmospherics received in frequency ranges 10–50 kc and 4–12 kc over a period of five years are analyzed and evaluated from the meteorological point of view.

551.594.6 2392
Stanford-Seattle Whistler Observations—J. H. Crary, R. A. Helliwell, and R. F. Chase. (*J. Geophys. Res.*, vol. 61, pp. 35–44; March, 1956.) Observations of times of occurrence of whistlers were made at Seattle, Wash., and Stanford, Calif., for two hours every week from October, 1951 to October, 1952, in order to determine the percentage of whistlers received simultaneously at both locations; the figure obtained was about 22 per cent. This result is examined in relation to theories of whistler origin and propagation; it gives support to Storey's theory (142 of 1954). See also 1666 of 1955 (Koster and Storey).

551.594.6:551.510.535 2393
The "Nose" Whistler—a New High-Latitude Phenomenon—R. A. Helliwell, J. N. Crary, J. H. Pope, and R. L. Smith. (*J. Geophys. Res.*, vol. 61, pp. 139–142; March, 1956.) Spectrograms of a type of whistler observed at College, Alaska, made by analyzing tape recordings, are reproduced and discussed. The initial, or "nose", frequency depends primarily on gyrofrequency, hence such observations should enable the effects of gyrofrequency and plasma frequency on dispersion to be separated, thus leading to more reliable estimates of the ionization density in the outer ionosphere.

551.594.6:551.510.535 2394
The Interpretation of Pulse Trains associated with Lightning Flashes—W. O. Schumann. (*Z. Angew. Phys.*, vol. 8, pp. 24–28; January, 1956.) A treatment of the propagation of atmospherics by earth and ionosphere reflections, assuming a radiating dipole source. See also 717 of 1955 (Hepburn and Pierce).

551.510.535 2395
Proceedings of the Fourth Meeting of the Mixed Commission on the Ionosphere [Book

Review]—Publishers: Union Radio-Scientifique Internationale, Brussels, 238 pp., 1954. (*Brit. J. Appl. Phys.*, vol. 7, p. 83; February, 1956.) Contains papers presented in Brussels in August, 1954, and discussions on them.

LOCATION AND AIDS TO NAVIGATION

621.396.933.1 **2396**
Beam Deflection Tube Simplifies Radio Compass—J. M. Tewksbury. (*Electronics*, vol. 29, pp. 166–167; May, 1956.) A design is described in which two miniature Type 6AR8 beam-deflection tubes replace seven ordinary tubes, thus reducing the size and weight of the equipment.

621.396.96+621.396.932 **2397**
Modernization of Radio and Radar Equipment in H.M. Telegraph Ships—W. Dolman and P. W. J. Gammon. (*P.O. Elect. Engrs. J.*, vol. 48, Part 4, pp. 204–207; January, 1956.)

621.396.96 **2398**
Radar P.P.I. Display uses Precision Interlace—A. Shulman. (*Electronics*, vol. 29, pp. 168–171; May, 1956.) Target position data from automatic tracking computers are presented continuously in the form of marker dots on the ppi screen by combining the two sets of scanning waveforms.

621.396.96 **2399**
Radar Polarization Power Scattering Matrix—E. M. Kennaugh and C. D. Graves. (PROC. IRE, vol. 44, p. 695; May, 1956.) Comment on 1425 of 1956 and author's reply.

621.396.963.3:621.396.822 **2400**
Visual Detectability of Signals in Noise—J. W. R. Griffiths. (*Wireless Engr.*, vol. 33, pp. 118–120; May, 1956.) The variation of output-signal/noise ratio as a function of "contrast" (using this term in the sense of a bias) is investigated for different input-signal/noise ratios. The results agree qualitatively with experimental results on probability of detection in very simple systems.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 **2401**
Photoconductivity of Some Cyanine Dyes—R. C. Nelson. (*J. Opt. Soc. Amer.*, vol. 46, pp. 10–13; January, 1956.)

535.215 **2402**
Sensitization of Photoconductivity in Cadmium Sulfide [by cyanine dyes]—R. C. Nelson. (*J. Opt. Soc. Amer.*, vol. 46, pp. 13–16; January, 1956.)

535.215:537.323:546.817.221 **2403**
Thermoelectric Force Measurements on Illuminated Lead Sulphide—H. A. Müser. (*Z. Phys.*, vol. 144, pp. 56–65; January 17, 1956.) Experiments indicate that the true thermoelectric force is not affected by illumination; apparent variations of thermoelectric force are due to photovoltaic effects.

535.215:546.482.12 **2404**
Photoelectromagnetic Effect in Insulating CdS—H. S. Sommers, Jr., R. E. Berry, and I. Sochard. (*Phys. Rev.*, vol. 101, pp. 987–988; February 1, 1956.) "The short-circuit photoelectromagnetic current in insulating crystals of cadmium sulfide has been measured in a batch of electroluminescent crystals. The product (mobility)$^3\times$(lifetime) is found to be 1 cm^6/volt^3 sec.^{2}. The sensitivity of the equipment is sufficient to detect the photoelectromagnetic effect for crystals whose product is as low as 10^{-5} cm^6/volt^3 sec.^{2}"

535.37 **2405**
Associated Donor-Acceptor Luminescent Centers—J. S. Prener and F. E. Williams. (*Phys. Rev.*, vol. 101, p. 1427; February 15, 1956.) Brief discussion of the properties of (Zn, Cd) (S, Se) phosphors activated with Cu, Ag, Au, P, As or Sb and coactivated with Cl, Br, I, Al, Ga, or In.

535.37:535.215 **2406**
Dielectric Changes in Inorganic Phosphors—S. Kronenberg and C. A. Accardo. (*Phy. Rev.*, vol. 101, pp. 989–992; February 1, 1956.) ZnS-CdS phosphors were investigated by using them as dielectrics in capacitors and exposing them to light. Observed capacitance changes are attributed partly to photoconduction effects and partly to true variations of dielectric constant.

535.37:535.215:546.472.21 **2407**
Determination of the Ratio of Effective Cross-sections of Capture and Recombination of Optical [photo-liberated] Electrons in ZnS-Cu, Co Crystal Phosphors—Syul Syul-Yun. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 818–821; February 11, 1956. In Russian.)

535.37:546.48.185.161 **2408**
Lead- and Manganese-Activated Cadmium Fluorophosphate Phosphors—R. W. Wollentin. (*J. Electrochem. Soc.*, vol. 103, pp. 17–23; January, 1956.)

535.376 **2409**
Influence of Temperature on the Electroluminescence of Zinc Sulphides—J. Mattler. (*J. Phys. Radium*, vol. 17, pp. 42–51; January, 1956.) An extended account of the investigation described previously (1362 of 1955).

535.376:546.472.21 **2410**
Time-Average Electroluminescence Output of some Zinc Sulfide Phosphors—S. Nudelman and F. Matossi. (*J. Electrochem. Soc.*, vol. 103, pp. 34–38; January, 1956.) "Dependence of time-average electroluminescence output on field strength and frequency is observed for frequencies up to 20 kc for green and blue emission. The field dependence can be described either by a power law or by an exponential law. The frequency dependence is discussed in terms of theoretical relations connecting the light output to recombination characteristics or to polarization effects. The polarization effects are of minor importance. Light outputs from sinusoidal and square wave excitation are compared."

535.376:546.472.21:537.226.2 **2411**
Dielectric Behaviour of Electroluminescent Zinc Sulfides—W. Lehmann. (*J. Electrochem. Soc.*, vol. 103, pp. 24–29; January, 1956.) Measurements were made at voltages up to 600 rms and frequencies up to >20 kc; the measurement cell and technique are discussed. Both the real and the imaginary parts of the dielectric constant vary inversely with frequency. The results support previous assumptions that the excitation mechanism for electroluminescence is different from that for photoluminescence, while the emission mechanisms are similar.

537.226/.227 **2412**
Behavior of Ferroelectric $KNbO_3$ in the Vicinity of the Cubic-Tetragonal Transition—S. Triebwasser. (*Phys. Rev.*, vol. 101, pp. 993–997; February, 1956.) Measurements on $KNbO_3$ single crystals are reported. From the values found for dielectric constant and spontaneous polarization a determination can be made of the first three terms of the power series expressing the free energy in terms of the polarization. The behavior of the crystals in the cubic and tetragonal phases is in reasonable agreement with predictions based on Devonshire's theory for $BaTiO_3$ (663 of 1950 and 1341 of 1952); the corresponding constants for the two materials are of the same order of magnitude.

537.226/.227:546.431.824–31 **2413**
A Microstructure Study of Barium Titanate Ceramics—F. Kulcsar. (*J. Amer. Ceram. Soc.*, vol. 39, pp. 13–17; January 1, 1956.) Polishing and etching techniques for preparing polycrystalline $BaTiO_3$ for metallographic examination are described. Photomicrographs are reproduced and discussed. A companion paper by Cook (*ibid*, pp. 17–19) analyses some of the domain patterns found.

537.226/.227:546.431.824–31 **2414**
A Modified Replica Technique and its Application to the Examination of Etched Single Crystals of Barium Titanate—D. S. Campbell and D. J. Stirland. (*Brit. J. Appl. Phys.*, vol. 7, pp. 62–65; February, 1956.)

537.227 **2415**
Properties of Guanidine Aluminum Sulfate Hexahydrate and some of its Isomorphs—A. N. Holden, W. J. Merz, J. P. Remeika, and B. T. Matthias. (*Phys. Rev.*, vol. 101, pp. 962–966; February 1, 1956.) Report of an experimental study of this new class of ferroelectrics (2987 of 1955). The crystals are trigonal, with the ferroelectric direction along the trigonal axis. The 60-cps hysteresis loops are often biased or double, the shape being correlated with the location of the specimen in the mother crystal. At room temperature the saturation polarization is about 0.35 μC/cm^2 and the coercive force at 60 cps in 1–3 kv/cm; these quantities increase with falling temperature. The small-signal dielectric constant is about 6 along the axis and about 5 perpendicular to it. The switching characteristics resemble those of $BaTiO_3$, but the present crystals are considerably slower.

537.228.1:548.5 **2416**
The Laboratory Production of Large Water-Soluble Crystals—E. A. Taylor. (*P.O. Elect. Engrs.' J.* vol. 48, part 4, pp. 219–223; January, 1956.) The production of synthetic crystals having useful piezoelectric properties is described.

537.311.3 **2417**
The Conductivity of an Antimony-Caesium Layer—L. I. Shafratova-Ekertova. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1357–1363; August, 1955.) If an Sb-Cs layer is in contact with two metallic electrodes and a constant potential difference is applied to the electrodes, the current through the layer increases with time. A report is presented on an experimental investigation into the physical nature of this phenomenon; the results obtained are interpreted theoretically. It is suggested that the phenomenon is due to polarization of the layer in a sense facilitating the passage of current.

537.311.3:539.23 **2418**
On the Measurement of Electric Constants of Thin Metallic Films—G. Bonfiglioli, E. Coen, and R. Malvano. (*J. Appl. Phys.*, vol. 27, pp. 201–203; March, 1956.) For thin films, whose detailed geometrical form is usually unknown, it is impossible to determine the Hall coefficient and conductivity separately, but their product—the Hall mobility of the carriers—can be determined. Measurements on Au films evaporated on to mica bases are reported; the mobility is independent of film thickness between 100 and 600 Å, but its value is only about a quarter of the bulk mobility. The structure of thin films is discussed in the light of this result.

537.311.3:546.841.4—31 **2419**
Polarization in Thorium Oxide Crystals—W. E. Danforth and J. H. Bodine. (*J. Franklin Inst.*, vol. 260, pp. 467–483; December, 1955.) Description and discussion of phenomena observed when a constant current is passed through a thoria crystal in vacuum at temperatures between 900° and 1300°C. Resistivity

varied from 6000 to 400 Ω\cm over this temperature range. Conduction seems to be almost entirely ionic, the electron current being <1 per cent of the total.

537.311.31+537.311.33 **2420**
On the Transport Properties of Metals and Semiconductors—D. Ter Haar. (*Physica*, vol. 22, pp. 61–68; January, 1956.) Simple kinetic theory is used to derive approximate expressions for the thermal conductivity, electrical conductivity, thermoelectric power, Hall constant, and magnetoresistance of metals and semiconductors.

537.311.31 **2421**
Modulation of Conductivity by Surface Charges in Metals—G. Bonfigioli, E. Coen, and R. Malvano. (*Phys. Rev.*, vol. 101, pp. 1281–1284; February 15, 1956.) Measurements were made on thin films of Au, Bi and Sb; a large surface charge was induced by application of an electric field. A tentative interpretation of the observed conductivity changes is offered. See also 2418 above.

537.311.32+536.21]:546.26 **2422**
The Thermal and Electrical Conductivities of Deposited Carbon—A. R. G. Brown, W. Watt, R. W. Powell, and R. P. Tye. (*Brit. J. Appl. Phys.*, vol. 7, pp. 73–76; February, 1956.) Measurements were made in the temperature range 20°–200°C. on commercial graphite and deposited carbon formed at 1800°, 1900°, 2000° and 2100°C. The thermal conductivities of the specimens deposited at 2000° and 2100°C. were respectively 20 per cent and 40 per cent greater than that of copper. The electrical resistivity at 20°C. was about 24.5×10^{-5}Ω. cm for carbon deposited at 2100°C. compared with 76.2×10^{-5} for commercial graphite or 323×10^{-5} for carbon deposited at 1800°C. The results are tabulated.

537.311.33 **2423**
Transport and Deformation-Potential Theory for Many-Valley Semiconductors with Anisotropic Scattering—C. Herring and E. Vogt. (*Phys. Rev.*, vol. 101, pp. 944–961; February 1, 1956.) A theory of transport phenomena is presented based on assumptions regarding the scattering processes which are less restrictive than those made previously [*e.g.*, 2642 of 1955 (Herring)]. Three relaxation times are assumed, corresponding to the three principal directions of the ellipsoidal energy surfaces. Expressions for carrier mobility, Hall effect, magnetoresistance, piezoresistance, and hf dielectric constant are derived in terms of the relaxation-time tensor. The deformation-potential approach of Bardeen and Shockley (3032 of 1950) is generalized to suit the many-valley model. The results are used for correlating mobility, piezoresistance, etc. for *n*-type Si and Ge.

537.311.33 **2424**
Effects of Pressure on the Electrical Properties of Semiconductors—D. Long. (*Phys. Rev.*, vol. 101, pp. 1256–1263; February 15, 1956.) Continuing previous work (*e.g.*, 161 of January), various measurements have been made on Ge, InSb, InAs, GaSb, Te, and Mg_2Sn at pressures up to 2000 atm. The results are used to deduce the pressure dependence of carrier concentration, mobility, energy gap, and effective mass. Whereas in Ge and Mg_2Sn the energy gap increases with increased pressure, in Te it decreases.

537.311.33 **2425**
Simultaneous Transport of Heavy and Light Holes in Semiconductors with a Degenerate Valence Band—E. S. Rittner. (*Phys. Rev.*, vol. 101, pp. 1291–1294; February 15, 1956.) "A theoretical study is presented of the motion in an *n*-type semiconducting filament of an injected narrow pulse of slow and fast holes subject to drift, diffusion, recombination, and reversible interband transitions. For low injection level and for interband transition times which are small compared to the recombination lifetime and to the observation time but large compared to the time between collisions, it is shown that both sets of holes propagate and broaden as a single pulse with a group mobility and diffusivity heavily weighted by that of the slower holes. This explains why only a single pulse is observed at the collector in drift mobility experiments."

537.311.33 **2426**
Degeneracy of the Electron Gas in Semiconductors—A. G. Samoilovich and L. L. Korenblit. (*Uspekhi fiz. Nauk*, vol. 57, pp. 577–630; December, 1955.) A survey of work on factors affecting and effects of the electron-gas degeneracy. 42 references, about half of which are to Russian literature.

537.311.33 **2427**
A Formula for the Voltage/Current Characteristic of an n-p Junction—V. L. Bonch-Bruevich and E. Ya. Pumper. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1520–1521; August, 1955.) If formula (1) proposed by Shockley (*Theory of Electronic Semiconductors*, 1953) is correct, then the ratio of the forward and reverse currents, for the same absolute value of the applied voltage, should be independent of the nature of the semiconductor. Such a general conclusion does not seem to be justifiable, hence the accuracy of the formula is questionable.

537.311.33:535.215 **2428**
Photoelectric Phenomena with Copper Phthalocyanine—H. Baba, H. Chitoku, and K. Nitta. (*Nature, Lond.*, vol. 177, p. 672; April 7, 1956.) Experiments are briefly reported, the results of which indicate that this material is a semiconductor exhibiting photoconductive and photovoltaic properties.

537.311.33:535.215:546.682.86 **2429**
Photoconductive and Photoelectromagnetic Effects in InSb—S. W. Kurnick and R. N. Zitter. (*J. Appl. Phys.*, vol. 27, pp. 278–285; March, 1956.) Measurements on single-crystal plates of *p*-type InSb were made at temperatures of 77° and 301°K. Electropolished and mechanically polished specimens yielded very different results. A simple two-dimensional theoretical model is used to interpret the results; it is assumed that hole-electron pairs are produced by the illumination at the surface only.

537.311.33:537.32:621.362 **2430**
Increasing the Efficiency of Semiconductor Thermocouples—A. F. Ioffe, S. V. Airapetyants, A. V. Ioffe, N. V. Kolomoets, and L. S. Stil'bans. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, p. 981; February 21, 1956. In Russian.) Results of theoretical considerations indicate that for maximum efficiency a high ratio of charge-carrier mobility to thermal conductivity is desirable. This can be achieved by using an isomorphous impurity to decrease the thermal conductivity; the charge-carrier mobility remains practically unchanged. The maximum efficiency occurs, theoretically, when the thermal emf of each branch of the thermocouple is about ±200 μv/degree.

537.311.33:546.23 **2431**
Influence of Light on the Dipole Absorption of E. M. Radiation by Selenium—Y. Meinnel, J. Meinnel, and Y. Balcou. (*J. Phys. Radium*, vol. 17, pp. 78–79; January, 1956.) Experiments with powder specimens of Se show that the activation energy is considerably lower for the illuminated than for the unilluminated material.

537.311.33:[546.28+546.289 **2432**
Electrolytic Shaping of Germanium and Silicon—A. Uhlir, Jr. (*Bell Syst. Tech. J.*, vol. 35, pp. 333–347; March, 1956.) Barrier effects and other phenomena occurring in the electrolytic etching of semiconductors, especially Ge, are described and techniques for using them in the shaping of semiconductor devices are discussed. Auxiliary techniques for localizing the action include optical illumination. Suitable electrolytes for etching Ge and Si are indicated.

537.311.33:[546.28+546.289 **2433**
Infrared Absorption and Oxygen Content in Silicon and Germanium—W. Kaiser, P. H. Keck, and C. F. Lange. (*Phys. Rev.*, vol. 101, pp. 1264–1268; February 15, 1956.) "An optical absorption band at 9μ has been correlated with the oxygen content in silicon. Pulled silicon crystals were found to contain up to 10^{18} oxygen atoms per cm³ which seem to originate from the quartz crucible. The oxygen concentration in silicon crystals prepared by the floating zone technique in vacuum was found to be less than 10^{16} oxygen atoms per cm.³ The 9μ absorption due to silicon-oxygen bond stretching vibrations provides a possibility for a quantitative oxygen analysis of high sensitivity. A corresponding absorption in germanium at 11.6μ is believed to be due to a germanium-oxygen vibration."

537.311.33:[546.28+546.289 **2434**
Surface States on Silicon and Germanium Surfaces—H. Statz, G. A. deMars, L. Davis, Jr., and A. Adams, Jr. (*Phys. Rev.*, vol. 101, pp. 1272–1281; February 15, 1956.) Measurements of the conductivity of *p*-type inversion layers on *n*-type crystals are reported and discussed. The steady-state conductance is related to a high density of surface states outside the oxide film which forms on the surface, while the nonsteady-state conductance is related to states located at the semiconductor/oxide interface; the energy levels of the latter states are 0.455 and 0.138 ev below the middle of the energy gap for Si and Ge respectively. The mechanism of charge transfer through the oxide film is not yet clear.

537.311.33:[546.28+546.289 **2435**
Effect of Dislocations on the Minority Carrier Lifetime in Semiconductors—A. D. Kurtz, S. A. Kulin, and B. L. Averbach. (*Phys. Rev.*, vol. 101, pp. 1285–1291; February 15, 1956.) "The density of random dislocations in germanium and silicon crystals has been measured by means of X-ray rocking curves and by etch pit counting. Data obtained by the two methods are in good agreement, and dislocation densities in the range 10^4-10^7/cm² were found. The minority carrier lifetime was shown to vary with the dislocation density, and the results could be expressed in terms of a recombination efficiency per unit length of dislocation line, $\sigma_R=1/N_D\tau$ (where N_D=dislocation density, τ=lifetime). σ_R was found to decrease with increasing resistivity of germanium and was higher for silicon than for germanium of comparable purity."

537.311.33:[546.28+546.289 **2436**
Ionization by Collision in Silicon and Germanium—E. Groschwitz. (*Z. Phys.*, vol. 143, pp. 632–636; January 10, 1956.) The ionization coefficient α of an electronic semiconductor is calculated as a function of the electric field strength. Comparison with experimental results [1079 of 1954 (McKay and McAfee)] indicates that ionization by collision is determined by the interaction of the conduction electrons with acoustic as well as optical energy quanta.

537.311.33:[546.28+546.289]:621.311.6 **2437**
The Electron-Voltaic Effect in Germanium and Silicon P-N Junctions—P. Rappaport, J. J. Loferski, and E. G. Linder. (*RCA Rev.*, vol. 17, pp. 100–128; March, 1956.) Extension of the investigation reported previously [1799 of 1954 (Rappaport)] of the voltage produced across the junction as a result of β-particle bom-

bardment. Measurements on large-area alloy junctions are reported. The maximum efficiency of conversion of the radioactive power is probably $\gg 5$ per cent for Si; measured values of 2.5 per cent were obtained. The life of the devices may be limited by the damage due to the high-energy bombardment.

537.311.33:546.28 2438

Photographs of the Stress Field around Edge Dislocations—W. L. Bond and J. Andrus. (*Phys. Rev.*, vol. 101, p. 1211; February 1, 1956.) Infrared photographs obtained with Si crystals are reproduced.

537.311.33:546.28:621.314.7 2439

Surface Treatment of Silicon for Low Recombination Velocity—A. R. Moore and H. Nelson. (*RCA Rev.*, vol. 17, pp. 5–12; March, 1956.) Surface recombination velocities comparable with those for Ge can be obtained for *p*-type Si by treating the surface chemically so as to produce films of aniline-like aromatic liquids or of salts of sodium dichromate type. Some details are given of the techniques involved and of the resulting improvement in the current amplification factor of *n-p-n* transistors. The treatment also eliminates channeling leakage at junctions. This is consistent with the theory that the films cause the energy bands to curve upwards at the surface.

537.311.33:546.289 2440

Effects of the Dislocations on Minority Carrier Lifetime in Germanium—J. Okada (*J. Phys. Soc. Japan*, vol. 10, pp. 1110–1111; December, 1955.) A quantitative study is made, based on measurements of dislocation density, recombination velocity, and carrier lifetime; the results are compared with theoretically deduced relations between these quantities. The density of active recombination centers along a dislocation is deduced to be about 6×10^5/cm.

537.311.33:546.289 2441

Effect of Electric Field on Surface Recombination Velocity in Germanium—J. E. Thomas, Jr. and R. H. Rediker. (*Phys. Rev.*, vol. 101, pp. 984–987; February 1, 1956.) Experiments are described confirming theoretical predictions of correlation between surface recombination velocity and surface potential for *n*-type Ge. The results also give support to explanations of semiconductor excess noise based on the assumption of surface traps. The influence of the ambient atmosphere is studied.

537.311.33:546.289 2442

Relaxation Effects in Recombination Velocity on Germanium Surfaces under Transverse Electrostatic Fields—A. Many, Y. Margoninski, E. Harnik, and E. Alexander. (*Phys. Rev.*, vol. 101, pp. 1433–1434; February 15, 1956.) Relaxation effects in surface recombination velocity were observed using the experimental technique described by Henisch and Reynolds (3652 of 1955). An explanation given previously in connection with surface conductivity [*e.g.*, 3649 of 1955 (Kingston)] is applicable in this case also. The experimental results are correlated with corresponding data on surface conductivity in a separate paper (*ibid.*, pp. 1434–1435).

537.311.33:546.289 2443

Delay Time of Plastic Flow in Germanium—J. R. Patel (*Phys. Rev.*, vol. 101, pp. 1436–1437; February 15, 1956.) Some experimental results are reported and discussed.

537.311.33:546.289 2444

Simple Method of Revealing p-n Junctions in Germanium—R. W. Jackson. (*J. Appl. Phys.*, vol. 27, pp. 309–310; March, 1956.) Details are given of an electrolytic etching method in which the Ge slice has a silver electrode painted on one side and the electrolyte is prevented from wetting this side.

537.311.33:546.289:621.396.822 2445

Excess Noise Spectra in Germanium—F. J. Hyde. (*Proc. Phys. Soc.*, vol. 69, pp. 242–245; February 1, 1956.) Spectral distributions are investigated in which the noise varies as f^{-m} over a wide range of frequencies, where m is not equal to unity. A diagram shows the spectrum for an *n*-type Ge filament, synthesized from two f^{-1} arc cot ($\omega\tau$) spectra with limiting relaxation times τ_1 and τ_2 of 2×10^{-4} and 5×10^{-7} second respectively. By superposing two basic spectrums of this type such as are postulated to arise from two recombination center levels, it is possible to generate an exponent which is practically constant at 1.24 over three decades of frequency. See also 2558 below.

537.311.33:546.561–31 2446

Lattice Defects and Dipole Absorption of E. M. Radiation by Cuprous Oxide—J. Meinnel, E. Daniel, and Y. Colin. (*J. Phys. Radium*, vol. 17, pp. 79–80; January, 1956.) Experiments have been made on a number of samples at temperatures from 77° to 350°K and at frequencies from 50 cps to 28 mc. Strong dipole absorption bands were observed in all cases. Activation energies between 0.18 and 0.44 ev are deduced.

537.311.33+538.22]:[546.823.171 +546.823.261]:539.23 2447

Some Electrical Properties of Titanium Nitride and Titanium Carbide—A. Münster and K. Sagel. (*Z. Phys.*, vol. 144, pp. 139–151; January 17, 1956.) Full report of experiments noted in 1386 of 1955 (Münster *et al.*). The semiconductor properties observed in TiN and TiC films deposited on SiO_2 may be due to inclusion of oxygen atoms.

537.311.33:546.863.683.231 2448

Electron Diffraction Determination of the Structure of $Tl_2Sb_2Se_4$—Z. G. Pinsker, S. A. Semiletov, and E. N. Belova. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 1003–1006; February 21, 1956. In Russian.)

537.311.33:546.873.241:536.21 2449

The Thermal Conductivity of Bismuth Telluride—H. J. Goldsmid. (*Proc. Phys. Soc.*, vol. 69, pp. 203–209; February 1, 1956.) Measurements over the temperature range 150°–300°K indicate that the electronic contribution to the heat conduction is considerably greater for specimens in which the charge carriers are intrinsically excited than for specimens in which most of the carriers arise from an impurity concentration. This result can be explained by a theory which takes into account the transfer of ionization energy down a temperature gradient.

538.22:546.3–1–71–59 2450

Ferromagnetic and Antiferromagnetic Properties of the System Gold-Manganese—A. Kussmann and E. Raub. (*Z. Metallkde*, vol. 47, pp. 9–15; January, 1956.) Experiments have established the existence of the ferromagnetic phase Au_4Mn. The phase Au_2Mn exhibits ferromagnetism and antiferromagnetism. The phase AuMn may also be antiferromagnetic.

538.221 2451

Intradomain Magnetic Saturation and Magnetic Structure of γ-Fe_2O_3—W. E. Henry and M. J. Boehm. (*Phys. Rev.*, vol. 101, pp. 1253–1254; February 15, 1956.) "The average moment for γ-Fe_2O_3 was found to be 1.18 Bohr magnetons per iron atom, which supports a preferential distribution of iron vacancies on octahedral sites in a spinel structure. A sample motion ballistic method was used for direct measurement of magnetic moments."

538.221 2452

Ferromagnetic Domain Nucleation in Silicon Iron—L. F. Bates and D. H. Martin. (*Proc. Phys. Soc.*, vol. 69, pp. 145–152; February 1, 1956.) "Powder deposit patterns obtained by the authors on single crystals of 3 per cent silicon iron are consistent with a description of the processes of domain phase creation in terms of nucleation at inclusions in the surfaces. Domain nucleation is discussed in relation to various magnetic properties."

538.221 2453

Variation of the Magnetic Anisotropy Energy of Ni and of Ni-Cu Alloys as a Function of Temperature—M. Sato and Y. Tino. (*J. Phys. Radium*, vol. 17, pp. 5–8; January, 1956.) Theory is based on consideration of the magnetic interactions between atoms in a crystal lattice. The anisotropy constant can be expressed as an exponential function of the square of the absolute temperature.

537.311.33:538.221 2454

Induced Ferromagnetism demonstrated by Addition of Lithium Ions to Nickel Oxide—N. Perakis, A. Serres, G. Parravano, and J. Wucher. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1275–1277; March 5, 1956.)

538.221:538.569.4 2455

Resonance of Ferrites at the Compensation Point in a Circularly Polarized Field (Ferrimagnetic Resonance)—J. Pauleve and B. Dreyfus. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1273–1275; March 5, 1956.) Measurements on ferrites of composition $Li_{0.5}Fe_{2.5-a}{}^{+++}Cr_a{}^{+++}O_4$ in a circularly polarized uhf field indicate that two types of resonance line can be distinguished in the neighborhood of the compensation temperature; the gyromagnetic ratio corresponding to one of these is of sign opposite to that of the electron.

538.221:621.318.1 2456

Ferromagnetism in Relation to Engineering Magnetic Materials—F. Brailsford. (*Proc. IEE*, part A, vol. 103, pp. 39–51; February, 1956.) ". . . a review is given of theoretical and experimental work mainly within the past ten years. This includes an account of ferromagnetic domains and of the small-particle theory of high coercivity. A description of the ferrites and of ferrimagnetism is given, and this is followed by a discussion of recent observations and ideas on the magnetic phenomena occurring at frequencies up into the microwave region."

538.632 2457

Significance of Hall-Effect Measurements on Alloys—B. R. Coles. (*Phys. Rev.*, vol. 101, pp. 1254–1255; February, 1956.) Hall-effect data for a number of alloys are examined in the light of the accepted theory of this effect. It is concluded that assumptions made about the relaxation time which seem appropriate when scattering by lattice vibrations predominates are not valid when scattering by solute atoms becomes significant.

539.234 2458

Influence of the Thickness of Thin [evaporated] Films on their Structure: Case of the Alloy Cu-Be—A. Viswanathan. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1586–1587; March 19, 1956.)

548.0:[546.47+546.289 2459

Investigations of Texture in Thin Zinc and Germanium Films—A. Segmüller. (*Z. Kristallog.*, vol. 107, pp. 18–34; January, 1956.) An electron-optical crystallographic investigation at temperatures between 90 and 770°K is reported. Good single-crystal texture may be obtained at relatively low temperatures by depositing the Zn or Ge on a zinc-blende cleavage plane.

548.0:549.514.51 2460

Laws of Intergrowth of Oriented Rutile Inclusions in Quartz—J. von Vultée. (*Z. Kristallog.*, vol. 107, pp. 1–17; January, 1956.)

621.3.066.6:537.311.4 2461

Contact Resistance and Surface of Contact—K. Millian and W. Rieder. (*Z. Angew. Phys.*, vol. 8, pp. 28–34; January, 1956.) An investigation of the influence of contact pressure, aging and surface treatment, including greasing, on the contact resistance of crossed cylinders of Cu, Ag and W. For Cu and for heat-treated W the contact resistance increases exponentially with time.

621.315.612 2462

Boron Nitride—K. M. Taylor. (*Mater. and Meth.*, vol. 43, pp. 88–90; January, 1956.) Physical properties of this ceramic insulating material are indicated.

621.315.615:537.52 2463

Breakdown Field Strength in Dielectric Liquids with Different Molecular Structure—E. Musset, A. Nikuradse, and R. Ulbrich. (*Z. Angew. Phys.*, vol. 8, pp. 8–15; January, 1956.) Report of measurements made on some 30 organic liquids at atmospheric pressure, using dc and 50-cps ac.

621.315.616:537.533.9 2464

Electrons produce High-Temperature Dielectric—J. B. Meikle and B. Graham. (*Electronics*, vol. 29, pp. 146–149; May, 1956.) A modified polyethylene irradiated by high-energy electrons has a volume resistivity of $10^{19}\Omega$.cm, dielectric constant of 2.3 and power factor <0.0007, while retaining the other physical properties of normal polyethylene; the new material is stable up to 300°C. Technique involved in the use of the electron accelerator in the processing of cable insulation etc. is outlined.

MATHEMATICS

517:519.241.1 2465

A Short Table of the Laguerre Polynomials—L. J. Slater. (*Proc. IEE*, part C, vol. 103, pp. 46–50; March, 1956.) A function referred to in connection with correlation analysis [2352 of 1955 (Lampard)] is tabulated.

517.5 2466

The Method of Stationary Phase—G. Braun. (*Acta Phys. Austriaca*, vol. 10, pp. 8–33; January, 1956.) The method presented is useful for evaluating an integral encountered in diffraction theory.

517.5 2467

On the Error Function of a Complex Argument—J. Kestin and L. N. Persen. (*Z. Angew. Math. Phys.*, vol. 7, pp. 33–40; January 25, 1956. In English.) A simplified method of dealing with the error function is presented, based on a transformation. Both the real and the imaginary components can be split into two parts, of which the first can be expressed by elementary functions while the second can be represented by two integrals which can be easily evaluated.

517.9 2468

Integration of a Nonlinear Integral Equation—P. Lévy. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1252–1255; March 5, 1956.) The equation discussed is that relating the kernel and the covariance of a random Laplace function $\phi(t)$.

519.281.2 2469

Two Methods of obtaining Least Squares Lines—I. H. Sher. (*Science*, vol. 123, pp. 102–104; January 20, 1956.)

518.2 2470

Index Mathematischer Tafelwerke und Tabellen aus allen Gebieten der Naturwissenschaften [Book Reviews]—K. Schütte. Publishers: R. Oldenbourg, Munich, 1955. 143 pp., (*Nature, Lond.*, vol. 177, p. 767; April, 1956.)

MEASUREMENTS AND TEST GEAR

529.7 2471

Atomic and Astronomical Time—L. Essen and J. V. L. Parry. (*Nature, Lond.*, vol. 177, pp. 744–745; April 21, 1956.) The National Physical Laboratory quartz clock has been calibrated at regular intervals by reference to the Cs resonator (3686 of 1955). A curve shows the frequency deviation over the period from June, 1955 to January, 1956, and a second curve shows the deviation from the Greenwich monthly revised values. The difference between these two curves represents the variation of the unit of astronomical time in terms of the Cs standard. Further curves show the Greenwich values reassessed after a long period, and astronomical observations, both in terms of the Cs standard.

621.3.001.3(083.74):621.318.423 2472

The Effect of Humidity on the Stability of Inductance Standards—G. H. Rayner and L. H. Ford. (*J. Sci. Instrum.*, vol. 33, pp. 75–77; February, 1956.) The changes in inductance of the National Physical Laboratory substandard inductance coils for a 10 per cent increase in relative humidity range from +0.7 part in 10^4 on a 100-μH coil to -0.25 part in 10^4 on a 10000-μH coil. The changes may be partly explained by dimensional changes in the formers of the coils.

621.317.3:621.396.822 2473

Fluctuations in a Loaded Line—V. S. Troitski. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1426–1435; August, 1955.) In measuring noise in a dipole connected to the noise meter by a long line conditions may be created under which the intrinsic noise of the meter input is reflected from the dipole, and thus the signal to be measured is changed by an unown quantity. Conditions necessary for the measurement of weak noise are discussed, sources of errors are indicated, and methods for their elimination are proposed.

621.317.3:[621.396.96+621.397.5 2474

Video Measurements employing Transient Techniques—H. A. Samulon. (Proc. IRE, vol. 44, pp. 638–649; May, 1956.) Equipment and waveforms suitable for testing television, radar, and other systems are reviewed, and methods of measurement and evaluation of the response characteristics are discussed. The effect on the transient response of small variations in the transfer function of the system is examined. Criteria for assessing transient response are listed.

621.317.3(083.74):621.372.029.3 2475

IRE Standards on Audio Systems and Components: Methods of Measurement of Gain, Amplification, Loss, Attenuation, and Amplitude-Frequency-Response, 1956—(Proc. IRE, vol. 44, pp. 668–686; May, 1956.) Standard 56 IRE 3.S1.

621.317.33.084.2:537.311.33 2476

A Four-Point Probe Apparatus for Measuring Resistivity—D. B. Gasson. (*J. Sci. Instrum.*, vol. 33, p. 85; February, 1956.) Description of apparatus with probe spacing of 1 mm and minimum surface leakage path of 1 cm which has been developed for measuring resistivities of semiconductors.

621.317.335.3 2477

Line Corrections in Permittivity Measurements at Frequencies below 50 Mc/s—R. Guillien. (*J. Phys. Radium*, vol. 17, pp. 52–56; January, 1956.) Classical line theory is used to determine the corrections necessary to take account of the line joining the specimen to the measuring apparatus in measurements at mc frequencies. Formulas are derived giving ϵ' and ϵ'' directly in terms of the measured value of the apparent dielectric constant.

621.317.335.3.029.64 2478

A Centimetre-Wave Parallel-Plate Spectrometer—P. H. Sollom and J. Brown. (*Proc. IEE*, part B, vol. 103, pp. 419–428; May, 1956.) The instrument described is based on the same general principles as that of Culshaw (1135 of 1954), but the radiation is enclosed between two parallel disks of diameter about 4 feet whose spacing, which is equal to the height of the specimen, is only 3/16 inch. The radiation is injected by means of a sectoral horn free to move over a 90° sector of the disk circumference, and is detected by a similar horn or a waveguide located anywhere on the circumference.

621.317.337:621.372.412:537.228.1 2479

A New Method for measuring the Quality Factor Q of Piezoelectric Crystals—H. Mayer. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1428–1430; March 12, 1956.) HF oscillations exciting the crystal at its resonance frequency are interrupted by means of a vibrator at 100 cps. During the interruptions the crystal executes damped oscillations; these are displayed on a cro and the Q factor hence determined. For glucinium sulphate the Q value obtained is 6700 in vacuum, 3300 in air.

621.317.34:621.317.729:621.372.2 2480

An Investigation into some Fundamental Properties of Strip Transmission Lines with the Aid of an Electrolytic Tank—Dukes. (See 2283.)

621.317.341.029.63:621.315.212 2481

Determination of the Attenuation of Coaxial Cables in the Frequency Range 300–1000 Mc/s by Measurement of the Input and Output Voltage—E. Scheffler and U. Queck. (*Nachrichtentech. Z.*, vol. 9, pp. 60–62; February, 1956.) The method described is suitable for tests on production lengths of cable. Details are included of a simple transformer section for use with specimens whose characteristic impedance differs from that of the test equipment.

621.317.342:621.317.755 2482

Group-Delay Measurements—C. J. Heuvelman and A. van Weel. (*Wireless Engr.*, vol. 33, pp. 107–113; May, 1956.) "A description is given of a simple group-delay meter which, in combination with any conventional wobbulator generator, gives the group-delay characteristic directly on an oscilloscope. An automatic-gain-control circuit, necessary to maintain a constant level at the output of the network under test, enables the tracing of the amplitude characteristic on a second oscilloscope at the same time. Calibration of amplitude and group-delay scales is possible for any oscilloscope used. A sensitivity of 1 mμs can be achieved. The frequency range is 20–45 mc."

621.317.361:621.384.612 2483

Bevatron-Frequency Measurement System—W. M. Brobeck and W. C. Struven. (*Electronics*, vol. 29, pp. 182–187; May, 1956.) Digital counter techniques in conjunction with a cro display are used to monitor the relation of oscillator frequency to magnetic-field intensity during the acceleration period of the bevatron.

621.317.373:621.317.755 2484

Phase-Angle Measurement—C. H. Vincent. (*Wireless Engr.*, vol. 33, pp. 113–117; May, 1956.) A cro method due to Fleming. (*J. IEE*, vol. 63, pp. 1045–1046; November, 1925.) is discussed in which X and Y deflections are adjusted for equality and the phase difference is deduced from the resulting ellipse. Analysis indicates that lack of accuracy due to geometrical factors [1740 of 1953 (Benson)] is avoidable if reasonable precautions are taken.

621.317.42 2485

Methods of Measurement of Magnetic Field—V. Andresciani. (*Ricerca Sci.*, vol. 26, pp. 25–63; January, 1956.) A survey including methods based on the Hall effect in semiconductors and on nuclear magnetic resonance. 60 references.

621.317.42 2486

General Conditions to be satisfied by an Exploring Coil for measuring an Arbitrary Magnetic Field at a Point, giving Fourth-Order Terms—P. Gautier. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1707–1710; March 26, 1956.) The possibility of eliminating the second-order terms with an exploring coil of finite volume is linked with the fact that the induction satisfies the Laplace equation.

621.317.42:621.317.715 2487

Characteristic Properties of Overdamped Galvanometers used as Partial Fluxmeters—É. Selzer. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1422–1425; March 12, 1956.)

621.317.7:621.374.3 2488

A Novel Circuit for Electronic Measurements—Abdel-Halim Ahmed. (*Elect. Engng., N. Y.*, vol. 74, p. 1049; December, 1955.) Digest of paper in *Trans. AIEE*, part I, *Communication and Electronics*, vol. 74, pp. 194–204; May, 1955. A square wave is applied to two tubes operated as cathode followers, in push-pull, so that they are cut off alternately in phase with the ac signal input and at the signal frequency. The resulting average anode current is a measure of $I \cos \phi$, or $V \cos \phi$, where I, V and $\cos \phi$ are the current, voltage and power factor of the circuit measured. The instrument may be adapted to measure any of these quantities, or the power.

621.317.72 2489

A Note on the Theory of Oscillating-Electrode Voltmeters—J. Rawcliffe. (*Proc. IEE*, part A, vol. 103, pp. 55–56; February, 1956.) "It has been tacitly assumed hitherto that the theory of the oscillating-electrode voltmeter derived for direct voltages applies equally well for alternating voltages if rms values are substituted for direct values. A rigourous treatment shows that this is not the case."

621.317.72:621.316.722.4 2490

A Voltage Divider containing a Nonlinear Unit—L. L. Alston. (*Proc. IEE*, part A, vol. 103, pp. 52–54; February, 1956.) Analysis is presented for a voltage divider in which the low-voltage arm is shunted by a SiC resistor; the arrangement was designed primarily for measuring the burning voltage of an arc.

621.317.729:621.372 2491

The Determination of Complicated Wave Fields by means of Multidimensional LC Networks—H. Schneider. (*Nachrichtentech. Z.*, vol. 9, pp. 70–76; February, 1956.) Networks of the general type described by Spangenberg *et al* (3066 of 1949) have been designed at Darmstadt, using cable sections for the individual elements. As an example of use of the model, the determination of the field near a slot antenna is described.

621.396.621.54:621.317.729 2492

New Field Intensity Measuring—Fuse and Soma. (See 2525.)

621.317.737:621.385.029.6 2493

An X-Band Magnetron Q-Measuring Apparatus—J. R. G. Twisleton (*Proc. IEE*, part B, vol. 103, pp. 339–343; May, 1956.) The apparatus described is of the type in which a directional coupler is used in conjunction with a frequency-swept oscillator and cro to monitor the power reflected from the magnetron cavity. The measurement accuracy attainable by this method is discussed.

621.317.763.029.6 2494

The Rod Wavemeter for the Frequency Range 180–80000 Mc/s—Construction and Measurement Results—U. Adelsberger. (*Arch. Elekt. Übertragung*, vol. 10, pp. 51–57; February, 1956.) A type of wavemeter developed at the Physikalisch-Technische Bundesanstalt comprises a coaxial line with sliding inner conductor (rod) of length less than that of the outer conductor; the line is thus open at one end and resonates when its length is an odd multiple of $\lambda/4$. This is advantageous from the point of view of accuracy, since the resonance positions for even generator harmonics do not coincide with those for odd harmonics. A model for mm-λ measurements has the outer conductor built up of quadrantal sections internally silvered and polished.

621.317.784.029.6:621.372.413 2495

A Resonant-Cavity Torque-Operated Wattmeter for Microwave Power—R. A. Bailey. (*Proc. IEE*, part C, vol. 103, pp. 59–63; March, 1956.) The force on a small vane a resonant cavity is a simple function of the Q factor, the power absorbed in the cavity, and the perturbation of the resonance frequency due to the vane. A sensitive measurement method based on these relations is described. Results obtained by this method and by the water-calorimeter method are compared. See also 220 of 1955 (Bailey *et al.*).

621.317.799:621.373.52:621.396.96.001.4 2496

Transistor Generator simulates Radar Target—W. Eckess, J. Deavenport, and K. Sherman. (*Electronics*, vol. 29, pp. 179–181; May, 1956.) Pulse-forming circuits for use in testing radar sets are described; either Ge or Si transistors may be used.

621.317 2497

Precision Electrical Measurements. Proceedings of a Symposium held at the National Physical Laboratory on 17–20 November, 1954 [Book Review]—Publishers: H. M. Stationery Office, London, 1955, 349 pp. (*Brit. J. Appl. Phys.*, vol. 7, p. 82; February, 1956.) The scope of the 26 papers ranges from precision methods of measuring physical properties, such as dielectric constant, to general surveys of testing techniques, such as impulse testing, and includes the appraisal of special properties of new materials, such as the microwave performance of magnetic dielectrics.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621–52 2498

An Introduction to the Analysis of Nonlinear Control Systems with Random Inputs—J. F. Barrett, and J. F. Coales. (*Proc. IEE*, part C, vol. 103, pp. 190–199; March, 1956.) In nonlinear systems with random inputs the probability distribution of the input has to be combined with that of the output in order to obtain that of the error function on which the output depends. When this is attempted, simultaneous integral equations are obtained; approximation methods of dealing with these are discussed. The possibility of using topological methods to investigate the stability of the systems is examined.

621–52 2499

A Three-Dimensional Machine-Tool Control System—(*Electronic Engng.*, vol. 28, pp. 204–207; May, 1956.) Programming instructions taken from design drawings are worked out by an independent digital computer and recorded on magnetic tape which is used in a control unit to operate the machine tool. For measurement purposes an optical diffraction grating system is associated with each plane of the tool and in combination with a photosensitive detector produces a pulse train which is locked to the command pulse train through a servomechanism.

621.317.39+621–52 2500

Electronics in the Process Industries—J. M. Carroll. (*Electronics*, vol. 29, pp. 138–145; May, 1956.) Measurement and control techniques used in chemical, petroleum, and other continuous-flow plants are described.

621.317.39:620.179.1 2501

Prescribed-Function Vibration Generator—P. M. Honnell. (*J. Brit. IRE*, vol. 16, pp. 187–198; April, 1956.) Details are given of an arrangement developed for testing mechanical systems. It comprises a mask profiled to represent the prescribed function (*e.g.*, a square or triangular waveform) rotating in front of a photocell, a vibrating platform being driven by the photocell amplifier.

621.317.79:539.1:538.569.4 2502

The Measurement, by Nuclear Resonance, of Light Water Concentration in Mixtures of Light and Heavy Water—A. M. J. Mitchell and G. Phillips. (*Brit. J. Appl. Phys.*, vol. 7, pp. 67–72; February, 1956.) Two techniques for inducing rf power absorption are described, one suitable for H_2O concentrations in the range 7–100 per cent, the other for low concentrations. In both cases the signal output rises almost linearly with H_2O concentration. A continuous flow method is also described.

621.384.6 2503

Design of the Pole Faces for Circular Particle Accelerators with the Electrolytic Tank—F. Amman and L. Dadda. (*Nuovo Cim.*, vol. 3, pp. 184–187; January 1, 1956. In English.)

621.385.833 2504

On the Magnification and Resolution of the Field-Emission Electron Microscope—D. J. Rose. (*J. Appl. Phys.*, vol. 27, pp. 215–220; March, 1956.) Analysis of electron trajectories indicates that the existence of small protrusions on the field—emission tip may give rise rise to local areas of enhanced magnification; a resolution of 3 Å may be attained at these regions, so that some of their atomic detail should be observable.

621.385.833 2505

A Favorable Condition for Seeing Simple Molecules in a Field-Emission Microscope—J. A. Becker and R. G. Brandes. (*J. Appl. Phys.*, vol. 27, pp. 221–223; March, 1956.) Experimental evidence is presented confirming the analysis developed by Rose (2504 above).

621.385.833 2506

General Expressions and Typical Curves of Electro-Optical Characteristics of Magnetic Electron Lenses—P. Durandeau. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1710–1712; March 26, 1956.)

621.385.833 2507

Phase Contrast and Interchromatic Contrast: New Observation Methods in Electron Microscopy—M. Locquin. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1713–1716; March 26, 1956.) Contrast is considerably increased by using an objective diaphragm with thinned edge in conjunction with illumination by a hollow cone of electrons.

621.385.833 2508

Investigation of Cylindrical Magnetic Lenses with Iron Armouring—V. M. Kel'man and S. Ya. Yavor. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1405–1411; August, 1955.)

621.385.833 2509

Electron-Optical Equations for Wide Beams, taking Account of Chromatic Aberrations, and their Application to the Investigation of the Motion of Particles in Axially Symmetrical Fields—Yu. V. Vandakurov. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1412–1425; August, 1955.)

621.386.8:621.383.2:620.179.1 **2510**
Intensification of the X-Ray Image in Industrial Radiology—A. Nemet and W. F. Cox. (*Proc. IEE*, part B, vol. 103, pp. 345–355; May, 1956. Discussion, pp. 355–359.) The influence of brightness, blurring, and contrast on the image resolution is examined. The improvement obtained by use of the image intensifier tube [1098 of 1953 (Teves and Tol)] is indicated.

621.387.424 **2511**
Analysis of Spuriousness of Geiger-Muller Tubes at High Temperatures—S. P. Puri and P. S. Gill. (*Indian J. Phys.*, vol. 30, pp. 1–9; January, 1956.)

621.383 **2512**
Die Anwendung der Photozellen [Book Review]—P. Goerlich. Publishers: Akademische Verlagsgesellschaft Geest and Portig, K. G., Leipzig, 468 pp., 1954. (*J. Opt. Soc. Amer.*, vol. 46, pp. 73–74; January, 1956.) Comprises six chapters, dealing with the properties of photocells, the basic associated circuitry, applications for switching and control operations, photometry, and recording, sound reproduction, and image reproduction.

PROPAGATION OF WAVES

621.396.11 **2513**
Estimating the Ratio of Steady Sinusoidal Signal to Random Noise from Experimental Data—M. L. Phillips. (Proc. IRE, vol. 44, p. 692; May, 1956.) A family of curves is presented and a procedure is indicated facilitating determination of the required ratio, which is of particular interest in studies of scatter propagation, from observed values of the envelope of received field intensities.

621.396.11:551.510.535:523.78 **2514**
Ionospheric Observations at Banaras during the Total Solar Eclipse on 20 June 1955—D. K. Banerjee, P. G. Surange, and S. K. Sharma. (*J. Sci. industr. Res.*, vol. 14A, pp. 517–521; November, 1955.) Field-strength measurements were made on signals from Colombo on 30.51 m λ and from Delhi on 30.71 m λ; vertical-incidence measurements were also made. The results are shown graphically and discussed; they indicate that the ionospheric eclipse lasted longer than the optical eclipse. An increase of the strength of the Colombo signals during the eclipse is attributed to reduction of absorption in the lower layers, while a reduction of absorption in the lower layers, while a reduction of the strength of the signals from Delhi is attributed to the reduced reflection coefficient associated with lowering of the electron density.

621.396.11.029.55:551.510.535 **2515**
Directional Observations on H.F. Transmissions over 2 100 km—E. N. Bramley. (*Proc. IEE*, part B, vol. 103, pp. 295–300; May, 1956.) Direction-of-arrival measurements have been made using a wide-aperture spaced-loop direction finder. Pulse transmissions were used for most of the experiments and first- and second-order F-reflections could usually be identified at 11 mc. The bearing fluctuations of these echoes included a lateral-deviation component of about the magnitude expected from previous experiments at shorter distances. The rapid fluctuations were appreciably larger than at 700 km (3389 of 1955); night-time observations on 5 mc indicated a standard deviation of 1.5° for individual bearings in an hourly period. The corresponding figure for the 1F echo in the daytime on 11 mc was only 0.6°. The results were unaffected by changes made in the transmitting antennas and by ionospheric and magnetic storms.

621.396.11.029.62:535.5 **2516**
Observations of Short Bursts of Signal from a Distant 50-Mc/s Transmitter—B. H. Briggs. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 171–183; March, 1956.) Analysis of reception over a distance of about 500 km strongly suggests that the majority, if not all, of the signals are reflected from meteor trails. In many cases the reflection point appears to be near the receiver rather than at the midpoint of the transmission path.

621.396.11.029.62:551.510.52 **2517**
Abnormal V.H.F. Propagation—A. H. Hooper. (*Wireless World*, vol. 62, pp. 295–298; June, 1956.) A method is indicated for constructing a graph showing variations of radio refractive index of the troposphere with height, using data derived from the daily meteorological observations of air pressure, dew point, and temperature.

621.396.81.029.62:621.396.3 **2518**
V.H.F. Trans-horizon Communication Techniques—Ringoen and Smith. (See 2532.)

621.396.812.3:551.510.535 **2519**
The Fading of Radio Waves of Frequencies between 16 and 2400 kc/s—S. A. Bowhill. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 129–145; March, 1956.) Nighttime experiments using continuous waves reflected from the lowest ionosphere at nearly vertical incidence are discussed. Shallow fading with a quasi-period of 7 minutes is found in the frequency range 16–70 kc; above 70 kc fading with a quasi-period of about 1 minute appears, increasing in depth with frequency. Ionosphere irregularities corresponding to ground dimensions of 5 km and 1 km, respectively, are indicated. The slow fading is due mainly to random ionospheric velocities of about 40 mps, superimposed on which is a smaller, very variable, drift velocity. A D-region model is suggested.

621.396.812.3:551.510.535 **2520**
The Fading Periods of the E-Region Coupling Echo at 150 kc/s—R. W. Parkinson. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 158–162; March, 1956.) From analysis of echoes returned from the coupling region lying below the main E layer, two dominant distributions of the fading periods have been found, centered on periods of 2 and 7 minutes. Possible causative mechanisms are discussed.

RECEPTION

621.376.33 **2521**
Theory of Detection of Frequency-Modulated Oscillations—V. K. Turkin and G. A. Levin. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 999–1002; February 21, 1956. In Russian.) The transmission of a fm signal through a tuned circuit or band-pass filter followed by a square-law detector is considered using new functions which are defined by series involving Bessel functions. The af output is calculated for a particular case.

621.396.3:621.396.82 **2522**
Waveform of Radiotelegraph Signals and Interference between Adjacent Channels—J. Marique. (*Ann. Télécommun.*, vol. 11, pp. 26–32; February, 1956.) The problem is attacked from the same general viewpoint as previously (3396 of 1955), consideration being given to the further case of pulses whose flanks are halves of Gaussian curves. The results indicate that no particular advantage accrues from use of this waveform.

621.396.621:621.376.3:621.396.82 **2523**
Impulse Noise in Narrow-Band F.M. Receivers—S. P. Lapin and J. J. Suran. (*Elect. Engng., N.Y.*, vol. 74, p. 1091; December, 1955.) Digest of paper in *AIEE*, part I, *Communication and Electronics*, vol. 74, pp. 450–454; September, 1955.) Loss of signal due to noise transients is minimized by rounding the band-pass characteristic of the receiver so that, after the initial maximum, the amplitude of the transient decays more rapidly than in filters having a flat band-pass response curve.

621.396.621.029.62:621.376.332 **2524**
Unconventional F.M. Receiver—M. G. Scroggie. (*Wireless World*, vol. 62, pp. 258–262; June, 1956.) Complementary details are given of the receiver embodying the pulse-clunter discriminator described previously (1861 of 1956). A single intermediate frequency of the order of 150 kc is used, to suit the discriminator.

621.396.621.54:621.317.729 **2525**
New Field Intensity Measuring—S. Fuse and S. Soma. (*Rep. Elect. Commun. Lab., Japan*, vol. 3, pp. 55–58; October, 1955.) The sensitivity of a double-heterodyne receiver used for measuring field strength and spurious signals is improved by narrowing the bandwidth. The problem of local oscillator frequency stability is dealt with by sweeping the frequency of the second oscillator.

621.396.621.54:621.398 **2526**
Telemetering Receiver conserves Bandwidth—M. S. Redden, Jr., and H. W. Zancanata. (*Electronics*, vol. 29, pp. 174–178; May, 1956.) A crystal-controlled double-superheterodyne receiver covering the range 216–247 mc is described. Two second-IF amplifiers are provided, one with a bandwidth of 100 kc, for reception of pwm/fm data, and the other with a bandwidth of 500 kc, for fm/fm data.

621.396.822 **2527**
Methods of solving Noise Problems—W. R. Bennett. (Proc. IRE, vol. 44, pp. 609–638; May, 1956.) "A tutorial exposition is given of various analytical concepts and techniques of proved value in calculating the response of electrical systems to noise waves. The relevant probability theory is reviewed with illustrative examples. Topics from statistics discussed include probability density, moments, stationary and ergodic processes, characteristic functions, semi-invariants, the central limit theorem, the Gaussian process, correlation, and power spectra. It is shown how the theory can be applied to cases of noise and signal subjected to such operations as filtering, rectification, periodic sampling, envelope detection, phase detection, and frequency detection."

621.396.823 **2528**
Limits for Radio Interference in Germany and Other Countries—G. Use. (*Elektrotech. Z., Edn A*, vol. 77, pp. 33–40; January 11, 1956.) The methods of measurement of radio interference and the recommended limits are summarized for nine European countries, North America, and Japan. 19 references.

STATIONS AND COMMUNICATION SYSTEMS

621.376.2:621.3.012.1 **2529**
A Vector Method for Amplitude-Modulated Signals—C. J. N. Candy. (*Proc. IEE*, part B, vol. 103, pp. 410–415; May, 1956. Discussion, pp. 415–418.) Notation and analysis are presented facilitating calculations of the output of a demodulator. Examples show that when modulation frequency is comparable with carrier frequency, the modulation (or signal) may be appreciably modified by impedances in the carrier channel. A complementary operational method for determining the response of circuits to nonsinusoidal am signals is also described.

621.39.001.11 **2530**
Some Terminology and Notation in Information Theory—I. J. Good. (*Proc. IEE*, part C, vol. 103, pp. 200–204; March, 1956.)

621.39.001.11 **2531**
The Arithmetical Characterization of Messages, and its Use for determining Correcting

Networks—M. Bayard and R. Roquet. (*Ann. Télécommun.*, vol. 11, pp. 33–45; February, 1956.) The message, or "modulation phrase," is uniquely characterized by an "index-number"; in the case of two-position telegraphy, which is examined in detail, this is given by $I=\sum^{n}_{n=0}\alpha_n 2^n$, where n is the number of individual signals making up the message and α_n is the amplitude of an individual signal. For a given transmission system a curve can be traced showing the distortion of a reference signal at zero time as a function of I; from such curves, which take account of the effect of future signals, the appropriate correcting networks can be determined for obtaining a desired modification of the "modulation phrase."

621.396.3:621.396.81.029.62 2532

V.H.F. Trans-horizon Communication Techniques—R. M. Ringoen and J. W. Smith. (*Electronics*, vol. 29, pp. 154–159; May, 1956.) Possible techniques are reviewed with emphasis on the equipment requirements. It should be feasible to establish teleprinter links over distances up to 1500 miles, in auroral regions, with reliability better than 99 per cent.

621.396.3.029.55 2533

Observations and Experience at the Frankfurt a. Main Radio Exchange with the TOM (Teletype on Multiplex) Equipment operated on Short-Wave Transmission Paths—M. Corsepius and K. Vogt. (*Nachrichtentech, Z.*, vol. 9, pp. 55–59; February, 1956.) The observations indicate the improvements obtained by use of an automatic error-correction system. See also 2507 of 1954 (Hayton *et al.*).

621.396.41:621.376.5:621.396.65 2534

Pulse-Time-Modulation Terminals for Music Transmission over Radio Links—R. F. Rous. (*Proc. IEE*, part B, vol. 103, pp. 283–292; May, 1956. Discussion, pp. 292–294.) A microwave link intended primarily for television is used alternatively to accommodate three music circuits and one engineer's circuit. Measurements indicate that signal/noise and signal/crosstalk ratios are satisfactory; operating over a 25-mile path there was a 20-db margin in the signal/noise ratio over the value recommended by the CCIR The multiplexing circuit is described and performance figures are given.

621.396.65:621.396.41 2535

Radio-link Network—S. Montagnani. (*Poste e Telecommunicazioni*, vol. 23, pp. 884–894; December, 1955.) Account of the Bologna-Pisa trans-Appennine extension to the network described previously [2423 of 1955 (Bernardi)].

621.396.65.029.62:621.396.41 2536

Very-High-Frequency Radio Link between Puerto Rico and the Virgin Islands—R. McSweeny. (*Elect. Commun.*, vol. 32, pp. 238–247; December, 1955. *Trans. AIEE*, part I, *Communication and Electronics*, vol. 74, pp. 781–785; January, 1956.) Illustrated description of a two-stage link system with fm operation at frequencies between 150 and 160 mc, providing several telephone and teleprinter channels.

621.396.65.029.63:621.396.41 2537

Microwave Relay System between Saint John and Halifax—H. C. Sheffield. (*Elect. Commun.*, vol. 32, pp. 214–236; December, 1955. IRE TRANS., vol. CS-4, pp. 144–167; May, 1956.) Detailed account of a time-division-multiplex radio link with five intermediate unattended repeaters. The 2-kmc transmission band is used. There are two multiplex equipments, each providing 23 voice channels, and three sets of rf equipment. Possible causes of interference are discussed and some performance figures are given.

621.396.71 2538

The New High-Frequency Transmitting Station at Rugby—C. F. Booth and B. N. MacLarty. (*Proc. IEE*, part B, vol. 103, pp. 263–278; May, 1956. Discussion, pp. 278–282;) Comprehensive description of this British Post Office radio-communication station. A short account was abstracted previously (569 of 1956).

621.396.932+621.396.96 2539

Modernisation of Radio and Radar Equipment in H. M. Telegraph Ships—W. Dolman and P. W. J. Gammon. (*P.O. Elect. Engrs.' J.*, vol. 48, Part 4, pp. 204–207; January, 1956.)

621.396.975:621.395.623.8 2540

Wireless Sound Systems—Hargens. (See 2278.)

SUBSIDIARY APPARATUS

621.311.6:537.311.33:[546.28+546.289 2541

The Electron-Voltaic Effect in Germanium and Silicon P-N Junctions—Rappaport, Loferski, and Linder. (See 2437.)

621.311.6:621.383.5:539.165 2542

The Effect of Radioactive Radiation on a Photocell—Pasynkov. (See 2567.)

621.311.6:621.39 2543

Automatic Control of Power Equipment for Telecommunications and other Essential Services—A. Watkins. (*J. Brit. IRE*, vol. 16, pp. 227–238; April, 1956.) "Two types of no-break generating sets are described: an all-electric battery operated equipment, and a diesel electric equipment. Detailed descriptions are given of three electronic devices used with these sets: a) a static exciter automatic voltage regulator making use of two saturated transductors; b) an alternator synchronizer in which the generator and mains voltages are compared in a triode circuit; c) speed regulator using a thyratron which feeds the control field of the dc motor and also incorporates alarm and protection devices."

621–526 2544

Servomechanism Analysis [Book Review]—G. J. Thaler and R. G. Brown. Publishers: McGraw-Hill Book Co., Inc., New York and London, 414 pp., 1953. (*Nature, Lond.*, vol. 177, p. 766; April 28, 1956.) Gives the mathematical background for the analysis and design of servomechanisms, including a clear introduction to the theory of Laplace transformations.

TELEVISION AND PHOTOTELEGRAPHY

621.397.241 2545

The Provision of Circuits for Television Outside Broadcast—M. B. Williams and J. B. Sewter. (*P.O. Elect. Engrs.' J.*, vol. 48, parts 3 & 4, pp. 166–170; October, 1955, and pp. 234–238; January, 1956.) Technique and equipment for transmitting video signals over telephone pairs is described, including a new video repeater [see also 275 of 1956 (Sewter and Wray)] and equipment for injecting vision signals into main coaxial-cable television links operated in the frequency ranges 3–7 mc and 0.5-4 mc.

621.397.26:621.396.65 2546

Microwave Television Radio Relay System—O. H. Appelt, K. Christ, and K. Schmid. (*Elect. Commun.*, vol. 32, pp. 248–254; December, 1955.) English version of paper originally published in German (538 of 1954).

621.397.5 2547

Phonevision—an Effective Method for Subscription Television—A. L. C. Webb and A. Ellett. (*J. Brit. IRE*, vol. 16, pp. 205–219; April, 1956.) Paper reprinted from PROC. IRE, *Aust.*, vol. 16, pp. 341–353; October, 1955.

621.397.5:535.623 2548

The Principles of N.T.S.C. Color Television—C. J. Hirsch. (*J. IEE*, vol. 2, pp. 89–97; February, 1956.)

621.397.5:778.5 2549

Television Studio Practices relative to Kinescope Recording—H. Wright. (*J. Soc. Mot. Pict. Telev. Engrs.*, vol. 65, pp. 1–6; January, 1956.) The problem of maintaining the correct brightness levels is considered in relation to the signal waveform.

621.397.5(083.74) 2550

Fundamentals of the Television Standards adopted in Uruguay—M. Giampietro. (*Rev. Telegr. Electronica*, vol. 44, pp. 11–12, 16; January, 1956.) The various considerations such as mains frequency, international interchangeability, requirements for color operation, and magnetic recording, involved in the choice of standards are indicated. The standards adopted are 525 lines 60 fields for monochromatic operation and NTSC standards for color, in accordance with FCC recommendations except in respect of the difference between maximum and minimum transmitter levels and the channel spacings.

621.397.61:621.318.57 2551

Electronic Switches for Television—A. M. Spooner. (*Electronic Engng.*, vol. 28, pp. 196–199; May, 1956.) Circuits for switching between video signals are discussed. The causes of switching transients are indicated and a transient-free switch is described in which a cathode-coupled triode pair, with a pentode for the common cathode impedance, feeds a series-connected double diode.

621.397.611.2 2552

The Problem of Inertia Effects in the Vidicon—W. Heimann. (*Arch. Elekt. Übertragung*, vol. 10, pp. 73–76; February, 1956.) Inertia effects were shown previously (2451 of 1955) to be due to photoelectric inertia in the target and to the nature of the charge-storage and signal-generating mechanisms, the latter being the more important. Experiments were made using target layers of different thicknesses and different scanned areas; the results confirmed the importance of obtaining the appropriate value for the capacitance of the picture elements. Measurements were made of the layer thickness by means of an interference microscope, and values of the capacitance were hence determined.

621.397.7 2553

Television Satellite Systems—C. B. Plummer. (IRE TRANS., vol. BTS-1, pp. 65–66; March, 1955.) A brief discussion of television coverage problems in the U.S.A. precedes separate papers dealing with particular installations, as follows:—

U.H.F. Satellite Transmitter-Receiver Design and Operation—L. Katz and T. B. Friedman abstract, PROC. IRE, (pp. 67–74; vol. 43, p. 640; May, 1955.)

The Engineering Aspects of a U.H.F. Booster Installation—J. Epstein (pp. 75–80; abstract as above).

A Report on U.H.F. Satellite Operation—J. R. Whitworth (pp. 81–82).

An Experimental On-Channel Satellite Booster System—J. H. DeWitt, Jr, G. A. Reynolds, and L. E. Rawls (pp. 83–102).

621.397.7:535.623:621.3.06 2554

Color Video Switching—W. B. Whalley and R. S. O'Brien. (*J. Soc. Mot. Pict. Telev. Engrs.*, vol. 65, pp. 16–19; January, 1956.) Precautions taken to minimize the effect of capacitances in the studio switching equipment on the phase and amplitude response to the chrominance signal are described.

621.397.743:621.317.2 2555

The New Independent-Television Network—S. H. Granger. (*P.O. Elect. Engrs., J.*, vol. 48, part 4, pp. 191–197; January, 1956.) The structure of the system is outlined, with descriptions of the London studio links and intercity networks and details of over-all performance.

TRANSMISSION

621.396.61:621.375.232 2556

A Method of deriving Overall Negative Feedback Voltage in Transmitters—D. Smart. (*J. Brit. IRE*, vol. 16, pp. 221–223; April, 1956.) Greater ease of frequency changing, greater linearity and reduced phase shift, drift, and cost result if the feedback voltage is derived from a resistor in the earth return of the power amplifier tube rather than by the conventional method of rectifying part of the modulated output.

TUBES AND THERMIONICS

621.3.011:621.396.822 2557

Physical Sources of Noise—Pierce. (See 2308.)

621.314.63:546.289:621.396.822 2558

Measurement of Noise Spectra of a Germanium p-n Junction Diode—F. J. Hyde. (*Proc. Phys. Soc.* vol. 69, pp. 231–241; February 1, 1956.) "Measurements have been made of the excess noise generated in a fused-alloy-type *p-n* junction diode in the frequency range 0.12 cps to 2 mc at 29°C. with the reverse direct current I as parameter. The observed noise spectral density may be synthesized from three well-defined types of component: a) an extensive component proportional to f^{-1} directly observed over five decades, and by synthesis assumed to exist over as many as seven decades of frequency; b) a component proportional to $(1+\omega^2\tau_1^2-1)$ associated with a single relaxation time τ_1; c) a uniform component associated with the "shot" noise of the measured current I, for frequencies less than $1/2\pi\tau_p$ where τ_p is the hole lifetime. τ_1 was found to increase with increasing I while the intensity of the f^{-1} component increased more rapidly than in proportion to I^2."

621.314.63:621.319.4 2559

A Variable-Capacitance Germanium Junction Diode for U.H.F.—L. J. Giacoletto and J. O'Connell. (*RCA Rev.*, vol. 17, pp. 68–85; March, 1956. *RCA Special Publication, Transistors I*, pp. 221–238; 1956.) The significance of various design factors for the preformance of alloy-junction diodes is considered with reference to use of the diodes as voltage-controlled capacitors when biased in the reverse direction. For a particular diode with a bias of 6 v the capacitance was 38 $\mu\mu$f and the voltage variation of capacitance was 3$\mu\mu$f/v; the lead inductance was 2.6 mμh and the effective series resistance 0.5 Ω. The Q value was very high at the lower radio frequencies, decreasing to 17 at 500 mc.

621.314.632:546.817.221 2560

Analysis of H_2S-Treated PbS Point-Contact Rectifiers—V. G. Bhide, J. N. Das, and P. V. Khandekar. (*Proc. Phys. Soc.*, vol. 69, pp. 245–248; February, 1956.) Measurements were made on natural and synthetic crystals originally of *n*-type but converted to *p*-type, to a certain depth, by heating in a H_2S atmosphere, and used in combination with various metal points. The values obtained for α, the slope of the semilog plot of the I/V characteristic, are near the theoretical value of 40 V^{-1}, whereas the values of α for the original *n*-type material are considerably lower. The significance of the results is discussed briefly.

621.314.7 2561

Some Experiments on, and a Theory of, Surface Breakdown [in transistors]—C. G. B. Garrett and W. H. Brattain. (*J. Appl. Phys.*, vol. 27, pp. 299–306; March, 1956.) The experiments were performed on *n-p-n* alloy junction transistors as used by Wahl and Kleimack (2243 of 1956). The procedure followed was to measure photocurrent point by point for voltages first below and then above breakdown value. Results indicate that surface breakdown, like body breakdown, is an avalanche process; the multiplication sets in at a particular spot. High breakdown voltage is encouraged by arranging that the polarity of the surface charge is such as to produce a "channel" over the material of the higher-resistivity side, and by surrounding the unit with a medium of high dielectric constant.

621.314.7 2562

Uniform Planar Alloy Junctions for Germanium Transistors—C. W. Mueller and N. H. Ditrick. (*RCA Rev.*, vol. 17, pp. 46–56; March, 1956. *RCA Special Publication, Transistors I*, pp. 121–131; 1956.) Technique is described for obtaining flat junctions, by separating the wetting from the alloying steps. As a consequence, the upper limit on operating frequency is considerably raised.

621.314.7 2563

P-N-P Transistors using High-Emitter-Efficiency Alloy Materials—L. D. Armstrong, C. L. Carlson, and M. Bentivegna. (*RCA Rev.*, vol. 17, pp. 37–45; March, 1956. *RCA Special Publication, Transistors I*, pp. 144–152; 1956.) "The addition of small percentages of gallium or aluminum to indium, for use as the emitter alloy, produces greatly improved high-current characteristics. As compared with pure indium, the use of gallium alloys improves emitter efficiency by about 3.5 times, and the use of aluminum-bearing alloys by about 10 times. Techniques for preparation of the alloys and results of tests on transistors using the various emitters are described. Volume lifetime is measured as a function of injection level to permit comparison with the theoretical equations for current amplification factor. These measurements are discussed briefly, and a revised equation for current amplification factor at high currents is given."

621.314.7:537.311.33:546.289 2564

Surface Treatment of Silicon for Low Recombination Velocity—Moore and Nelson. (See 2439.)

621.383.4:546.289:621.396.822 2565

Technique for Improving the Signal-to-Noise Ratios of Single-Crystal Photoconductive Detectors—R. M. Page, R. W. Terhune, and J. Hickmott, Jr. (*J. Appl. Phys.*, vol. 27, pp. 307–308; March, 1956.) Correlation technique adapted from that used by Montgomery (122 of 1953) is briefly described. The detector is a Ge filament, and the light source a neon lamp modulated at 1 kc.

621.383.5:537.311.33 2566

Photovoltaic Effect in GaAs *p-n* Junctions and Solar Energy Conversion—D. A. Jenny, J. J. Loferski, and P. Rappaport. (*Phys. Rev.*, vol. 101, pp. 1208–1209; February 1, 1956.) Measurements on several cells are reported and the results are compared with theoretical predictions. The highest value obtained for the efficiency of conversion of solar energy is 6.5 per cent, which is of the same order as values obtained for Si and CdS, but higher values are to be expected as the technique of preparing the cells is improved.

621.383.5:539.165:621.311.6 2567

The Effect of Radioactive Radiation on a Photocell—V. V. Pasynkov. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1376–1385; August, 1955.) Experiments conducted on a Se barrier-layer photocell indicate that the effect of β-radiation is similar to that of a light beam. It is suggested that with the aid of artificial radioactive isotopes small low-power electric batteries can be constructed which would not be damaged by short circuits and which would have a very long working life.

621.385:537.533 2568

Instability of Electron Beams subjected to a Magnetic Field—B. Epsztein. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1425–1428; March 12, 1956.) Analysis is presented to explain the breaking-up of a tubular beam observed by Webster (*J. Appl. Phys.*, vol. 26, pp. 1386–1387; November, 1955.) Instability can also occur without a magnetic field. The energy brought into play is derived entirely from the mutual repulsion of the electrons.

621.385.001.4 2569

A Simple Optical Method for determining Grid-Cathode Spacing in Electronic Valves—W. Guber and W. Stetter. (*Nachrichtentech. Z.*, vol. 9, pp. 77–79; February, 1956.)

621.385.029.6 2570

The Design of High-Power Traveling-Wave Tubes—M. Chodorow and E. J. Nalos. (PROC. IRE, vol. 44, pp. 649–659; May, 1956.) The design of tubes for pulsed powers of the order of 1 mw is discussed. Some success has been achieved with an experimental tube operating in the 3-kmc band, using a disk-loaded waveguide for slowing the wave. From results of measurements on this tube it is inferred that bandwidths of 10–20 per cent should be attainable with reasonable efficiency. Modifications in attenuator, focusing and coupling systems likely to lead to improved performance are indicated.

621.385.029.6 2571

A Large-Signal Theory of Traveling-Wave Amplifiers—P. K. Tien. (*Bell Syst. Tech. J.*, vol. 35, pp. 349–374; March, 1956.) The work of Nordsieck (2497 of 1953) and Tien *et al.* (1822 of 1955) is extended to take fully into account the effects of space-charge repulsion and of a finite coupling between electron beam and circuit. The energy associated with the backward wave is calculated and its effect on efficiency is discussed. Assuming typical values for the various parameters, values ranging from 23 per cent to 40 per cent are found for the saturation efficiency. Curves and tables are presented showing the voltage and phase of the circuit wave, the velocity spread of electrons and the fundamental component of the charge-density modulation.

621.385.029.6 2572

Travelling-Wave Tubes—F. N. H. Robinson. (*Research, Lond.*, vol. 9, pp. 27–31; January, 1956.) Recent developments are briefly surveyed; space-periodic beam-focusing arrangements, nonreciprocal attenuators, improvements in noise figure, and backward-wave tubes are mentioned. Over 30 references.

621.385.029.6 2573

Phase-Angle Distortion in Traveling-Wave Tubes—W. R. Beam and D. J. Blattner. (*RCA Rev.*, vol. 17, pp. 86–99; March, 1956.) A theoretical study is made of variations of phase velocity with supply voltages, signal level, and matching conditions; Pierce's first-order theory (*Traveling-Wave Tubes*, 1950) is used. The results are in good agreement with measurements on a tube operating at 3 kmc, in which the following phase shifts were observed: 50° for 1 per cent change in helix voltage; 2° for 1 per cent change in first-anode voltage; about 0.1(for 1 per cent change in magnet current; 6° for each mw increase of output power; 6° for each mw of power lost by reflection at the output coupler.

621.385.029.6 2574
Bihelical Traveling-Wave Tube with 50-dB Gain at 4000 Mc/s—W. P. G. Klein. (*Elect. Commun.*, vol. 32, pp. 255–262; December, 1955.) A 5-w tube suitable for microwave relay systems is discussed. In order to maintain the gain as near as possible to the small-signal value, two helices are provided, having separate dc circuits and coupled only by the beam; an attenuating section is arranged between the two helices, the second of which is shorter than the first. Measurements on several experimental tubes are reported.

621.385.029.6 2575
On the Space Charge affected by the Magnetic Field—Y. Yasuoka. (*J. Phys. Soc. Japan*, vol. 10, pp. 1102–1109; December, 1955.) A study has been made of effects due to dense space charge constrained by a strong magnetic field. Special experimental tubes were used, with a tungsten-filament cathode and copper-block anode, and the cathode back-heating was measured in the absence and in the presence of the magnetic field. The measurements confirm the theory that the outermost electrons of the space-charge cloud are scattered by mutual interaction and have excess energies, thus giving rise to the back heating. When the anode voltage is made sufficiently high, these electrons are captured and oscillations, probably of plasma type, occur.

621.385.029.6 2576
Development of Traveling-Wave Tubes for 4000-Mc/s Band—K. Sato, D. Kobayashi, A. Kondo, and J. Koyama. (*Rep. Elect. Commun. Lab., Japan*, vol. 3, pp. 11–16; October, 1955.) Details are given of the design, construction and performance of several helix-type tubes developed in Japan.

621.385.029.6 2577
The Cascade-Bunching of Electrons in Application to the Theory of the Multi-resonator Magnetron—V. N. Shevchik. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1462–1470; August, 1955.) Attempts have been made in the literature to consider the operation of a magnetron from the standpoint of the interaction of the electron stream with the hf fields localized directly at the anode slots, as distinct from the method of the traveling wave; this method is developed further. A detailed analysis of the operation of a magnetron with a "thick" cathode is given, the results are in good agreement with experimental data.

621.385.029.6:621.317.737 2578
An X-Band Magnetron Q-Measuring Apparatus—Twisleton. (See 2493.)

621.385.029.6:621.373.423 2579
A Reflex-Klystron Oscillator for the 8–9-mm Band—D. J. Wootton and A. F. Pearce. (*Proc. IEE*, part C, vol. 103, pp. 104–111; March, 1956.) Detailed description of the Type-VX5023 klystron; it is continuously tunable over the wavelength range 8–9 mm and is suitable for use in superheterodyne receivers and as a source for laboratory measurements.

621.385.029.65 2580
A Millimetre-Wave Magnetron—J. R. M. Vaughan. (*Proc. IEE*, part C, vol. 103, pp. 95–103; March, 1956.) A detailed description is given of a fixed-tuned rising-sun magnetron Type-VX5027, for high-power pulse operation at 8.6 mm λ, designed in accordance with established principles. Design problems, performance, and test procedure are discussed.

621.385.032.2:537.533 2581
A Detailed Analysis of Beam Formation with Electron Guns of the Pierce Type—W. E. Danielson, J. L. Rosenfeld, and J. A. Saloom. (*Bell Syst. Tech. J.*, vol. 35, pp. 375–420; March, 1956.) The theory of Cutler and Hines (2154 of 1955) is extended to cover the case when spread caused by thermal electron velocities is not small compared with nominal beam size; a lens correction for the finite size of the anode aperture is worked out. Charts are presented facilitating the choice of design parameters to produce a prescribed beam and experimental results confirming the theory are given.

621.385.032.21 2582
Some New Thermionic Cathodes—F. A. Vick. *Sci. Progr.*, vol. 44, pp. 65–71; January, 1956.) A brief progress review.

621.385.032.213.1 2583
Mutual Heating in Transmitting-Valve Filament Structures—W. J. Pohl. (*Proc. IEE*, part C, vol. 103, pp. 224–230; March, 1956.) Methods of calculating the effects of mutual heating between individual elements are described. A set of universal curves to facilitate application of the results to cylindrical structures is presented.

621.385.032.216 2584
Cathode Interface Impedance Desimplified—H. B. Frost. IRE TRANS., vol. RQC-5, pp. 27–33; April, 1955. Abstract, PROC. IRE, vol. 43, p. 896; July, 1955.) To represent the cathode-interface impedance accurately, a RC network containing four elements is required. Gradual and sudden tube failures due to the growth of this impedance are discussed in relation to system reliability.

621.385.032.216 2585
Conduction Mechanism in Oxide-Coated Cathodes—F. B. Hensley. (*J. Appl. Phys.*, vol. 27, pp. 286–290; March, 1956.) "Measurements have been made on a system composed of two parallel planar cathodes so arranged that their surfaces may be pressed together or separated by a small gap. Low-field conductivity measurements show that above approximately 700°K, the conductance of the system does not depend on physical contact between the cathode surfaces. This result supports the theory that the high-temperature conductivity is a property of the electron gas in the cathode pores. The ratio of conductivity to thermionic emission was measured under conditions designed to preserve the state of activation of the cathode surface. The results agreed with the theoretically predicted ratio and demonstrate that the higher values previously reported were caused by a lower activation on the surface than in the interior of the cathode."

621.385.032.216 2586
Equivalent Circuit for the Oxide-Coated Cathode—R. E. J. King. (*J. Appl. Phys.*, vol. 27, pp. 308–309; March, 1956.) A new interpretation is presented of the equivalent circuit proposed by Tomlinson (3409 of 1954) to represent the two-mechanism conduction process in the cathode coating.

621.385.032.7 2587
Electron Bombardment of the Glass Envelope of a Receiving Valve—G. H. Metson and D. J. Sargent. (*Proc. IEE*, part B, vol. 103, pp. 334–338; May, 1956.) On being bombarded by electrons, the potential of the tube envelope is driven towards cathode potential if its initial value is such that the secondary emission coefficient δ is <1, and towards anode potential is δ is >1; this latter condition, termed "lock-up," is stable. Experimental and theoretical investigations for a pentode with an oxide cathode indicate that the mechanisms by which the envelope acquires the initial positive potential corresponding to $\delta>1$ include anode-voltage leakage, capacitive transfer of anode voltage, and photoelectric emission.

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A Grounded-Grid Valve System with High Stability Characteristics—F. Exley and R. E. Young. (*Electronic Engng.*, vol. 28, pp. 202–203; May, 1956.) An arrangement suitable for an oscillator or rf amplifier using a "lighthouse" type tube is described in which the anode and cathode coaxial lines extend on either side of the grid plane.

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Some Half-Tone Charge-Storage Tubes—R. S. Webley, H. G. Lubszynski, and J. A. Lodge. (*Proc. IEE*, part B, vol. 103, pp. 395–397; May, 1956.) Discussion on 3453 of 1955.

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Progress in the Development of Post-Acceleration and Electrostatic Deflection—K. Schlesinger. (PROC. IRE, vol. 44, pp. 659–667; May, 1956.) Practical forms of the single-deflection-center "deflectron" system (3290 of 1952) are formed by printing the electrodes on the insides of glass cylinders or cones. A post-acceleration, or intensifier, system suitable for use with this deflection system comprises a long drift space terminated by a double metal mesh close up to the screen, generally as described by Allard (1032 of 1951). Undesired secondary emission from this mesh is suppressed by providing an insulating coating on the first surface of the mesh while connecting the support metal to a potential lower than that of the drift space.

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Transistors and Other Crystal Valves [Book Review]—T. R. Scott. Publishers: Macdonald and Evans, London, 1955, 258 pp., (*Brit. J. Appl. Phys.*, vol. 7, pp. 82–83; February, 1956.) "This book has been written for engineers who may have to use crystal tubes, rather than for those engaged in developing or designing them."

MISCELLANEOUS

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Transactions of the IRE 1955 Index—(PROC. IRE, vol. 44, 24 pp. following p. 732; May, 1956.) Includes contents lists, author and subject indexes and "nontechnical index."

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Reliability in Complex Electronic Equipment—G. H. Scheer. (*Elect. Engng., N.Y.*, vol. 74, pp. 1062–1065; December, 1955.) A statistical study is presented of operational failures of various types of components used in airborne military equipment; ways of reducing failures are briefly indicated.

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The Definition of Terms of Interest in the Study of Reliability—C. R. Knight, E. R. Jervis, and G. R. Herd. IRE TRANS., vol. RQC-5, pp. 34–56; April, 1955. Abstract, PROC. IRE, vol. 43, p. 896; July, 1955.)