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THE COVER—The emergence of important new devices that utilize the atomic properties of matter is exemplified by an ultrastable cesium atomic-beam frequency standard developed by the National Company. Heart of the equipment is the atomic-beam chamber (foreground) in which a stream of atoms moves upward through an rf field. When the field frequency is at exactly cesium resonance frequency, the energy level of the atoms is changed. The target chamber at the top detects the number of changed atoms and feeds back a frequency-correcting signal to the rf generator when it drifts off the cesium resonance frequency. The device has an accuracy of 1 part in 10^9 with a frequency stability of 5 parts in 10^{10}.

Photo—National Company, Malden, Mass.

John G. Brainerd

Director, 1956–1957

John G. Brainerd was born in Philadelphia, Pa., August 7, 1904. From the Moore School of Electrical Engineering, University of Pennsylvania, he received the bachelor's degree in 1925 and his doctorate in 1934.

From 1922 to 1925 he was a reporter on the *North American*, a Philadelphia newspaper which by a happy coincidence continued publication until one month before his college graduation. He then became associated with the Bell Telephone Company of Pennsylvania until his return to his alma mater, the Moore School of Electrical Engineering, as a staff member. For five years he was Chairman of the Division of Physical Sciences of the University's Graduate School, and today he is Director of the Moore School of Electrical Engineering.

During World War II, Dr. Brainerd supervised numerous research and development projects; the largest project was the development and construction of the ENIAC, the first electronic large-scale general-purpose digital computer.

Following his assumption of the chairmanship of the IRE Standards Committee in May, 1949, he devoted much effort to obtaining agreement to publish IRE Standards in the Proceedings. Late that year, publication was secured, and since then the widespread distribution of IRE Standards has enhanced the influence of the IRE. Previously, Dr. Brainerd had been chairman of the Circuits Committee, and the chief organizer and first chairman of the Professional Group on Circuit Theory. In 1954 he was chairman of the Philadelphia Section of the IRE, and chairman of the 1955 Eastern Joint Computer Conference.

Dr. Brainerd is co-author of two books in the electronics field. He edited two volumes of the *Annals of the American Academy of Political and Social Science*, and he holds membership in Tau Beta Pi, Sigma Xi, and Eta Kappa Nu. He is a fellow of the American Institute of Electrical Engineers and the American Association for the Advancement of Science, and the IRE representative on the council of the latter organization.

He became an Associate of the IRE in 1933, and changed his status to that of a Member in 1939. In 1943 he rose to the grade of Senior Member, and in 1951 he was elected a Fellow.

Poles and Zeros

Witness. Since over 50,000 members of the IRE have, in the process of achieving their current grades of membership, relied on as many as five other members as references, it is fair inference that the most widespread activity in the Institute is filling out reference forms. In so doing the referee must perform an act of fealty not alone to the candidate but also to the Institute. Faithfulness to the membership standards is clearly essential if our society is to perform properly one of its basic functions, recognition of professional advancement.

It may also be fairly inferred, since IRE members are human, that the temptation to be overly loyal to the friend or acquaintance who is seeking admission or advancement to a higher grade is strong. It's easy to praise, particularly when overstatement carries no direct penalty. But the damage incurred in shading recommendations in favor of the applicant is borne by every member who fairly earned his admission or transfer.

Readers who feel that this is idle preachment should look over the shoulders of the members of the Admissions Committee, those devoted servants who study over a thousand applications forms every month and who have learned to distinguish the careful, objective evaluation from the routine and the thoughtless. These men *know* that a better job should be done by the majority of the referees, and they know it can be done, because the minority does a good job. They also know how to protect the candidate against overzealous and picayune references and they refer all cases to the Membership Relations Coordinator, who reviews them and brings his recommendations to the Executive Committee for final action. All this effort cannot be effective if the information offered by the referees is incomplete or inaccurate.

The Admissions Committee therefore asks that referees take special pains to bear true and full witness. The answers to the questions on the form should be responsive. If we know the applicant and his record well, the job is simply one of weighing the record against the qualifications for the grade in question, which are printed on the form, and citing corroborative facts from personal knowledge which establish the standing of the candidate. If we need more complete information to fill out the picture Headquarters will gladly send a copy of the application data, and we should ask for it. If we don't know the candidate's background and can make no personal contribution, we should say so. This will not block the candidacy; it will merely require the applicant to name another member who knows him better. All such negotiations are handled with discretion and the referees' statements are, of course, completely confidential.

The reference form is a chore unless we do it well. A little extra effort can make it a satisfying contribution to IRE.

Second Round. The special issue on single sideband techniques (December, 1956) was not as singleminded as the casual reader might imagine, because hidden away among thirty-three papers on SSB was one on non-SSB. This was "Synchronous Communications" by J. P. Costas which took the position that all the other papers in the issue were barking up the wrong tree. This case of strange bedfellows was entirely advertent. Messrs. Kaar and Honey, who coordinated the issue, (see P and Z, December) were fully aware of the problem posed by Mr. Costas' contribution and decided, very wisely we think, to bind up the argument in one set of covers. This flexible point of view has since had its reward. The world of military, aeronautical and mobile communications has been embroiled, since the SSB issue appeared, in its most spirited technical discussion since the war. We take no sides but are happy to provide the forum, noting the fact that the participants can carry their positions around with them in one issue of the PROCEEDINGS, when it might have taken two.

Such jollity aside, we are happy to present in this issue (p. 534) no fewer than thirteen letters of discussion on the SSB special issue. The first of these is by Mr. Costas. In it he addresses himself specifically to the argument and he does with a style and invective that should dispel the notion that engineers cannot write forceful prose. No doubt there will be rebuttals and surrebuttals. We invite all readers to dig in. Mr. Costas is wearing the purple trunks.

Fellow. The award for Fellow Grade, nominations for which are due at Headquarters not later than the end of this month, is governed by the following new requirement, recently voted by the Board of Directors: "The grade of Fellow shall be conferred only upon a person who is either a Senior Member of the IRE or meets all the requirements for Senior Member, and has been a member in any grade for a period of five years preceding the year of nomination, except that these provisions in individual cases may be waived for cause by the Board of Directors, as in the case of members-at-large who are not members of any established IRE Section."

Bridge. Norris Tuttle, one of the most thoughtful observers of the scene of network technology, pointed out at NEREM last fall that there is a serious gap between the practitioners of network theory, who devise rigorous but extremely complex and tedious methods of network synthesis, and the practical circuit designers who are impatient with the specialized jargon of the theorists and go right on designing approximate networks based on rules a quarter century old. So it is a particular pleasure to present in this issue (p. 454) a paper by A. J. Grossman which goes far toward bridging this gap for a particular class of filter of practical utility.—D. G. Fink.

Scanning the Issue

A Tribute to Five Outstanding Men (p. 437)—Five of the annual IRE awards are named after men who have been outstanding figures of their time: Morris N. Liebmann, Browder J. Thompson, Harry Diamond, Vladimir K. Zworykin and W. R. G. Baker. The Editorial Department has asked five persons who were especially well acquainted with these men to prepare brief, personalized accounts of the remarkable careers which inspired these important awards.

Management of Large R & D Organizations (Hall, p. 451)—The size and complexity of some present-day engineering organizations give rise to some formidable organizational and management problems. A case in point is a 5000-man research and development operation, one of the country's largest, considered in this article. A timely and down-to-earth discussion is presented of the major problems confronting an organization of this size; namely, obtaining and then retaining top-flight technical talent, organizing them efficiently, and managing them effectively.

Synthesis of Tchebycheff Parameter Symmetrical Filters (Grossman, p. 454)—One of the major contributions to the field of network theory was Darlington's classic paper on the insertion loss method of designing four-terminal reactance networks, written eighteen years ago. As is true of much of the work in this highly mathematical field, the inherent complexities of the subject have made it difficult for the "man-in-the-street" designer to understand and apply Darlington's results to his every-day problems. This paper presents in tutorial fashion a detailed step-by-step explanation of one of the important contributions in Darlington's paper, backed up by a valuable set of design charts for determining the component values of a filter after its performance characteristics are once specified. The editors believe this to be the most complete and compact design reference on symmetrical equal-ripple Tchebycheff filters that has ever been published, and hope that it will do much to bridge the gap between theorist and practitioner.

A New Semiconductor Photocell Using Lateral Photoeffect (Wallmark, p. 474)—It has been discovered that when a semiconductor junction is exposed to nonuniform light a lateral photovoltage is produced parallel to the junction. This novel effect has been utilized by the author in a new photocell that can measure with extreme accuracy (less than 0.1 second of arc) the direction of a point source of light imaged on the cell by a lens. When the direction of the light coincides with the axis of the cell, no signal is produced. As the direction changes from one side of the axis to the other, the signal goes from one polarity, through zero, to the opposite polarity. An interesting feature of the device is that it can be made to electronically sweep an area instead of having to be mechanically aimed.

Design Considerations for Broad-Band Ferrite Coaxial Line Isolators (Duncan, et al., p. 483)—Until now, the use of ferrite isolators has been restricted to waveguides, since they alone can support modes of propagation which have the necessary characteristics to produce nonreciprocity in ferrites. In this paper, the authors make the important finding that by partially filling a coaxial line with a dielectric, the coaxial propagation mode will be so distorted that it, too, will have the required characteristics for use with ferrites. Thus, the use of nonreciprocal devices can now be extended to coaxial structures. It is interesting to note that a further extension to multiple wire transmission lines is reported in the following paper.

Analysis of Nonreciprocal Effects in an N-Wire Ferrite Loaded Transmission Line (Boyet and Seidel, p. 491)—Applications of microwave ferrite devices have heretofore been confined to waveguide systems. Other transmission systems would be of considerable interest, too, especially at frequencies below a few thousand megacycles where waveguides become unduly large. The preceding paper reported an important extension of ferrite isolators to coaxial systems, useful in the 2 to 4 kmc range. The present paper reports still another interesting extension of ferrites, this time at even lower frequencies—1 to 2 kmc and below. The structures proposed here consist of four-wire and eight-wire transmission lines, either embedded in or surrounding a ferrite rod. It is shown that these act, respectively, as a gyrator and a circulator, resulting in nonreciprocal devices that can be used at new low frequencies and are yet compact.

Backward-Wave Oscillator Experiments at 100 to 200 Kilomegacycles (Karp, p. 496)—A traveling-wave tube, operating as a backward-wave oscillator, has been built which will produce electronically tunable oscillations at wavelengths of 1.5 millimeters. This breakthrough into the region above 100 kilomegacycles represents an important advance in our techniques of generating extremely high frequencies, shedding new light on the personality of these tubes in this region of the spectrum.

Transistor Junction Temperature as a Function of Time (Mortenson, p. 504)—A novel and thorough study is presented of the variation of junction temperature with time for a given pulse excitation. Basic information is developed which will be of importance to circuit designers dealing with transistors that operate at low frequencies (below 2000 cycles), especially in switching and pulse circuits, where there is considerable temperature variation during the course of a single cycle.

Shutter Image Converter Tubes (Linden and Snell, p. 513)—Image converter tubes—tubes that pick up an image on a photoemissive cathode and project it onto a phosphor screen—have found important applications in converting infra-red images into visible ones (the snooperscope), in light amplifiers, and especially in ultraspeed photography. In the latter application, the image is rapidly pulsed on and off by pulsing the accelerating voltage of the tube. This paper presents an important simplification for both magnetically and electrostatically focussed tubes. A mesh is inserted between the cathode and the screen which acts both as a shutter and a focusing electrode. Thus the image can be pulsed on and off simply by pulsing the low mesh voltage (about 100 volts) rather than the 5 to 12 kv accelerating voltage as heretofore.

Minimizing Incidental Frequency Modulation in Amplitude-Modulated UHF Oscillators (Schaffner, p. 524)—Small variations in transit time within an AM oscillator tube give rise to unwanted variations in the frequency of the output, which at ultra-high frequencies can reach proportions of practical concern. This study shows that incidental FM can best be minimized by a proper selection of parameters of the feedback and cathode circuits and by using cathode or grid modulation instead of plate modulation. These conclusions will be of particular interest to designers of low power uhf transmitters in meeting FCC specifications on incidental FM in as economical a manner as possible.

Improved Keep-Alive Design for TR Tubes (Gould, p. 530)—Just one year ago the PROCEEDINGS published a paper which investigated why seemingly good tr tubes were failing to protect the crystals of radar receivers from the transmitted pulse. It was found that the keep-alive discharge, which readies the tr tube against the initial surge of the transmitted pulse, occasionally broke down from a glow to an arc and, as a result, momentarily quenched itself, exposing the receiver crystal to excessive damage. In this paper it is found that as severe a condition as glow to arc transition is not necessary for crystal damage, that all it takes is a wandering of the glow discharge along the wall of the anode. An improved keep-alive structure is developed which avoids this important source of trouble.

A Tribute

Last month at the Annual Banquet the IRE bestowed five important awards bearing the names of outstanding men of their times: Morris N. Liebmann, Browder J. Thompson, Harry Diamond, Vladimir K. Zworykin, and W. R. G. Baker. In paying tribute each year to these awards and to those who receive these high honors, we are apt to lose sight of the five men for whom the awards are named. Accordingly, brief sketches of their careers have been published on the following pages, in the order in which the awards were founded, in the belief that their lives will provide both inspiration and a better appreciation of the awards.

These personalized accounts represent considerably more than the usual biographies. They were prepared by persons uniquely qualified by long and close personal association to give us the true character of these men and the significance of their accomplishments. For these labors of love we are deeply indebted to:

Emil J. Simon, a Charter Member of the IRE whose personal donation to the IRE made possible the establishment of the Morris N. Liebmann Memorial Prize;

Edward L. Bowles, expert consultant to the Secretary of War during World War II, out of whose office Browder J. Thompson was working at the time he was killed while on a special mission;

Wilbur S. Hinman, Jr., Technical Director of the Diamond Ordnance Fuze Laboratories, 1956 recipient of the Harry Diamond Memorial Award, and a close associate of Harry Diamond during the latter's 20-year career at the National Bureau of Standards;

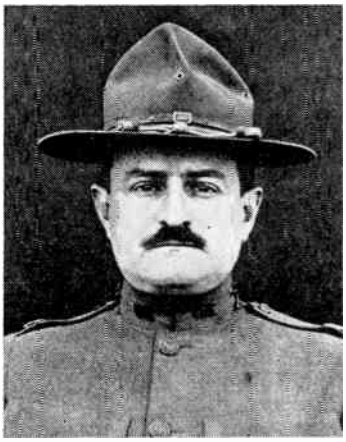
Irving Wolff, Vice-President, Research, of RCA Laboratories, where he has been a close associate of Vladimir K. Zworykin for a quarter of a century, and

Arthur V. Loughren, who as a radio receiver design engineer, and later as Vice-Chairman of the National Television System Committee and Director and Past President of the IRE has enjoyed an intimate professional association with W. R. G. Baker for more than 15 years.

—*The Editor*

Life of Colonel Morris N. Liebmann

EMIL J. SIMON, FELLOW, IRE



MORRIS N. LIEBMANN

"WITH firmness in the right as God gave him to see the right" Morris Nathaniel Liebmann in World War I fought and died on the battlefields of Flanders in defense of democracy.

Desirous of honoring his memory, a friend established the Morris N. Liebmann Memorial Prize in 1919.

Liebmann was an engineer and a soldier, and devoted to both professions. His engineering career was concerned chiefly with manufacturing. His aim was to make radio equipment more reliable. He joined the IRE soon after its formation and became its 68th member. Liebmann's first military training was in the Spanish-American War. A few years later he joined the New York National Guard.

A brief sketch of Colonel Liebmann's life may be of interest to the membership.

Morris Nathaniel Liebmann was born in New York, N. Y., on July 13, 1878. His father, Morris Liebmann, had emigrated to this country from Frankfort, Germany, in 1866 when he was 20 years of age and settled in New York. In 1875, he married Addie Henderson, a daughter of Nathan C. and Lydia Henderson who came from Ohio and Indiana, respectively. They had two sons, Morris and Walter.

When Morris was two years old, the family went west and settled in Deadwood, Dakota Territory. Gold had been discovered in the Black Hills and Deadwood became a booming mining town. Father Liebmann opened a general dry goods store in this frontier town and in time became a successful merchant. The children went to school in nearby Spearfish where their father maintained a branch store.

In June, 1895, Morris graduated from the Spearfish State Normal School and entered the University of Nebraska. He graduated with the degree of Bachelor of Science in electrical engineering in June, 1900.

The following year the family returned to New York and Morris Liebmann joined the firm of Foote Pierson & Company at 160 Duane Street, New York. This firm was well-known as a pioneer manufacturer of fire-alarm and telegraph equipment and possessed a reputation for producing superior products. Liebmann soon became Chief Engineer and later Vice-President.

The Wireless Specialty Apparatus Company founded by the late distinguished Greenleaf Whittier Pickard, former IRE president, became one of Foote Pierson's principal customers. Through this association Colonel Liebmann became interested in radio.

He contributed his talents toward perfecting the design and construction of Pickard apparatus. Notable examples were the "IP-76" receiver and the "Perikon" crystal detector. This apparatus became standard equipment in practically all radio stations, government and commercial, during the 1910's. "IP" stood for interference prevention; a rather optimistic designation in retrospect.

Foote Pierson also pioneered the manufacture of William Dubilier's earliest mica transmitting condensers. These soon replaced the leyden jar. The first one-half-kw and 1-kw quenched spark panel-type transmitting sets were developed and built at Foote Pierson. These were supplied in considerable quantities to the United States Navy before and during World War I.

When the Spanish-American War broke out, Liebmann was a student at Nebraska and joined a western volunteer regiment. After returning to New York he joined the 23rd New York National Guard, as a private. Rising rapidly in rank, he became a corporal and then a sergeant in 1904. In 1908 he was commissioned lieutenant and in 1913 captain of Company I.

He served throughout the Mexican Border campaign of 1916 and became regimental Adjutant. When the United States entered World War I, the 23rd Regiment became the 106th U. S. Infantry. Liebmann was promoted to lieutenant colonel. During the eight months of training at Camp Wadsworth in Spartanburg, N. C., Colonel Liebmann was transferred to second in command of the 105th U. S. Infantry. These two regiments, together with 107th, went overseas as part of the 27th Division in May, 1918. The Division moved to the battlefront in July and was brigaded with the British Army in Flanders occupying the "East Poperlinghe Line" opposite the Hindenberg Line at Mt. Kemmel, Belgium, a position that was subjected day and night to heavy artillery fire by the enemy.

The day that King George V of England was inspecting the 27th American Division, August 6, 1918, an enemy shell hit Colonel Liebmann and he was instantly killed. He was buried with full military honors in the Abeele Aerodrome Military Cemetery at Abeele, Belgium, on August 18, 1918.

In 1921, the Belgium Government posthumously bestowed on Colonel Liebmann the Croix de Guerre.

Liebmann's life no doubt was greatly influenced by his boyhood years in the west. Here life had been really rugged and he was brought up among pioneers of whom many were exiles from civilization. His outdoor life built up his bodily frame and he possessed both stature and physical strength, which served him well as a soldier. His love of the strenuous life and his admiration for Theodore Roosevelt undoubtedly influenced his career. Many qualities in Liebmann's character, especially his rugged Americanism, seem to parallel those of the President.

Firm, staunch, and resolute, he possessed qualities that endeared him to his friends and that commanded the respect and admiration of his comrades in arms. Liebmann was a quiet, tolerant man with infinite patience. A born leader of men, his word and commands were always respected.

With courage and fortitude he responded to the call of duty and finally gave his life for his country.

Are you dead? No comrade, No!
The dead lie only with the foe.
You sleep, 'tis true, but yet you live;
You gave your life, yet did not give
Your deeds to be forgotten thus
When bone and sinew turn to dust.
In Flanders Fields.

Bibliography and Sources

[1[*Polk's Minnesota State Gazeteer and Business Directory*. Including Dakota Territories. 1880–1881, p. 830, 1884–1885, p. 891, and 1888–1889, p. 1464.
[2] Bennett, Estelline. *Old Deadwood Days*. 1928. Photo of Main Street of Deadwood in 1888, p. 212.
[3] *Peterson's Illustrated Atlas of South Dakota*, 1904, pp. 199–200.
[4] 1905 Census, County Court House, New York, N. Y. Liebmann family address: 409 Amsterdam Ave., New York, N. Y.
[5] 1915 Census, County Court House, New York, N. Y. Liebmann family address: 185 West 135th St., New York, N. Y.
[6] *New York Times*, November 2, 1911, obituary of Addie Liebmann, mother of Colonel Liebmann.
[7] Simon, E. J., and Israel, L. L. "The Operating Characteristics of a 3-phase 500-Cycle Quenched Spark Transmitter," Proceedings of the IRE, Vol. 2 (September, 1914), pp. 217–236. See Figs. 13–15, p. 232.
[8] *New York Times*, August 15, 1918, "Lieutenant Colonel Liebmann Killed in Action."
[9] *New York Herald Tribune*, August 15, 1918, obituary and photograph of Colonel Liebmann.
[10] "A Short History and Illustrated Roster of the 105th Infantry," Edward Stern and Co., Philadelphia, Pa.; 1918, containing photograph of Colonel Liebmann and burial in Belgium.
[11] Obituary of Colonel Liebmann, Proceedings of the IRE, Vol. 6 (October, 1918), p. 235.
[12] *New York Times*, June 6, 1921, "Belgium Decorates Heroes of the 27th."
[13] O'Ryan, John F. *The Story of the 27th Division*. 1922, pp. 219–221.
[14] Swetland, M. J., and Swetland, L. *These Men*. 1940, pp. 138–139.
[15] Military records, Adjutant General's Office, State of New York, Albany, N. Y.
[16] Bureau of Vital Statistics, Health Department, New York, N. Y.
[17] Last Will and Testament of Morris Liebmann, filed Hall of Records, New York, N. Y. Liber 1092; page 229.
[18] *New York Times*, September 8, 1928, obituary of Morris Liebmann, father of Colonel Liebmann.
[19] Records of Pine Lawn Cemetery, Pine Lawn, N. Y., burial ground of Colonel Liebmann's parents.
[20] Letter from the Registrar, University of Nebraska, dated December 17, 1956.

Recipients of the Morris N. Liebmann Memorial Prize

Awarded to a member of the IRE for a recent important contribution to the radio art.

L. F. Fuller	1919	Edmond Bruce	1932	P. C. Goldmark	1945
R. A. Weagant	1920	Heinrich Barkhausen	1933	Albert Rose	1946
R. A. Heising	1921	V. K. Zworykin	1934	J. R. Pierce	1947
C. S. Franklin	1922	F. B. Llewellyn	1935	S. W. Seeley	1948
H. H. Beverage	1923	B. J. Thompson	1936	C. E. Shannon	1949
J. R. Carson	1924	W. H. Doherty	1937	O. H. Schade	1950
Frank Conrad	1925	G. C. Southworth	1938	R. B. Dome	1951
Ralph Bown	1926	H. T. Friis	1939	William Shockley	1952
A. H. Taylor	1927	H. A. Wheeler	1940	John A. Pierce	1953
W. G. Cady	1928	P. T. Farnsworth	1941	Robert R. Warnecke	1954
E. V. Appleton	1929	S. A. Schelkunoff	1942	Arthur V. Loughren	1955
A. W. Hull	1930	W. L. Barrow	1943	Kenneth Bullington	1956
Stuart Ballantine	1931	W. H. Hansen	1944	O. G. Villard, Jr.	1957

Browder Julian Thompson

EDWARD L. BOWLES, FELLOW, IRE

BROWDER J. THOMPSON

THE name of Browder J. Thompson, recipient of the Morris N. Liebmann Memorial Prize in 1938 and Fellow of the Institute of Radio Engineers, is known to many through his numerous articles in the professional literature of electronics, particularly in the field of vacuum tubes. His contributions to the art, both numerous and significant, began a remarkably short time after his graduation from the University of Washington in 1926 and continued throughout his brief but conspicuous career which ended tragically in his death in the summer of 1944 while serving his country. The particular mission which cost him his life was one of his own determination in the interest of applying electronic techniques to the interdiction of enemy transport at night.

In the span of not quite twenty years, B. J., as he became known to his friends, had lived a full and productive life. He enjoyed the satisfaction of truly professional accomplishment wherein he perceived and solved many vexing problems—problems which had escaped the intellect of others. He had known the satisfaction of applying his particular perceptive skills, his unusual capacity for ingenious simplification of problems, and his inventive resourcefulness. Aside from these accomplishments, I suspect, and I speak as a close friend, this man valued above all the friendships he had attracted in the course of his life.

One of Thompson's first contributions was to the realization of a novel space charge grid type of tube with unusually high insulation to act as an electrometer device. This new tube was to fill a great need in research for means of measuring minute currents. The article by Thompson describing one development of this tube attributed to it a capability of measuring 10^{-17} ampere. The lucid, direct analysis of the problems to be solved in achieving such a tube was to become symbolic of this man's ability to get at the meat of a problem and to express the critical issues with discerning incisiveness. His next bold step was to develop what became the "acorn" tube of the trade. Here was an outstanding contribution to ultra-high-frequency techniques where it had been generally conceded there was little hope for the extension of the capability of triodes through innovation. As Dr. Engstrom put it recently, "The production of this tube was an important factor in breaking the log jam which theretofore had held up uhf development."[1]

Here was a demonstration of what could be accomplished in what was accepted as a worn out field by intelligent, untrammeled analysis and an awareness of limiting parameters including transit time.

There was other research and publication on screen grid tubes, power pentodes, transit time and noise—contributions which in many respects shaped the art. This constructive curiosity, this feel of the perfectionist and record of achievement coupled with his gifted personality blessed with deep human understanding and ethical standards led inevitably to multiple responsibilities of greater moment. Browder's leadership was inspirational. Many were the products of his intellect in the abstract and in substance. Always he had the fine quality of giving others the feeling that they themselves had done the job.

[1] Engstrom, Transcript of Testimony, Television Inquiry, Senate Interstate and Foreign Commerce Committee, p. 1429; March 15, 1956.

Thompson had been recruited by the General Electric to work on test. He was soon heading a small group in the Vacuum-Tube Engineering Department at Schenectady. From 1931–1940 he was associated with the Radiotron Division of the RCA Manufacturing Company in charge of the Research Section, Research and Engineering Department. Here he was engaged in ultra-high-frequency tube development including television tubes in their many aspects. In 1940 he became associate director of the department, then Research Director of the new RCA Laboratories at Princeton, N. J.

Browder J. Thompson was a genteel soul of noble character and broad understanding. Though exacting in his requirements of others, as well as himself, he was tolerant and sympathetic. His every action was characterized by an enviable urbanity. Those who were privileged to know him were ever impressed by his quiet, reasoned approach to a problem. His precision of expression was impressive, yet it was simplicity itself, without pretense or embellishment or affectation. His dress, as if reflecting his inner mind, was likewise precise, correct and without affectation. The dignified manner when necessary, along with the well-measured Homberg, always appeared in character and in keeping with the man. There was ready always a contagious smile prompted often by a subtle insight into human nature and an indomitable sense of humor.

Browder Julian Thompson was born in Roanoke, La., on August 14, 1904. He went to private school in Lake Charles, La., where he lived with his father, B.J. Thompson Senior, his mother, Julia Thompson, and an older sister, Marguerite, until he was nine. The family migrated to Minot, N. Dak., where he was graduated from high school at the age of fifteen. In this period his mother operated a small business to provide for the family and plan for the children's future. Finances were helped in this period by B. J.'s part-time job in a haberdashery establishment. It was a period of hard work for all, brightened by the influence of a mother who had the stimulating quality of facing adversity and making life, family life, interesting and deeply rewarding. In this period young Browder was given responsibility for his older sister—a task he evidently took seriously, if I may judge from his sister's interpolations in response to my request for information on the family's history. There seems to have been an early inception of the sense of responsibility which was to be a great strength in Thompson's life. At every step in this struggle, it was a model struggle of a healthy family with a desire to live.

There came the time for University traning. Thompson had dreamed of M.I.T. but this was out of the question because of the expense. The solution was to pull up stakes in Minot and go to Seattle, where the two children could attend the University of Washington. In the initial move, funds took these three determined minds only as far as Spokane. Here they settled down, the mother with her shop, the children working in the fields. Seattle was ultimately reached. Working his way through school, including time in Alaska on railroads and other constructive interludes, Browder Thompson was graduated in 1926.

In checking dates to explain hiatuses, I found that Browder apparently was sensitive about his comparative youth in the G. E. environment. He may have felt that being thought of as older than his looks implied would help him professionally. On the occasion of an office party on his twenty-sixth birthday, his sister Marguerite was sought out by the group to give them the number of candles the cake should bear. She left them to guess. The cake appeared at the appropriate moment as a 30 candle-power job. He did not destroy the illusion, nor did his sister betray him.

In the course of the last war, I was drafted by Secretary Stimson to help him in the application of radar and other electronic techniques to military operations. My responsibilities went beyond my ability to carry them. I therefore sought assistance on selected problems.

There were urgent problems needing expert attention having to do with magnetron development and procurement and the need for understanding of coordination of tubes and radar equipment programming. Another critical problem was that of proximity fuses. Someone at Secretary level was needed to make a thorough examination in these two areas. I could think of no person better qualified for this task than B. J. Thompson. Consultation with Dr. Engstrom and Thompson himself resulted in the appointment of this man as Expert Consultant to the Secretary of War. His letter of assignment, bearing my signature, is dated September 19, 1943. The memorandum which Thompson prepared on Magnetron Requirements and Facilities, March 9, 1944, proved a masterpiece of comprehension and skillful handling of an extraordinarily involved and sensitive interservice-industry problem.

The effectiveness of Thompson's performance led me to give him also the assignment of auditing the guided missile activities, some of which showed promise of application if adequately supported. Soon Thompson had got to the bottom of this confused problem and was ready with constructive recommendations.

By June of 1944 an important operational problem had arisen. Military progress in Italy by the Allies had resulted in the interdiction of rail supplies to the enemy front. Therefore all substantial movement was by transportation on the highways by night. Critical to our imminent success was the destruction of this line of supply. Questions were raised as to how to do the job by aircraft applying radar or other technological methods. Following a conference with the Secretary and General Marshall, I conferred with Thompson on the subject. He expressed not simply a desire to work on this problem but a vital interest in it. He asked to be able to go over to Italy to make his own examination of the problem. I made arrangements with the Theater Commander and the Tactical Air Commander so that he would have a free hand in studying the problem.

On his arrival in England, I was dispatched the following message in hurried pencil on a page torn from his notebook:

Monday

CLB, —

After a slow start (plane + train got me to London at 9 PM Saturday), it seems I'll get to work tomorrow. These little affairs they're sending over seem to have most everyone's attention at the moment. I'm sure things will be all right as soon as I can get well started.

Please ask one of the girls to write my sister ~~to~~ that the flying bombs around ~~nothing is~~ worrying me or keeping me from sleeping. I don't know what impression the newspapers give but I just find things interesting. The German reports are greatly exaggerated. I don't feel there is any danger

BJT

At our last conference before he left Washington, Thompson put into my hands an extract from a communication he had just given his mother:

"If anything happens to me on this job, you can have the satisfaction of knowing that it was in the active service of our country and that I would never hesitate at anything I should do because of risks. You taught me that. I'd rather die this way than most others."

In the late afternoon of July 4, 1944, Thompson presented his credentials to the Commanding Officer of the 57 Bombardment Group operating out of the Grosseto Avea, Italy. He expressed specific interest in technical aids, including radar, for the location of tactical targets during night flying and conditions of poor visibility. I shall quote from a report,

"Dr. Thompson indicated an ardent desire to see for himself the problems confronting tactical units in locating targets at night. Dr. Thompson indicated a strong desire to participate personally in an operational flight over enemy territory. It was suggested to Dr. Thompson that he participate in a local training flight scheduled for the evening, but Dr. Thompson stated such a flight would be of no value to him. He said in substance, 'I should like to go over the lines to observe enemy traffic on the roads. It would be of no value to me to see our own traffic with lights on, as I can see that in the States. I want to see enemy traffic operating under blackout conditions.' He stated only by a personal flight could he get the true picture of visibility conditions over target areas under changing conditions of clouds and moonlight and that such a flight would greatly aid him in successfully carrying out the mission which had been given him by the War Department."

Thompson was briefed on the combat operations for the intruder missions of that evening. He chose the flight which was over the area which involved attacks only on road movements. He took off around 10 P.M. The weather over the target area was not hazardous, there was a full moon and scattered clouds. Only light and inaccurate flak had been encountered in this area in the past. There had been no trouble with enemy aircraft. Five other similar aircraft were assigned missions in the same or adjoining areas. Two were in the same area as the Thompson plane. All but the Thompson plane returned. The returning craft reported no enemy aircraft, very slight and ineffective flak. No fires were noted that could be that of a burning aircraft.

Later information indicated that the missing plane went down at Pontedera, near the airstrip—four or five miles from Florence. Browder Thompson's body is interred in the U. S. Military Cemetary at VADA.

Of this man, Secretary Stimson observed:

"It is a great tribute to him that, under no other compulsion than his own great desire to do all in his power to aid in the solution of a most urgent problem facing our own Army and those of our allies, and thereby to assist in bringing the war to an early conclusion, he deemed it necessary to do as he did."

RECIPIENTS OF THE BROWDER J. THOMPSON MEMORIAL PRIZE

Awarded to an author under 30 years of age at date of submission of manuscript, for a paper recently published by the IRE which constitutes the best combination of technical contribution and presentation of the subject.

G. M. Lee	1946	J. F. Hull	1950	Richard C. Booton, Jr.	1953
C. L. Dolph	1947	A. W. Randalls	1950	R. L. Petritz	1954
W. H. Huggins	1948	A. B. Macnee	1951	Blanchard D. Smith, Jr.	1955
R. V. Pound	1949	H. W. Welch, Jr.	1952	Jack E. Bridges	1956
		D. A. Buck	1957		

Portrait of Harry Diamond

WILBUR S. HINMAN, JR., SENIOR MEMBER, IRE

HARRY DIAMOND

HARRY DIAMOND was born in Russia on February 12, 1900, and emigrated to the United States as a child. He was graduated from the Massachusetts Institute of Technology in 1922 and received the Master's degree in electrical engineering from Lehigh University in 1925. From 1927 until his death in 1948, his career was a long succession of major technical achievements in radio and electronics. He was active in IRE affairs throughout his life. The Harry Diamond Memorial Award was created by his many professional friends and colleagues because they believed Mr. Diamond typified the ideal public servant in the technical activities of the Government.

In retrospect, one is inclined to dwell on Diamond's successes, his brilliance, his resourcefulness, and his resolute qualities, and to ignore the more human qualities which made the man. These other qualities are the ones which made him such a genial companion and which made for the strong loyalties of his associates and subordinates. These are all part of the portrait, but the technical record comes first.

Dellinger, Pratt, Lowell, and Dunmore had already devised the first radio range beacon system for the nation's airways when Harry Diamond joined the staff of the National Bureau of Standards at the little airport at College Park, Md. After one year, when Pratt turned to other work, Diamond was made chief of the activity, and new developments continued to evolve at a rapid rate until 1933, when the work was terminated by transfer to the Aeronautics Branch of the Department of Commerce.

This work resulted in the establishment of the whole system of radio range beacons and locations which mark the airways. In this period, the shielded ignition system for aircraft and the stub antennas were developed. The work culminated in the development of the first complete instrument system for landing airplanes "blind," as in heavy fog. Harry Diamond acted as radioman on the first completely blind cross-country flight. The instrument landing system used today is the grandchild of the landing beam system originated by Diamond's staff in the early thirties.

The next major achievement of Diamond and his staff was the radiosonde. These small balloon-carried automatic weather instruments make several ascents daily, not only from stations all over the United States, but at all points in the world. About two million radiosondes have been launched to date, and the original system is still the primary means for measuring the conditions of the atmosphere which make the weather.

The radiosonde led to the development of one of the first permanent-site automatic weather stations. It was designed to report weather data from remote locations for long periods without servicing.

About a year before Pearl Harbor, the United States became interested in the development of the proximity fuze. It was calculated (and borne out in combat) that a fuze which would explode a projectile near a plane, or at the best height above a target on the surface, would increase lethality by a factor of five or ten.

Initial work was at the Carnegie Institution's Department of Terrestrial Magnetism. Two or three months later, the National Bureau of Standards was brought into the program, and Harry Diamond was given responsibility for this phase of the Bureau's work. Within about four months of the start of the program, Diamond's group established feasibility of the radio proximity fuze through conclusive tests in bombs dropped at the Naval Proving Ground at Dahlgren, Va. Throughout World War II, this group acted as the central laboratory of Division 4 of the National Defense Research Committee, and Diamond was the central figure of the group. Much of the basic proximity fuze technology was developed under his direction.

By the end of the war, he had built up a strong technical staff, which continued in the field under sponsorship of the Army Ordnance Corps until 1953. In that year, five years after his death, the organization was transferred to the Ordnance Corps and named the Diamond Ordnance Fuze Laboratories.

With the close of World War II, Diamond became interested in still more advanced electronic systems and in new electronic component developments. Under his guidance, a strong staff on industrial electronics was built up at the National Bureau of Standards. Printed circuits were given their first real start in this program, and considerable progress was made in automatic assembly processes, high-polymer potting compounds, and electrical transducers and controls. The Bureau's initial work on digital computers was done under his direction, and SEAC, one of the nation's better-known digital computers, evolved from this program.

Diamond maintained his interest in air navigation. One of his last acts was the preparation of the section on "Radio Aids to Aviation" in Keith Henney's "Radio Engineering Handbook."

As to the nature of the man, perhaps the most outstanding trait was confidence—confidence in himself and in the ability of his staff. This gave him the courage to undertake new and challenging problems, even in fields with which he was not familiar. A minor but illustrative example is an elementary text in meteorology, a field in which Diamond had little knowledge or experience. Since a text for meteorology was needed, Diamond studied standard texts and produced an excellent text for an elementary course.

Some of Diamond's technical gambles would have completely dismayed a more cautious man. It must have been his keen technical perception, backed by good technical training, and a willingness to work without regard to the clock that made so many of his ventures "pay off." Diamond even contracted to describe a new radiosonde system at the Annual Convention of the American Meteorological Society before he had a system to describe. The system was completed on Christmas Eve, only three days before the Convention.

In appearance, Diamond was well set up—about five feet ten inches tall, one hundred eighty pounds. He had a shock of jet black hair, and regular features. He was vitally alive and seldom at rest. His later years were characterized by careful attention to the development and the welfare of his staff. Diamond was modest but never retiring, and he was careful to credit and promote the individual accomplishments of others. His competence and general character induced very strong staff loyalties.

His associates who knew him best realized too late that his personal drive and devotion to his job caused him to work beyond his physical limit. This was probably a major cause of his untimely death in 1948.

Recipients of the Harry Diamond Memorial Award

Awarded to a person in government service for outstanding contributions in the field of radio or electronics as evidenced by publication in professional journals.

A. W. Haeff............ 1950
M. J. E. Golay......... 1951
Newbern Smith........ 1952
Robert M. Page........ 1953
Harold A. Zahl......... 1954
Bernard Salzberg....... 1955
Wilbur S. Hinman...... 1956
Georg Goubau.......... 1957

Vladimir K. Zworykin

IRVING WOLFF, FELLOW, IRE

VLADIMIR K. ZWORYKIN

FROM the time when the Pilgrims landed on this continent every disturbance in Europe which has threatened the freedom of the individual has made its contribution to our life in the form of those talented men and women who came here to start a new life. The Communist revolution of 1917 is particularly noteworthy for its gifts of outstanding scientists and engineers. Among them are such notables as Sikorsky and Von Karman in aeronautics, Timoshenko in mechanical engineering, Tykociner in electrical engineering, and V. K. Zworykin in electronics. In the impact on our daily life, electronic television, which Zworykin has done so much to promote, must certainly take high rank.

I first became well acquainted with him when he was forty years old; he is now over sixty-five. At the end of this period he has the same qualities which contributed to his success as a younger man. He has the same perseverance and crusading spirit in overcoming obstacles, whether man-made or technological, the same uninhibited originality, the same grandiose planning of technological revolutions, and the same enthusiasm.

Although V. K. Zworykin has made many contributions to electronic technology, he is best known for his inventions in electronic television. The extent to which this work has been a part of his life will be made more apparent in the following brief biography.

He was born in 1889, in Mourom, Russia. His father operated a fleet of boats on the Oka River. By 1910 he was studying engineering at the Institute of Technology in St. Petersburg. His courses included some laboratory work in physics. Here he quickly completed all the experiments assigned to him and asked his professor what to do next. The same thing happened when a series of additional experiments was assigned. The laboratory soon ran out of student problems and his professor, Boris Rosing, asked him whether he would like to help in his own laboratory. To the young student this was a tremendous thrill and opportunity and he accepted the offer eagerly. It may surprise many that Rosing was at that early date already trying to develop a method of electronic television using the Braun cathode-ray tube. The young student was not able to pursue his television research very actively because of academic duties, but the contact with this advanced development had a permanent influence on his later life.

Shortly after graduation, followed by a period of advanced study in physics under Langevin in Paris, World War I came along and he became an inspector of communications equipment for the Czar's government. I have been told that some of Dr. Zworykin's time during this inspection period was devoted to trying to induce those whose activities he was inspecting to work on electronic television. After the war ended he returned to his research activity. However, as an officer in the Czar's Army he was forced to leave suddenly when the Communists took over. After adventures which took him twice around the world he came to this country in 1919.

His first technical association in the United States was in the Research Laboratory of the Westinghouse Electric and Manufacturing Company in Pittsburgh, Pa. Here, in 1923, he applied for a basic patent on an all-electronic television system, which he demonstrated experimentally to his superiors. Following this demonstration he was advised to do "something more useful" and directed his attention to research on photocells, facsimile, and sound movies. However, before long he persuaded his supervisor to let him continue his work on electronic television, and in November, 1929, he presented the first results of his experiments at a meeting of the IRE. Thus, while one economic boom was collapsing, the technical foundations for a major contribution to the next one were being laid; significantly, some twenty years elapsed between the time of the initiation of electronic television in the United States and the time it became successful commercially.

In 1930, Zworykin was transferred to the Radio Corporation of America. The television equipment designed in the 1920's had a cathode-ray tube for showing the picture, but a mechanical pickup system. Zworykin did not feel content until he had developed a practical completely electronic system, elaborating the ideas which he had demonstrated in 1923. In the early 1930's the successful development of the iconoscope completed the fundamental elements for the electronic television system which we have today. Many improvements in circuitry, in pickup tubes, in resolution, and in viewing tubes have been made since that day and Dr. Zworykin has shared in many of these. With the development in the early 1930's, however, it became clear that electronic television was practical.

Dr. Zworykin has told me many times that his goal when he started in television was not television for entertainment purposes, and, although entertainment happens to be the television which we know best today, he has always insisted that the real goal of television is the extension of sight. Following up on this idea, he proposed in the early 1930's, while the guided missile was still in the realm of popular science fiction, a missile which would through the television camera in its nose be able to transfer the vision of the target to a remote controller and thus enable him to guide its impact through radio control signals. This was truly an extension of man's sight. This project to me is typical of the kind of thinking and imagination that I so admire in Dr. Zworykin. Perhaps the idea may seem commonplace today, but in 1934 it was radical. True, the concept was well ahead of the ability to accomplish the details at the time, but the thinking assessed accurately our fundamental ability to accomplish the objective, the necessity for doing it, and the trend of the art. At that time we did not have controllable missiles, we did not know how to navigate them, we did not have television of adequate sensitivity, nor television which was nearly small enough, and the concept had no place in the thinking of our military planners. But the idea was sound, as subsequent events have proven, and the ability to overcome details was just a question of time.

It is characteristic of the man that once something has been developed to a practical state he essentially loses interest in it; this has not changed with age. Entertainment television, both black and white and color, and guided missiles using television cameras in the nose are ideas of the past. He has now shifted his interest to newer fields, many of which may seem just as visionary as electronic television did in 1910 and television-guided missiles in the early 1930's.

In the television area, Dr. Zworykin will not feel that his job is done until we have television cameras so cheap that they can be used in every home, farm, and industrial organization to extend sight so effectively that by pressing an appropriate button one will be able to see any place he desires.

Dr. Zworykin through his own efforts and through the efforts of associates has had a part in the development of many important electronic devices such as the electron microscope, the electron multiplier, the electronic image tube, the Vidicon, and some of the fundamental components used in present-day computers. Although these developments have had an important impact throughout the world of electronics and in other areas of science, I think of him mostly in terms of some of the ideas which have not yet been accomplished.

Shortly after the war at a time when some destructive hurricanes struck the East Coast, Dr. Zworykin conceived the idea that it should be possible to divert hurricanes with possibly a small expenditure of power at the time when the hurricane was in its formation stage. He discussed this problem with Dr. John Von Neumann, of the Institute for Advanced Study at Princeton. Out of such discussions grew the concept of the electronic computer for weather prediction, since it soon became apparent that the mathematics were too involved to handle in any other fashion. The prediction of weather by electronic means is now a problem on the way to solution. The second part of the project, learning how to destroy or divert the hurricane, is still to be accomplished, but the foundation of knowledge for doing this is now being built up.

Automatic control of automobiles on the highways is another typical concept. With the advent of the superhighways and turnpikes where crossings are nonexisting and where driver drowsiness is one of the foremost causes of accidents, he believes that we have come to the point where automatic control of automobiles on trunk highways is not only feasible but is a necessity. With his associates he has demonstrated possible methods for obtaining this control.

One can think of many reasons why hurricanes cannot be diverted from their paths and automobile traffic cannot be controlled automatically but in 1910 there were just as many reasons why electronic television was impossible and in 1934 there were just as many reasons why the television missile could not be accomplished.

Dr. Zworykin's major interest at present is in medical electronics. In the direction of laboratory work he is active both at RCA and the Rockefeller Institute. In addition he is devoting a major effort to the stimulation of greater interest and cooperation between electronic engineers and the medical profession. He realized that the IRE Professional Group on Medical Electronics could well serve as the focal point for this joint effort, but he found that IRE rules for Professional Group membership were interfering with his program. His impatience with these traditional membership provisions, his refusal to accept them as inevitable, and his salesmanship in having them modified is typical of his life.

What I have written up to now concerns to a great extent Dr. Zworykin as a scientist, inventor, and promoter, but a story like this would not be complete without saying something about him as a man.

Along with all the rest of us, Dr. Zworykin has his faults. These faults are characterized, however, by a certain spontaneity and transparency and over a period of years I have come to like him as much for his faults as for his virtues.

His youth was spent in an environment of the well-to-do class in Czarist Russia. In education and in social graces this group had no superior, and in his personal life he continues to typify to me many of the characteristics of that environment. Although Dr. Zworykin is very sociable, he is not one who is happy or comfortable in large groups of people. He is most content with his family or when he is with small groups of friends and would rather entertain at home than be entertained. His home at Taunton Lakes has served as focal point for many stimulating week-end discussions with friends and younger men on his staff at intervals between swimming or other athletic activity.

Travel is one of his major recreations and there are few places on the earth's surface where he has not been. The stories he brings back of his travels and other occurrences in an eventful life have a distinctive flavor and sense of humor which would be worthy of the stage.

Dr. Zworykin has been well recognized throughout the world for his accomplishments. Among the medals he has been awarded are the Medal of Honor and the Morris N. Liebmann Prize of the Institute, and the Chevalier Cross of the French Legion of Honor. He is a Member of the National Academy of Science. In setting up the provisions for the Award which bears his name, he has tried particularly to have it serve as a stimulus to the younger engineers now active in the field of electronic television on which he embarked almost a half-century ago.

Recipients of the Vladimir K. Zworykin Television Prize

Awarded to a member of the IRE for important technical contributions to electronic television.

B. D. Loughlin	1952	Alda V. Bedford	1954	Frank J. Bingley	1956
Frank Gray	1953	Harold B. Law	1955	Donald Richman	1957

W. R. G. Baker—An Appreciation

ARTHUR V. LOUGHREN, FELLOW, IRE

W. R. G. BAKER

WALTER BAKER was born in Lockport, N. Y., on November 30, 1892. He attended both grade and high school in Lockport and moved to Schenectady in 1907, where he obtained a job in the local telephone office, receiving and recording trouble reports from customers.

Although he was not what would be regarded as an ardent scholar in his youth, he preferred reading to many other activities. Even today he frequently carries in his briefcase a book on one subject or another. When Baker was unable to find out from his fellow workers at the telephone office precisely how they located the faults in the line, he found a textbook which explained much of the mysteries of the telephone business. This study eventually led to the job of Assistant Wire Chief to a man named C. A. Hoxie, who was later to become prominent for his work in talking movies in the General Electric Company.

Walter saw the need of further education and in 1912 entered Union College, working part time at the local telephone company. Upon graduation in 1916, he was offered the opportunity of going to Albany as Assistant Division Engineer for the telephone company, but chose rather to go to New York, entering the employ of the Western Electric Company. In June, 1917 he joined the General Engineering Laboratory of the General Electric Company, working with Hoxie on a high-speed photoelectric recorder for wireless telegraphic code signals. While engaged in this work, he chanced to observe Chester Rice working with what was then called a pliotron by the General Electric Company. This was a three-element, high-vacuum tube developed by Dr. Langmuir. Rice's explanations, augmented by discussion with Langmuir and Eli Kinney, another telephone man who was deeply interested in radio and was experimenting with the De Forest Audion in his home, stimulated Baker's interest. As a result, Dr. L. T. Robinson, then head of the General Engineering Laboratory, gave Baker a new assignment in 1918 which soon involved the development of radio telephone transmitters for potential sale to the government.

This activity led to a proposal to develop a company-owned station. Although Baker was interested in the broadcasting aspects because of the activities of the then Station KDKA, the Company management were not convinced that this was a profitable venture. Thus, the first Station WGY was really planned, at least on the official approvals, as a voice communication channel to handle Company work between its other plants.

Following the war, Baker transferred to the newly-established Radio Department. Here, during the early 1920's, a major impact was felt from the change of emphasis in radio communication away from operation by code to operation by voice—away from point-to-point operation into broadcast operation. Voice transmitters of relatively high power were needed; radio receivers capable of being operated by unskilled listeners were needed. With this rapid growth of needs Baker became first Designing Engineer in charge of transmitters in 1920, and, in 1926, he received the responsibility for the design of all General Electric's radio apparatus. He was made Manager of the Radio Department in 1928. Prior to this time, Baker was instrumental in establishing the South Schenectady transmitter development site where most of the short wave work of the General Electric Company was done. He has also been instrumental in establishing Company stations KOA at Denver and KGO in Oakland.

While the problems of broadcast transmitter design of the mid-twenties seem pretty simple writing now in 1957, the man who was first willing to see a broadcast transmitter designed which required the simultaneous operation of some 20 water-cooled tubes was a man who brought a good deal of courage to his daily task.

At the end of World War I, a complex patent situation had led the Navy Department to sponsor the organization of a U. S. corporation to acquire all U. S. rights to patents needed for effective radio communication. Radio Corporation of America thus came into existence with relations by way of stock ownership, rights under patents, etc., with a group of companies which ultimately came to include the General Electric Company, Westinghouse Electric Company, American Telephone and Telegraph Company, and Wireless Specialty Apparatus Company. The agreements which formed the basis for RCA's initial operation had not been worked out with any expectation that the public service of radio broadcasting would become an important matter. As a consequence, the radio apparatus which was initially sold to the public by the companies which had relations with RCA was all sold through RCA as the sales agent, with the manufacturing exclusively by others than RCA. By the late 1920's, this arrangement had come to exhibit its cumbersome nature and RCA took over the responsibility for the manufacture of radio consumer goods from the companies who had been its suppliers up to that time. Dr. Baker was asked to become Vice-President in Charge of Engineering for the new manufacturing subsidiary of RCA, which was called the RCA-Victor Corporation; accordingly, he moved to Camden late in 1929 to take over this new responsibility. As the Camden operation expanded, he became Vice-President in Charge of Engineering and Manufacturing.

The transfer of engineering and manufacturing activities from Westinghouse Electric Company and General Electric Company to RCA had been accomplished by an undertaking made in good faith and with the belief that it was in full conformance with the country's laws, that the two larger companies would refrain from competing in this rather limited field with the subsidiary which was so largely owned by them. However, in due course, the courts held that this undertaking was in fact not in accordance with the laws of the country and Westinghouse and General Electric renounced such portion of the agreements with RCA as the courts construed as being in contravention of the statutes. One consequence of this litigation was that the two major electrical manufacturing companies were required to divest themselves of their substantial ownership in RCA and, as a corollary, were freed to resume radio apparatus manufacturing and sale on their own behalf.

Dr. Baker rejoined G.E. in 1935 to supervise the reestablishing of his original company's place in this field. He was made Managing Engineer of the Radio Receiver Section in 1936 and, in 1939, was made manager of the company's radio and television department. During this time, the field which had earlier been called "radio and television" was showing more and more signs that this was too narrow a terminology. The broad term was commencing to be recognized as "electronics." In 1941, Dr. Baker was elected a Vice-President of the General Electric Company with responsibility for a department which, in later years, became known as the Electronics Division. His record, and that of his associates in that department of his company, is written in many places in the history of the U. S. effort in World War II.

A man's career may consist of many threads interwoven in a complex pattern. The thread which finds Dr. Baker with the IRE starts with his becoming an Associate in 1919 and continues with the publication of his first technical papers in the August and December issues of our PROCEEDINGS in the year 1923. This thread now goes to his activities in our committees, where it starts with service on the Standardization Committee in 1925; with a tremendous history of service as officer or as member of an administrative committee or a technical committee during the ensuing 30 years. While mumbers do not measure this, the record shows service on one or another technical committee for 14 years, and a total amount of service on administrative committees and as officer of the Institute aggregating 60 one-year terms of service. Not all these one-year terms represent equal amounts of burden; the year as President is far and away the greatest, but the 10 years of service on the Board of Directors and on its Executive Committee represent contributions far beyond that of service on one of the less burdensome administrative committees.

Another thread in the career which should be picked up perhaps at this point is that of active work on behalf of trade associations. While this thread takes Dr. Baker to a number of these, his work for the association that has now come to be called the Radio-Electronic-Television Manufacturers Association has represented by far the greatest demand on his time of any of his pieces of trade association work. He has served as Chairman of the Engineering Division or as Director of Engineering continuously from 1934 to the present, and, for the same length of time, has been a Director of the Association. He is currently serving as its President as well.

To pick up another thread, Dr. Baker's original Bachelor of Engineering degree of 1916 had been supplemented by the degree of Master of Electrical Engineering in 1918. In 1935, his Alma Mater, Union College, conferred upon him the honorary degree of Doctor of Science. In 1951, Syracuse University was to confer upon him the honorary degree of Doctor of Engineering for his many services to his profession.

Another thread which needs to be woven into our story is one which tells of the successful managing of the preparation of unbiased technical advice for the use of a government agency in situations where members of the industry were almost "prepared to pull each others' hair." In the late 1930's, the Radio Manufacturers Association had made recommendations to the Federal Communications Commission for standards for monochrome tv broadcasting. There was enough dissent from these recommendations to lead the FCC to refuse to act upon the recommendations. With the support of the then Chairman of the FCC, the Honorable Lawrence Fly, Dr. Baker organized the National TV System Committee, served as its Chairman, and brought its work along to the point where a substantially unanimously accepted report was submitted to the FCC late in 1940. The adoption of this report by the FCC laid the foundation for our present highly successful monochrome tv broadcasting.

This thread of service to governmental agencies concerned with regulation continued almost without let-up from that time on. Even during World War II it became recognized that much of the regulatory structure would need revision at the end of the war and so a Radio Technical Planning Board was formed to give at least preliminary consideration to matters of this sort. As usual, someone who was willing to put forth the effort had to do it and, again, Dr. Baker served as Chairman of this organization.

When the interest of the television broadcasting and tv receiver manufacturing industry in color television became highly active toward the end of the 1940's, proposals for Federal establishment of broadcasting standards commenced to appear. Standards for color tv broadcasting were in fact established by the FCC late in 1950, but the technical information upon which these standards were based did not represent the consensus of the entire body of engineers of experience in this field. A second National Television System Committee was assembled to review this whole matter and to develop a new set of recommendations for standards for color tv broadcasting. The work of this committee, of its panels and subcommittees, and the work done in their support in the laboratories of the interested organizations is estimated to have represented well over one million man-hours of engineering effort. As a result of this work, recommendations submitted to the FCC by the NTSC and supported by a demonstration given by the NTSC led to the adoption of standards for color tv broadcasting based on the NTSC recommendations. Our prospect for color tv broadcasting with an extremely high grade of ultimate performance potentiality in this country is wholly a consequence of the recommendations of this second NTSC.

While Dr. Baker was doing these other things, he somehow found time to establish the Electronics Division of his Company in new, and, initially, very spacious quarters in Electronics Park just outside the city of Syracuse, N. Y. This operation was a very forward-looking step; the organization did not outgrow the new quarters for 3 or 4 years! But in the course of that outgrowing, or perhaps shortly after it, the Electronics Division moved into a place of major importance in its own company and, in addition, into places of major importance in a number of the fields in which it supplied products. Indeed, so extensive is the radio enterprise built by Dr. Baker that it is now larger than the General Electric Company itself was only a decade ago.

The thread of service to IRE has not been fully explored in the preceding remarks. As World War II drew to a close it became obvious that the IRE would need a new headquarters in order to keep pace with the rapid increase in membership. Dr. Baker was a key figure in making the Building Fund Campaign a success, obtaining funds from manufacturers and individuals in sufficient amounts to establish the IRE in its present headquarters. At about the same time, the view commenced to develop in the minds of a few of the far-sighted members of IRE that our profession was not only increasing greatly, but also diversifying its interests at a tremendous rate. It started to be recognized that either an executive means of dealing with this would have to be found or else there would no longer be a single radio or electronic engineering society. Probably Dr. Heising was the first to fully recognize the need of the situation and to suggest a series of steps for dealing with it. And among the early supporters of what became the IRE Professional Group system was Dr. Baker. His interest in this development within our society has continued to grow and he has served as Chairman of the the Professional Groups Committee continuously from 1950 to the present time.

An indication of the esteem in which Dr. Baker has been held is the presentation to him of the IRE Medal of Honor in 1952, with the citation, "In recognition of his outstanding direction of scientific and engineering projects; for his statesmanship in reconciling conflicting viewpoints and obtaining cooperative effort and for his service to the Institute."

As engineer, as industrialist, as wise administrator of and counselor to a professional society, and as leader of men, the donor of the Baker Award represents an example which those considered as potential recipients of the Award in the future may well look up to.

Acknowledgment

The author is especially grateful to Ellsworth D. Cook, a Fellow of the IRE, who through his long association with Dr. Baker at the General Electric Co., both as a friend and a colleague, was able to contribute substantially to the accuracy and completeness of this account.

Recipients of the W. R. G. Baker Award

Awarded to the author(s) of the best paper published in the IRE Transactions of the Professional Groups. This Award is being given for the first time this year.

R. J. Kircher........... 1957   R. L. Trent............ 1957   D. R. Fewer........... 1957

Management of Large R & D Organizations*

NATHAN I. HALL†, FELLOW, IRE

Summary—**This paper deals primarily with the organization and management of large industrial research and development organizations. However, many of the techniques described are equally applicable to smaller organizations.**

The paper discusses the importance of building a high caliber scientific staff and then providing it with interesting research and development projects in the midst of a pleasant professional environment. An organizational technique emphasizing adequate organization charting is described. The subjects of salaries, bonuses, and special privileges are considered, and special techniques for minimizing problems in these areas are suggested.

The paper concludes with a discussion of what it is that an R and D director expects of his managers at various levels, and the attributes a man must possess to advance rapidly into the managerial ranks of a large R and D organization.

THE LAST few years have seen a tremendous increase in the size and complexity of electronic systems. This is particularly true in the case of military weapon systems. For example, the giant search radars, data transmission systems, computing centers, surface-to-air guided missiles, interceptor squadrons, radar fire-control systems, and air-to-air guided missiles of our air defense system are extremely complex. The development of these systems and subsystems requires integrated engineering organizations of many hundreds, and often many thousands of technical and supporting people.

One of the largest organizations in the country devoted to electronic systems development is at Hughes, where over 5000 R & D people are working as a single unit on closely related electronic systems. In a rapidly growing organization of this type, the major problems are those of obtaining outstanding engineers and physicists, holding onto them after they are obtained, organizing them into an efficient organizational structure, and then managing them effectively.

The Scientific Staff

Given a reasonably enlightened management and reasonably adequate financing, the success of today's electronics company will be largely a function of the quality of its scientific staff. Management's foremost problem is to acquire and hold increasing numbers of well-trained technical people. This problem has been increased by a large factor in recent years because of the shortage of engineers and physicists.

The engineering shortage exists at all quality levels. However, it is the man in the top 5 or 10 per cent bracket who is in greatest demand. It is such men who give one company an edge over its competition. Therefore when we talk about obtaining and holding technical personnel, we are interested to a degree in all quality levels, but particularly in the higher levels.

What Engineers Want

The best way to get and keep a topnotch man is by discovering what it is that he really wants, and then equal, if not beat, one's competitors in giving it to him. Incidentally, this is much easier to do if management has come up through the engineering ranks and not only understands but is composed of engineers.

How important is a high salary to an engineer? Undoubtedly, it is quite important. Money is the thing he uses to buy his house and to send his kids to college. Money brings the things which symbolize success, and engineers are few and far between who are so dedicated to scientific pursuits that income is unimportant to them. If anyone doubts this, he need only look at the problem which faces our universities, caused by their lower-than-industry salaries.

Because of our high income taxes, today's corporations are providing an ever increasing percentage of total income in the form of fringe benefits which are not taxable, or which enjoy lower tax rates. Stock option plans are extremely popular these days. Company-financed retirement plans, longer paid vacations, added sick benefits, and larger insurance policies are the rule of the day.

Income is clearly an important factor in obtaining and keeping a high grade technical staff. However, its importance in comparison with other factors should not be overestimated. It is observed that certain engineering organizations which are noted for paying substantially higher than average salaries also have higher than average turnover ratios. Furthermore, the astute engineer with several offers of employment frequently passes up the higher income offers and accepts the offer of a company which provides those things which are accumulatively more important to him than the extra money.

In considering what an engineer really wants, interesting work and lots of it ranks high on the list. The engineer who is given a steady flow of work which he finds intensely interesting, and for the doing of which he receives adequate recognition, seldom goes around looking for another job. In this connection, however, he wants to see evidence of advance planning, research, new proposals, and the like so that he feels assured of his future. It is most important that management keep its engineers adequately informed, not only of its current successes but of its forward planning. The manager who feels that such matters are none of the engineers' business is quite out of date by today's standards.

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† Hughes Aircraft Co., Culver City, Calif.

An engineer can be paid a good salary with regular increases, assigned work of great interest to him, and yet become dissatisfied if the years roll by and he appears to be not advancing in the organization. This problem would not exist if it were possible to promote every good man to a successively higher organizational position every year or two. Even for a company that is growing rapidly, there just aren't enough supervisory positions available to even come close to such an ideal.

It is a wise organizational procedure to have a second in command available who can step into his boss's shoes. For large organizational units, more than one alternate is advantageous. An outstanding section head who is overdue for a promotion need not be kept waiting until the position of department head is open. His talents may be utilized to wider advantages by making him actually an assistant department head, as well. This gives the section head the recognition among his fellows which he deserves, while the very real added assistance he brings to the department head may well make it possible to lighten the latter's burdens so that his full talents may be more broadly capitalized upon.

If the growth of a department and the development of the men within it obviously warrants, still an additional, and somewhat more elevated position—that of associate department head—may be created, thus opening up still another very real and apparent promotion and spreading the burden of growing executive responsibility still further.

The point here is that the R & D organization which has the problems of its outstanding employees constantly in mind will not sit idly by and let these people grow dissatisfied. It will deliberately seek and it will find means of recognizing good performance in ways that will not only add to the general efficiency but make it plain to all that the man with ability and determination does get ahead.

In many engineering organizations the only way to get ahead is to work up through the supervisory ranks. This discriminates against the man who makes a large technical contribution and whose talents would not be properly used in a high supervisory position. Here again, salary increases alone do not usually fulfill the man's desire to get ahead. His salary is not public knowledge. He wants recognition which can be seen by his peers, his wife, and his friends. The senior technical man should be recognized with an appropriate title such as senior staff engineer, senior staff physicist, etc., and he should not only receive added privileges but should take part in management discussions.

The senior technical man should be encouraged to write and present technical papers. Part-time teaching in local universities provides many benefits to the man, to his company, and to the university. One of the greatest advantages from the company's standpoint is the inside track which teaching provides with respect to hiring topnotch graduates.

The subject of recruiting scientific personnel is too broad to be covered in this paper. However, it should be mentioned that one of the best recruiting techniques is that of instituting university fellowship programs.

A discussion of what an engineer wants would not be complete without touching upon the nature of the facility in which he works. The engineer should be proud of his company. He will find this to be difficult if he is stuffed into a large barn-like area with his desk in a row of 100 others. The top R & D laboratories of the country have found by years of experience that creative work is best performed in offices containing one, two, or perhaps four people. They have also learned that money spent on air conditioning, sound proofing, adequate telephone service, and the like, in engineering areas, will in the long run result in larger profits for the company. For the more senior people, executive dining room privileges, name parking spaces, private secretaries, and other privileges of this nature are of great importance in encouraging outstanding work, contentment, and a low turnover ratio.

Organization

As an organization grows larger and larger, the importance of organizing efficiently becomes very great. Consider a large R & D organization operating at the rate of 100 million dollars a year. It is all too easy to allow inefficiencies to creep in which reduce the effectiveness of the average person in the organization to 80 or 90 per cent of his capability. This in turn means a loss in output of ten to twenty million dollars a year. In such a case, even a small increase in organizational efficiency may save many millions of dollars.

The purpose of organization is to place each person in a position for which he is well suited, to keep him continuously informed as to what is expected of him, and to supervise him sufficiently so that he will indeed work at maximum efficiency. It is much more difficult to effectively supervise 1000 people, than it is to supervise 20. Therefore, a popular way to avoid the complication of managing large numbers of people in a single unit is to separate them by projects. The virtues of such an arrangement are obvious. A project engineer supervising a few dozen people will not only be able to know every person personally, but he will know in considerable detail what each person is doing.

The real problem occurs when the projects get very large and interrelated. Consider an R & D organization which is simultaneously developing half a dozen different guided missiles with anywhere from 100 to 500 people working on each missile. If the organization were divided into half a dozen different project groups, the result will be six separate groups of aerodynamicists, six propulsion groups, six airframe groups, etc. Such an arrangement would be quite inefficient.

In order to attract and hold topnotch experts in such fields as aerodynamics or electronics, it is necessary that the top man in the group be an outstanding man in his field. Imagine the difficulty of obtaining half a dozen nationally-known aerodynamicists to head that many small aerodynamics groups, just so that each missile project might have its own group. It is much more practical to hire one top expert and have him direct the aerodynamics work for all the guided missile projects.

Consider the highly technical field of systems analysis. Capable physicists and engineers with experience in this field are particularly hard to find. A high proportion of them have doctor's degrees. Should a company try to separate its men of this type into many separate groups, each associated with a separate missile project? The speaker's experience indicates that it is highly desirable to group all of the systems analysis people in one department, directed by an outstanding man in this field. This mode of operation not only makes more effective use of the talents of the individuals, but the people involved are happier with this arrangement.

Still another advantage of departmentalization by skills is that it promotes desirable standardization among the projects. If a single department is designing power supplies for a number of guided missiles, it is likely that acceptable compromises will be found which will allow a single power supply to be used in several different missiles. Such compromises would be much less likely to occur if separate project groups were each designing their own power supplies.

A weakness of dividing an organization into project groups is the morale problem which results as a project nears completion and the work load falls off. The project engineer has an inclination to dig up unnecessary jobs to keep his people busy as long as possible. The people become restless and try, perhaps prematurely, to transfer to expanding projects. Departmentalization by skills almost completely solves this problem.

Project Managers

When an organization with many projects is departmentalized according to skills, project coordination becomes of great importance. Our experience has shown a project-manager system to be quite effective. Men are chosen for project-manager position who have seniority and capabilities equal to those of the department heads. A project manager may be concerned with only a single project if it is a very large one. Generally, however, a project manager is able to concern himself with two to three closely related projects.

Consider, as a specific example, our Guided Missile Laboratories. This organization, composed of about 1500 people, is headed by a director, an associate director, and a technical director. Reporting directly to the directors are eight department heads, each a specialist in his field. The head of the Electronics Department is responsible for the electronic portions of all of the missiles under development, and the head of the Flight Operations Department is responsible for all flight testing, etc.

Three project managers, who also report to the directors of the Guided Missile Laboratories, have cognizance over a number of guided missiles now under development. The project managers have no one reporting directly to them other than their secretaries and one or two assistants. They maintain liaison with the customers, see that design objectives are set and met, monitor contracts, chairman interdepartmental technical meetings, etc.

Management Techniques

One of the most effective aids in the management of a large R & D organization is the organization chart. Several of the needs for an accurate charting of the organization are so obvious that they need not be mentioned. A very important, and less obvious need, is the constant needling that organization charts give to the technical department heads to improve the efficiency of their departments. The technically competent department head has a tendency to do too much of the work himself and spend too little time organizing his department for maximum efficiency. When required to reissue his organization chart periodically and have it approved by his superiors, a department head must give his organizational structure the attention which it would probably not otherwise receive.

A uniform system of personnel titles and a close control of the approval of promotions to titled positions is essential where thousands of people are involved. In our Weapon Systems Laboratories, this control is considered so important that each and every promotion into or within the titled ranks (about 170) must be personally approved by the director of the laboratories. This is done in weekly coordination meetings involving the directors of all the major units of the organization.

A salary curve system has been found to be a valuable aid in maintaining a uniform salary structure throughout a large organization. Let us assume that Joe and Bill are of equal value to the company but no one person in the organization knows them both. The problem is how to pay Joe and Bill the same salary and give each the same raise at review periods.

A family of salary curves may be used in which salary is plotted against experience in years. The various curves represent persons of varying capabilities, and once a man is established as belonging on a particular curve, the amount of each periodic salary increase is automatically established. A man's rating may of course be changed if it appears at any time to be out of line with his abilities. The department heads should compare notes with each other occasionally so as to have a uniform idea as to what constitutes a 75 man or a 90 man. The salary curve structure, along with departmental salary budgets, forms a superior method of salary control.

Qualities of Leadership

Occasionally one finds a genius who has successfully scaled the organizational ladder to a high executive position. However, geniuses often find it tough going in a large R & D organization, since they generally lack certain important qualities of leadership.

A good executive must have confidence, but he must also have humility. He will surround himself with smart people and will see to it that they get full credit for what they do. The successful executive seeks and makes good use of the advice of others.

The outstanding engineering executive has learned to mould his organization around his key people rather than to fit the people to the organization. He knows how to successfully use a "prima donna" where a lesser man would have fired him. The question is not, "Is he a prima donna?" but, "Can he sing?"

Your speaker has worked in large R & D organizations for over 20 years and has observed many men climb up to the top. Few could be termed brilliant. However, all had what is known as drive—that inner urge which will not admit failure. When such men have tried nine times to solve a problem and have failed, they are all ready to try a tenth time.

There is an old saying, "Hitch your wagon to a star." This is probably good advice if the "star" is properly chosen. However, the author has observed that those who are promoted most rapidly are those who devote themselves completely to the job at hand, and who will tell you that they aren't anxious to take on larger responsibilities. On the other hand, it has been observed that the unhappy engineer is generally the one who overestimates his capabilities and is constantly maneuvering for a promotion for which he is not ready.

To be considered really successful, an engineer must learn how to live as well as how to make his way up the supervisory ladder. Charles Schwab, chairman of one of the largest steel companies, lived on borrowed money for the last five years of his life and died broke. Samuel Insull, president of one of the largest utilities, died penniless in a Paris subway station. One of the greatest wheat speculators of all times, Arthur Cutten, died abroad and insolvent. Albert Fall, a cabinet member, was pardoned from prison so he could die at home. Leon Fraser, president of a great international bank, committed suicide. And so did Ivar Kreuger, the match king. All of these men had learned how to make money, but not one of them had learned how to live.

Synthesis of Tchebycheff Parameter Symmetrical Filters*

ALEXANDER J. GROSSMAN†, MEMBER, IRE

The problem of bridging the gap between theory and practice has proved particularly difficult in the field of circuit theory. After the theoretical analysis of a problem has been published it may be years before methods of computation are worked out which can be used in a straightforward manner by practical engineers. It is hoped that the following paper, showing how to carry through the design of Darlington filters in important practical cases, finally will make the superior performance of these filters generally available. This paper is heartily welcomed and it is hoped that papers will be forthcoming showing how to apply other developments in circuit theory to the solution of practical problems.—*The Editor*

***Summary*—This paper consists of two parts. The first part is tutorial and describes at an elementary level one of the contributions made by Darlington.[1] This is the design theory of electrically symmetrical reactive (lossless) networks with particular attention to filters which exhibit Tchebycheff type performance in the pass and stop bands. The second part of the paper is the presentation, in "handbook" style, of step-by-step procedures to be followed in the design of filters of the above type. Emphasis is placed on the use of rapidly converging series in the computations in place of elliptic function tables.**

A METHOD of designing four-terminal reactance networks on an insertion loss basis is described by Darlington.[1] The purpose of the present paper is two-fold. The main part of the text is an elementary and detailed explanation of one of the important contributions contained in the original. It is hoped that this will give the average reader a better understanding of this particular part of Darlington's rather compactly written paper. The second purpose is to provide a set of design charts which enables one to go from the specification of filter performance to the component values by following a step-by-step procedure. The scope of the paper is severely limited both

* Original manuscript received by the IRE, November 20, 1956.
† Bell Telephone Labs., Inc., Murray Hill, N. J.
[1] S. Darlington, "Synthesis of reactance 4-poles," *J. Math. and Phys.*, vol. 18, pp. 257–353; September, 1939.

in respect to network configuration and type of insertion loss characteristic. It is assumed that the network is electrically symmetrical, that is, it is not possible to distinguish between the two pairs of terminals (or the two ports) by measurements made at these pairs of terminals. This implies that the filter is to be inserted between equal resistance terminations. It is further assumed that the network is reactive (or lossless), that is, the components are pure inductors and capacitors. The particular insertion loss characteristic to be considered is known as the Tchebycheff parameter type. It is defined as follows: the minima of loss between the infinite loss peaks in the attenuation band are equal to a value specified in advance; the loss in the pass band ripples between a minimum of zero and a maximum value that is likewise specified in advance. In other words, the filter introduces a specified minimum discrimination between frequencies in the attenuation band and those in the pass band and has a prescribed maximum allowable distortion in the pass band.

General Relations for a Symmetrical Network

The first step in the insertion loss design method is the choice of a function which not only satisfies the desired insertion loss requirements but is also physically realizable. In the case of an electrically symmetrical network, the conditions which such a function must satisfy are established most conveniently by analysis of the lattice network, even though an equivalent ladder may be preferred for actual construction. As shown in Fig. 1, the

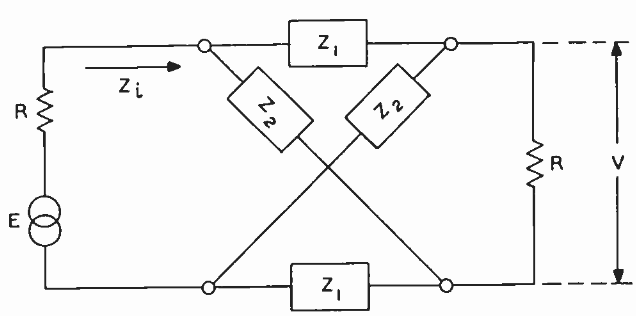


Fig. 1—Symmetrical terminated lattice network.

lattice, consisting of the two-terminal reactance networks designated Z_1 and Z_2, is inserted between the equal resistances, R. The complex insertion voltage ratio is defined as:

$$e^{\theta} = \frac{V_0}{V},$$

where e denotes the base of naperian logarithms; $\theta = \alpha + j\beta$, with α the insertion loss and β the insertion phase shift; V_0 is the voltage appearing across the load resistance when the generator and load are connected directly; V represents the voltage across the load after the network is inserted between the generator and load. It is easy to demonstrate that:

$$e^{\theta} = \frac{V_0}{V} = \frac{R^2 + R(Z_1 + Z_2) + Z_1Z_2}{R(Z_2 - Z_1)}. \tag{1}$$

The insertion loss, α, is found from the "insertion power ratio," which is defined as the square of the magnitude of the complex voltage ratio, or:

$$e^{2\alpha} = \left|\frac{V_0}{V}\right|^2. \tag{2}$$

The necessary conditions that this function must satisfy may be derived from the fact that the impedances Z_1 and Z_2 are pure reactances. According to Foster's theorem, each may be represented as a ratio of two polynomials in the frequency variable $p = j\omega$. The coefficients are real constants and the degree of the numerator is one more or less than the degree of the denominator. Thus,

$$\frac{Z_1}{R} = \frac{pN_1}{D_1}, \tag{3a}$$

$$\frac{Z_2}{R} = \frac{N_2}{pD_2}, \tag{3b}$$

when N_1, N_2, D_1, D_2 are *even* polynomials in p with real coefficients. It is noticed that these expressions are written in an arbitrary way; that is, the numerator of Z_1 is an odd polynomial and the denominator an even polynomial, and conversely for Z_2. If this does not happen naturally in a particular case, it can be effected by multiplying numerator and denominator by the variable p. This convention is introduced to get the final results in a standard form.

Eq. (1) for the insertion voltage ratio becomes:

$$\frac{V_0}{V} = \frac{(N_2D_1 + p^2N_1D_2) + p(N_1N_2 + D_1D_2)}{(N_2D_1 - p^2N_1D_2)}. \tag{4}$$

By computing the square of the absolute value, we obtain the important result that:

$$e^{2\alpha} = 1 + \left[\frac{\omega S}{P}\right]^2, \tag{5}$$

where,

$$S = D_1D_2 - N_1N_2$$
$$P = N_2D_1 - p^2N_1D_2,$$

and consequently are *even* polynomials in the frequency variable and have real coefficients. This equation states that it must be possible to express a given insertion loss characteristic in this particular form if the corresponding network is to be realizable as an electrically symmetrical reactive network inserted between equal resistances. A simple illustration may serve to emphasize this point. Suppose we wish to design a low-pass filter which has a maximally flat insertion loss characteristic. Assume that the design requirements are satisfied by:

$$e^{2\alpha} = 1 + \omega^6.$$

Comparison with (5) shows that $S=\omega^2$, and $P=1$, and so it is possible to find a symmetrical network which will introduce this insertion loss. On the other hand, if a somewhat greater loss is needed in the attenuating band, one might be tempted to try:

$$e^{2\alpha} = 1 + \omega^8.$$

Here, however, $S=\omega^3$, an odd polynomial in terms of the frequency variable. This characteristic cannot be realized by a symmetrical reactive network.

After the desired insertion power ratio is expressed in the normal form (5), we are ready to find the values of the network components. The procedure proposed by Darlington is based on the following reasoning: there is a relation between the insertion power ratio and the square of the magnitude of the reflection coefficient of a reactive network inserted between resistance terminations; it is possible to determine the complex reflection coefficient from the square of its magnitude; the lattice impedances can be expressed in terms of the polynomials appearing in the insertion power ratio and the complex reflection coefficient; finally, a knowledge of these impedances leads to the component values.

The relation between the insertion power ratio and the magnitude of the reflection coefficient may be established in the following way. Let the input impedance of the terminated symmetrical network be $Z_i=R_i+jX_i$, as indicated in Fig. 1. The input current is

$$I_1 = \frac{E}{R + R_i + jX_i}.$$

The input power is:

$$P_i = \frac{|E|^2 R_i}{(R + R_i)^2 + X_i^2},$$

and the available or reference power is:

$$P_m = \frac{|E|^2}{4R}.$$

The difference between these is the reflected power. The ratio between reflected and available power is:

$$\frac{P_m - P_i}{P_m} = \frac{(R - R_i)^2 + X_i^2}{(R + R_i)^2 + X_i^2} = \left|\frac{R - Z_i}{R + Z_i}\right|^2,$$

which is the square of the magnitude of the reflection coefficient. Now, for a purely reactive network, the power delivered to the load is equal to the power supplied to the network. Also, the insertion power ratio for a symmetrical network is the ratio of available power, P_m, to delivered power, P_i. Therefore, the relation which was to be established is:

$$\left|\frac{R - Z_i}{R + Z_i}\right|^2 = 1 - e^{-2\alpha} = \frac{-p^2S^2}{P^2 - p^2S^2} \qquad (6)$$

where the polynomials P and S are prescribed in advance as indicated by (5).

The connection between the magnitude of the reflection coefficient and the network components is furnished by the complex reflection coefficient. To show this, we start with the input impedance of the terminated lattice, Fig. 1:

$$Z_i = \frac{R(Z_1 + Z_2) + 2Z_1Z_2}{2R + Z_1 + Z_2}.$$

The branch impedances of the lattice are written in the form of (3a) and (3b) to get:

$$\frac{Z_i}{R} = \frac{(N_2D_1 + p^2N_1D_2) + 2pN_1N_2}{(N_2D_1 + p^2N_1D_2) + 2pD_1D_2}.$$

Then the complex reflection coefficient is:

$$\frac{R - Z_i}{R + Z_i} = \frac{pS}{A + pB}, \qquad (7)$$

where,

$$A = N_2D_1 + p^2N_1D_2$$
$$B = D_1D_2 + N_1N_2,$$

are *even* polynomials in the frequency variable and have real coefficients. Now there remains the problem of finding the polynomial $A+pB$ based on the fact, required by (6) and (7), that $A^2-p^2B^2=P^2-p^2S^2$.

The solution to this problem is based on the observation that the complex insertion voltage ratio (4) may be written in the form:

$$\frac{V_0}{V} = \frac{A + pB}{P}. \qquad (8)$$

The roots of the numerator in terms of the frequency parameter, p, are the natural modes of the terminated network. It is well known that the real part of a natural mode cannot be positive since this is the damping constant. Hence, the roots of $A+pB$ are those roots of $P^2-p^2S^2$ that do not have a positive real part. These selected roots specify the polynomial completely except for a constant multiplier.

The selection of the roots proceeds in the following way. Since $P^2-p^2S^2=(P+pS)(P-pS)$, and P is an even and pS is an odd polynomial, the roots of $(P+pS)$ are the negatives of the roots of $(P-pS)$. Hence, if we calculate the roots of one factor, those of the other factor are known at the same time. It is recalled that the even polynomials S and P are obtained from the specified insertion power ratio,[2] and it is a simple matter to form the polynomial $(P+pS)$. The roots of this polynomial must be determined. Then they are sepa-

[2] For completeness, a plus and minus sign should be associated with S and P. The various choices correspond to a network and its inverse and the interchange of the lattice impedances Z_1 and Z_2. These options will be disregarded here.

rated into two groups. The first group contains those roots that do not have a positive real part; they are then considered to be the roots of a polynomial which is denoted by P_1+pS_1. The second group contains those roots that do have a positive real part; they are used to form a polynomial designated as P_2-pS_2. Thus, we have:

$$P + pS = (P_1 + pS_1)(P_2 - pS_2). \tag{9}$$

Since the roots of the polynomial $(P-pS)$ are the same as those of $(P+pS)$ except for a reversal in sign, we also have:

$$P - pS = (P_1 - pS_1)(P_2 + pS_2). \tag{10}$$

With these relations, we may write:

$$(A + pB)(A - pB) = (P_1 + pS_1)(P_2 - pS_2)(P_1 - pS_1)(P_2 + pS_2).$$

By selecting those polynomial factors whose roots do not have a positive real part, the following solution is obtained:

$$A + pB = (P_1 + pS_1)(P_2 + pS_2). \tag{11}$$

This important result states that the polynomial $(A+pB)$ is formed from those roots of $(P+pS)$ which do not have positive real parts and the negatives of those roots of $(P+pS)$ that do have positive real parts.

There is now sufficient information available for determining the lattice impedances as given by (3a) and (3b). Using the definitions of S and P written below (5) and those of A and B below (7), it is seen that:

$$\frac{Z_1}{R} = \frac{A - P}{p(B + S)}$$
$$\frac{Z_2}{R} = \frac{A + P}{p(B + S)}. \tag{12}$$

It turns out that these results may be simplified by using the expanded forms of (9) and (11), namely:

$$\begin{aligned} P &= P_1P_2 - p^2S_1S_2 \\ S &= S_1P_2 - S_2P_1 \\ A &= P_1P_2 + p^2S_1S_2 \\ B &= S_1P_2 + S_2P_1. \end{aligned} \tag{13}$$

Upon introducing these expressions in (12), the lattice impedances are specified by:

$$\frac{Z_1}{R} = \frac{pS_2}{P_2}$$
$$\frac{Z_2}{R} = \frac{P_1}{pS_1}. \tag{14}$$

A recapitulation and an illustration may be helpful at this point. The design process consists of the following steps:

1) The desired loss characteristic must be such that the corresponding insertion power ratio can be expressed in the form of (5), and it must be written in this form;
2) Determine the roots of the polynomial $(P+pS)$;
3) From those roots that do not have a positive real part, form the polynomial (P_1+pS_1);
4) Reverse the sign of those roots that do have a positive real part and form (P_2+pS_2);
5) The lattice impedances are calculated from (14).

As an illustration, consider the maximally flat insertion loss characteristic, for which the insertion power ratio is:

$$e^{2\alpha} = 1 + \Omega^6 = 1 - p^6,$$

where $\Omega=\omega/\omega_0$, $p=j\Omega$, and ω_0 is the reference frequency (in radians per second) at which loss is 3 db. This is in the form of (5) with $S=\Omega^2=-p^2$ and $P=1$. The roots of

$$P + pS = 1 - p^3$$

are the values of p for which the value of this polynomial is equal to zero. They are:

$$\begin{aligned} p_0 &= 1 \\ p_1 &= -0.500 + j0.866 \\ p_1^* &= -0.500 - j0.866. \end{aligned}$$

Since p_1 and p_1^* do not have positive real parts, they are used to form:

$$P_1 + pS_1 = (p - p_1)(p - p_1^*) = p^2 + p + 1.$$

The sign of p_0 is reversed to form:

$$P_2 + pS_2 = [p - (-p_0)] = p + 1.$$

By sorting out the even and odd parts of these polynomials, we have:

$$\begin{aligned} P_1 &= p^2 + 1 \\ P_2 &= 1 \\ pS_1 &= p \\ pS_2 &= p. \end{aligned}$$

The lattice impedances, in accordance with (14), are:

$$\frac{Z_1}{R} = p,$$
$$\frac{Z_2}{R} = \frac{p^2 + 1}{p}.$$

Hence, the lattice arm Z_1 is an inductor having the value R/ω_0 henries, and Z_2 is a series tuned circuit made up of an inductor equal to R/ω_0 henries and a capacitor equal to $1/R\omega_0$ farads. If the inverse network consisting of a capacitor in the Z_1 arm, and an antiresonant circuit in the Z_2 arm, is preferred, this may be evaluated by using the reciprocals of the above expressions. Two other possibilities exist and they are obtained by interchanging Z_1 and Z_2 in each case.

Tchebycheff Parameter Insertion Loss Characteristic

For ease of exposition, attention will be directed to the design of a two-section low-pass filter. The Tchebycheff parameter insertion loss characteristic is shown in detail in Fig. 2. The pass band extends from zero to the frequency f_p; the loss ripples between zero and the prescribed equal maxima α_p. The attenuation band extends from the frequency f_a to infinity; the loss oscillates between an infinite value and the prescribed equal minima α_a. These bands are joined by the transition band extending from f_p to f_a. It is convenient to define a reference frequency, $f_0=\sqrt{f_p f_a}$, which is somewhat analogous to the cutoff frequency used in image parameter filter design. Again, as a matter of convenience, we work with the normalized frequency variable $\Omega=\omega/\omega_0=f/f_0$, and shall call this (inaccurately) the frequency. Then, as indicated in Fig. 2, the pass band terminates at $\Omega=\sqrt{k}$, and the attenuation band begins at $\Omega=1/\sqrt{k}$, where $k=f_p/f_a$, the selectivity parameter, by definition.

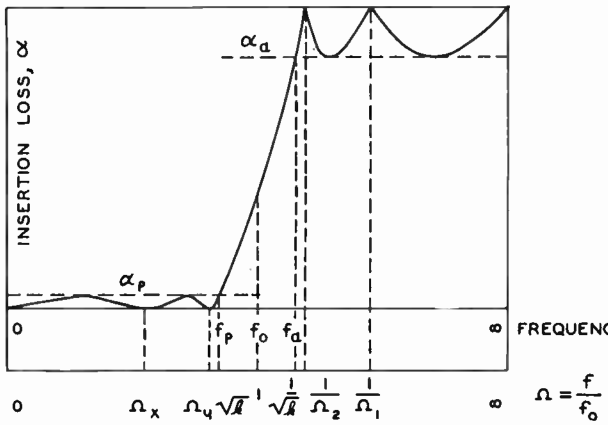


Fig. 2—Tchebycheff parameter insertion loss characteristic.

This loss characteristic may be represented by the insertion power ratio:

$$e^{2\alpha} = 1 + (e^{2\alpha_p} - 1)\left[F(\Omega)\right]^2. \tag{15}$$

The function $F(\Omega)$ has the following characteristics: 1) it is equal to zero at $\Omega=0$, Ω_x, Ω_y; 2) its magnitude is unity at $\Omega=\sqrt{k}$, and at two frequencies which are separated by the zero loss points; 3) it is infinite at $\Omega=1/\Omega_1$, $1/\Omega_2$, ∞; 4) its magnitude is equal to $\sqrt{(e^{2\alpha_a}-1)/(e^{2\alpha_p}-1)}$ at $\Omega=1/\sqrt{k}$, and at two frequencies which are separated by the infinite loss points. In addition, it must be of the form $(\Omega S/P)$, where S and P are even polynomials in Ω, if it is to be realized by an electrically symmetrical network. All these requirements can be satisfied by

$$F(\Omega) = F_0 \frac{\Omega(\Omega^2 - \Omega_x^2)(\Omega^2 - \Omega_y^2)}{(1 - \Omega_1^2\Omega^2)(1 - \Omega_2^2\Omega^2)} \tag{16}$$

provided F_0 and the frequencies of zero and infinite loss are chosen properly.

The determination of these parameters is accomplished in a rather indirect manner by studying certain functions derived from (16). The function $(1-F^2)$ has a single root at $\Omega=\sqrt{k}$ and double roots at those frequencies within the pass band where the magnitude of F is unity. Another derived function, $(1-k_1^2F^2)$, where $k_1^2=(e^{2\alpha_p}-1)/(e^{2\alpha_a}-1)$, has double roots at the equal minima of F^2 between the frequencies of infinite loss and a single root at $\Omega=1/\sqrt{k}$. A third function, $dF/d\Omega$, has single roots at the maxima of F^2 in the pass band and at the minima of F^2 in the attenuation band. Now, if this derivative is divided by the square root of the product of the other two functions, one will find that

$$\frac{dF/d\Omega}{\sqrt{(1-F^2)(1-k_1^2F^2)}} = \frac{\pm M_0}{\sqrt{(1-\Omega^2/k)(1-k\Omega^2)}},$$

because the single roots of $dF/d\Omega$ cancel the square root of the double roots of the other functions, leaving only the single roots at $\sqrt{k}$ and $1/\sqrt{k}$ and a constant multiplier, M_0. This equation may be expressed in terms of definite integrals as follows:

$$\int_0^F \frac{dx}{\sqrt{(1-x^2)(1-k_1^2x^2)}} = \pm M_0\sqrt{k}\int_0^{\Omega/\sqrt{k}} \frac{dy}{\sqrt{(1-y^2)(1-k^2y^2)}} + C_1$$

where x and y are variables of integration and C_1 is a constant of integration. These integrals are known as elliptic integrals of the first kind. By making the transformations:

$$x = \sin\theta_1$$
$$F = \sin\phi_1$$
$$y = \sin\theta$$
$$\Omega/\sqrt{k} = \sin\phi$$

this is put in the more desirable form:

$$\int_0^{\phi_1} \frac{d\theta_1}{\sqrt{1-k_1^2\sin^2\theta_1}} = \pm M_0\sqrt{k}\int_0^{\phi} \frac{d\theta}{\sqrt{1-k^2\sin^2\theta}} + C_1.$$

The solution of this equation, giving the relation between F and Ω may be written in the form of a pair of simultaneous equations:

$$\Omega = \sqrt{k}\,\mathrm{sn}\,(u, k)$$
$$F = \mathrm{sn}\,(\pm Mu + C_1, k_1) \tag{17}$$

where

$$u = \int_0^{\phi} \frac{d\theta}{\sqrt{1-k^2\sin^2\theta}}$$

$$\pm Mu + C_1 = \int_0^{\phi_1} \frac{d\theta_1}{\sqrt{1-k_1^2\sin^2\theta_1}}$$

$$\mathrm{sn} = \text{elliptic sine}.$$

The constants in (17) are evaluated by examining some of the properties of the elliptic functions. A brief discussion of elliptic functions, sufficient for the present purpose, is given in the appendix.

The real period of sn (u, k) is $4K$, where K is the complete elliptic integral of modulus k. The sn function goes through a complete cycle of values between $+1$ and -1 as u varies from 0 to $4K$, and this cycle is repeated as u takes on larger values or becomes negative. Hence, the values of Ω obtained from (17) for u real lie between $+\sqrt{k}$ and $-\sqrt{k}$. This corresponds to the pass band and its image at negative frequencies. In particular, the pass band at positive frequencies is traversed once by values of u lying between 0 and K, the quarter period. On the other hand, the real period of sn $(\pm Mu+C_1, k_1)$ is $4K_1$, where K_1 is the complete elliptic integral of modulus k_1. The course of this function is the same as that of sn (u, k) except that its period is different. If it is to correspond to $F(\Omega)$ as given by (16), it must be zero for $u=0$ and increase to $+1$ as its argument increases to K_1, and then decrease to zero at Ω_x (as yet unknown) where its argument is $2K_1$; it must trace out a similar negative cycle between Ω_x and Ω_y and then increase to $+1$ at $\Omega=\sqrt{k}$, or $u=K$. The desired behavior is obtained by setting $C_1=0$, choosing the plus sign for M, and equating M to $5K_1/K$. This replaces (17) by

$$\Omega = \sqrt{k}\,\mathrm{sn}\,(u, k)$$
$$F(u) = \mathrm{sn}\,(5uK_1/K, k_1). \tag{18}$$

These functions are plotted in Fig. 3. It is observed that

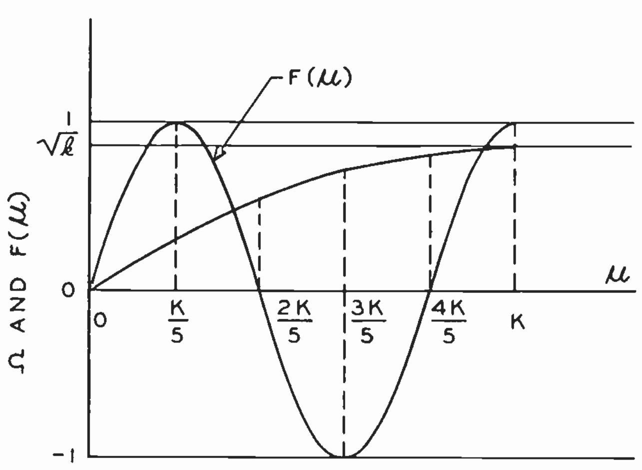


Fig. 3—Comparison between the frequency variable and the insertion loss function.

the zero loss points are located at $u=0$, $2K/5$, $4K/5$, and the unit maxima at $u=K/5$, $3K/5$, K. The frequencies of zero loss are:

$$\Omega_s = \sqrt{k}\,\mathrm{sn}\,(2sK/5, k), \qquad (s = 0, 1, 2). \tag{19}$$

Fig. 3 reveals that a plot of $F(\Omega)$ against Ω has its oscillations crowded toward the upper edge of the pass band.

At this point, a function has been found which exhibits the type of performance desired in the pass band, and the location of the zero loss points of $F(\Omega)$ appearing in (16) has been determined. There remains the specification of the infinite loss points and the constant F_0. Also, there is apparently complete independence in the choice of the selectivity parameter, k, and the discrimination parameter, k_1. These questions, except the determination of F_0, will be resolved by studying the behavior of $F(u)$ in the attenuation band.

It is shown in the appendix that the argument u in (18) is replaced by $(u+jK')$ in the attenuation band, *i.e.*, for the frequency interval, $1/\sqrt{k} \leqq \Omega \leqq \infty$. Here, u is real $(0 \leqq u \leqq K)$ and K' is the complementary complete integral, or the complete integral of modulus k'. In place of (18), we have:

$$\Omega = \frac{1}{\sqrt{k}\,\mathrm{sn}\,(u, k)},$$
$$F(u) = \mathrm{sn}\,[5(u + jK')K_1/K, k_1]. \tag{20}$$

In order that this function describe the loss characteristic shown in Fig. 2, it is required to be infinite at infinite frequency and at two as yet undetermined frequencies. Since $u=0$ corresponds to Ω equal to infinity, the requirement of an infinite loss peak at infinity means that $F(0)=\mathrm{sn}\,(5jK'K_1/K, k_1)$ should be infinite. It is seen from (49) and (48) in the Appendix that this will be the case for

$$\frac{5K'}{K} = \frac{K_1'}{K_1}$$

where K_1' is the complete elliptic integral of modulus $k_1'=\sqrt{1-k_1^2}$. Since K and K' are functions of k, the selectivity parameter, while K_1 and K_1' are similar functions of k_1, the discrimination parameter, we find that these parameters are related and they cannot all be chosen arbitrarily. This result will be given in a more useful form later.

The two finite frequencies of infinite loss may be determined by introducing the above condition in (20) to get,

$$F(u) = \mathrm{sn}\left(5u\,\frac{K_1}{K} + jK_1', k_1\right).$$

An examination of Fig. 21, with K and K' replaced by K_1 and K_1', respectively, shows that poles of $F(u)$ are located at values of the argument equal to $2sK_1+jK_1'$ ($s=$ integer). Hence, to calculate u, we set

$$5uK_1/K = 2sK_1$$

and obtain

$$u = 2sK/5.$$

Therefore, the frequencies of infinite loss are specified by

$$\Omega_s = \frac{1}{\sqrt{k}\,\mathrm{sn}\,(2sK/5, k)}, \qquad (s = 0, 1, 2). \tag{21}$$

Comparison of (21) and (19) shows that the frequencies of infinite loss in the attenuation band are the reciprocals of the frequencies of zero loss in the pass band. It is evident also from Fig. 21 that the minima of $F(u)$ in the attenuation band (which are equal in magnitude to $1/k_1$) occur at values of the argument equal to $(2s+1)K_1+jK_1'$, corresponding to $u=(2s+1)K/5$. Hence, the frequencies of minimum loss in the attenuation band are the reciprocals of the frequencies of maximum loss in the pass band.

Eq. (16) may now be written as:

$$F(\Omega) = F_0 \frac{\Omega(\Omega^2 - \Omega_1{}^2)(\Omega^2 - \Omega_2{}^2)}{(1 - \Omega_1{}^2\Omega^2)(1 - \Omega_2{}^2\Omega^2)}. \qquad (22)$$

This function has equal ripples in the pass band and equal minima in the attenuation band provided that the Ω_s are selected in accordance with (19). However, until the constant F_0 is evaluated, the values of the equal maxima and equal minima are not specified. This evaluation may be made, well enough for our purposes, by equating $F(\Omega)$ to $F(u)$ at the value of u corresponding to a particular Ω. A convenient frequency at which this computation can be performed is the reference frequency, $\Omega=1$, where $F(\Omega)=F_0$ according to (22). This frequency lies in the transition band, $\sqrt{k}\leqq\Omega\leqq 1/\sqrt{k}$. In this interval, as discussed in the appendix, the argument u in (18) is replaced by $(K+jv)$ and the frequencies are computed from

$$\Omega = \sqrt{k}/dn(v, k').$$

The ends of the transition band are given by $v=0$ and $v=K'$. Hence, one might expect that the reference frequency, $\Omega=1$, corresponds to $v=K'/2$. This is in fact the case as may be checked by the elliptic function relation which reads:

$$\mathrm{dn}^2(v/2) = \frac{\mathrm{cn}\, v + \mathrm{dn}\, v}{1 + \mathrm{cn}\, v}.$$

We are now in a position to evaluate F_0 by making the substitution, $u=K+jK'/2$ in (18). After this is done, remembering that $5K'/K=K_1'/K_1$ and the real period of this sn function is $4K_1$, we get

$$F_0 = \mathrm{sn}\,(K_1 + jK_1'/2, k_1) = \frac{1}{\mathrm{dn}\,(K_1'/2, k_1')}.$$

Upon using the formula above, this turns out to be

$$F_0 = 1/\sqrt{k_1} = [(e^{2\alpha_a} - 1)/(e^{2\alpha_p} - 1)]^{1/4}. \qquad (23)$$

This completes the explanation of how the various parameters which appear in the insertion power ratio for a two-section filter are determined. The results for the general case of an m-section filter are readily apparent. The quantity m is used to denote the number of infinite loss peaks (at finite frequencies) for low and high pass filters, and one-half the number of infinite loss peaks for band pass and band elimination filters.

The insertion power ratio for a symmetrical Tchebycheff parameter filter is:

$$e^{2\alpha} = 1 + \sqrt{(e^{2\alpha_p} - 1)(e^{2\alpha_a} - 1)} \cdot \left[\frac{\Omega(\Omega^2 - \Omega_1{}^2) \cdots (\Omega^2 - \Omega_m{}^2)}{(1 - \Omega_1{}^2\Omega^2) \cdots (1 - \Omega_m{}^2\Omega^2)}\right]^2 \qquad (24)$$

where α is the insertion loss in nepers, α_p the maximum pass band loss, α_a the minimum attenuation band loss. The frequencies at which the loss is zero, and the reciprocals of the frequencies at which the loss is infinite are specified by:

$$\Omega_s = \sqrt{k}\, \mathrm{sn}\left[\frac{2sK}{2m+1}, k\right], \quad s = 0, 1, \cdots, m. \qquad (25)$$

The selectivity parameter $k=f_p/f_a$ and the discrimination parameter $k_1=\sqrt{(e^{2\alpha_p}-1)(e^{2\alpha_a}-1)}$ are related in such a way that the condition,

$$(2m+1)\frac{K'}{K} = \frac{K_1'}{K_1} \qquad (26)$$

is satisfied.

Determination of Network Impedances

The impedances of the network are determined by following the procedure outlined previously. The first step is the calculation of the roots of the polynomial $(P+pS)$. We start with the insertion power ratio,

$$e^{2\alpha} = 1 + (e^{2\alpha_p} - 1)\, \mathrm{sn}^2\, [(2m+1)uK_1/K, k_1]$$

which is (15) after (18) has been written for the general case of an m-section filter. This expression may be regarded as the product of two factors, one of them being,

$$1 + j\sqrt{(e^{2\alpha_p} - 1)}\, \mathrm{sn}\, [(2m+1)uK_1/K, k_1]. \qquad (27)$$

The roots of this factor are the negatives of the roots of the other (or conjugate) factor. This means that the values of u which satisfy the equation:

$$\mathrm{sn}\, [(2m+1)uK_1/K, k_1] = j/\sqrt{e^{2\alpha_p} - 1} \qquad (28)$$

and their negatives are the roots of the insertion power ratio. The solutions of (28) may be regarded as the roots of $(P+pS)$ and their negatives as the roots of $(P-pS)$. By making the substitution $u=jv_0$, we obtain:

$$\frac{\mathrm{sn}\, [(2m+1)v_0K_1/K, k_1']}{\mathrm{cn}\, [(2m+1)v_0K_1/K, k_1']} = \frac{1}{\sqrt{e^{2\alpha_p} - 1}}, \qquad (29)$$

from which it is possible to evaluate v_0. Therefore, one value of the argument which satisfies (28) is $(2m+1)jv_0K_1/K$. Since the real period of this elliptic sine is $4K_1$, all the different solutions are given by:

$$(2m+1)jv_0\frac{K_1}{K}+4sK_1$$
$$=(2m+1)\frac{K_1}{K}\left[jv_0+\frac{4sK}{2m+1}\right],$$

where $s=0, 1, 2, \cdots, 2m$. Thus, we have the result that the roots of $(P+pS)$ in terms of u are:

$$u=jv_0+\frac{4sK}{2m+1}\cdot \quad (s=0, 1, \cdots, 2m). \tag{30}$$

Network impedances, however, are specified in terms of the frequency parameter $p=j\Omega$. Hence, the roots are required in terms of p. They are found by means of the transformation,

$$p=j\sqrt{k}\,\mathrm{sn}\,(u, k).$$

Corresponding to jv_0, there is the real root,

$$a_0=j\sqrt{k}\,\mathrm{sn}\,(jv_0, k)=-\sqrt{k}\frac{\mathrm{sn}\,(v_0, k')}{\mathrm{cn}\,(v_0, k')}\cdot \tag{31}$$

The remaining roots are given by:

$$a_s+jb_s=(-1)^s j\sqrt{k}\,\mathrm{sn}\left[jv_0\pm\frac{2sK}{2m+1}, k\right],$$
$$(s=1, \cdots, m) \tag{32}$$

which recognizes the fact that values of the elliptic sine differ only in sign for values of the argument which differ by one-half the real period and that the roots come in conjugate complex pairs. With the aid of the summation formula [see (53)] and the standard relations between the elliptic functions, this may be written as:

$$a_s+jb_s=\frac{(-1)^s a_0V_s\pm j\Omega_sW}{1+a_0^2\Omega_s^2}, \quad (s=1, \cdots, m) \tag{33}$$

where,

$$W=\sqrt{(1+ka_0^2)(1+a_0^2/k)}$$
$$V_s=\sqrt{(1-k\Omega_s^2)(1-\Omega_s^2/k)}$$
$$\Omega_s=\sqrt{k}\,\mathrm{sn}\left(\frac{2sK}{2m+1}, k\right).$$

It is convenient in the determination of the network impedances to regard a_0 as a positive number. This causes no difficulty since the effect is merely to replace the network by its inverse. With this convention, the set of roots that are obtained from (33) for odd values of the index s are assigned to the polynomial (P_1+pS_1), while a_0 and those roots corresponding to even values of s are reversed in sign and assigned to (P_2+pS_2). How this works out for a particular case is illustrated readily by a two-section filter. We have, except for possible constant multipliers,

$$P_1+pS_1=(p+a_1+jb_1)(p+a_1-jb_1)$$
$$P_2+pS_2=(p+a_0)(p+a_2+jb_2)(p+a_2-jb_2).$$

The branch impedances of the lattice network are given at once by:

$$\frac{Z_1}{R}=\frac{p^3+(a_2^2+b_2^2+2a_0a_2)p}{(a_0+2a_2)p^2+a_0(a_2^2+b_2^2)},$$
$$\frac{Z_2}{R}=\frac{p^2+a_1^2+b_1^2}{2a_1p}.$$

Approximate Formulas

All the formulas required in the design of a Tchebycheff parameter filter have been presented. In principle, a particular design can be undertaken with the aid of elliptic function tables. However, these tables must be double-entry tables since the value of an elliptic function depends not only on the argument but also on the modulus. Interpolation between tabulated values is laborious and subject to error. Fortunately, the functions that appear in the numerical computations can be expressed in terms of certain auxiliary functions called theta functions. These are represented by rapidly converging series. It will appear that the need for elliptic function tables can be removed completely.

The notation that is used in the literature on theta functions is quite varied. We choose the following definitions:

$$\vartheta_0(u/2K, q)=1+2\sum_{n=1}^{\infty}(-1)^n q^{n^2}\cos 2n\frac{\pi u}{2K}$$

$$\vartheta_1(u/2K, q)=2q^{1/4}\sum_{n=0}^{\infty}(-1)^n q^{n(n+1)}\sin(2n+1)\frac{\pi u}{2K}$$

$$\vartheta_2(u/2K, q)=2q^{1/4}\sum_{n=0}^{\infty}q^{n(n+1)}\cos(2n+1)\frac{\pi u}{2K}$$

$$\vartheta_3(u/2K, q)=1+2\sum_{n=1}^{\infty}q^{n^2}\cos 2n\frac{\pi u}{2K}$$

where

$q=e^{-\pi K'/K}=$ modular constant

$K=$ complete elliptic integral of modulus k

$K'=$ complete elliptic integral of modulus $\sqrt{1-k^2}$.

The elliptic functions are expressed in terms of the theta functions as follows:

$$\mathrm{sn}\,(u, k)=\frac{1}{\sqrt{k}}\cdot\frac{\vartheta_1(u/2K, q)}{\vartheta_0(u/2K, q)}$$

$$\mathrm{cn}\,(u, k)=\sqrt{\frac{k'}{k}}\cdot\frac{\vartheta_2(u/2K, q)}{\vartheta_0(u/2K, q)}$$

$$\mathrm{dn}\,(u, k)=\sqrt{k'}\cdot\frac{\vartheta_3(u/2K, q)}{\vartheta_0(u/2K, q)}\cdot \tag{34}$$

In the usual case, the filter design problem consists of the specification of the number of filter sections that is required to attain prescribed minimum attenuation band loss, a prescribed waste or transition band, and a tolerable pass band ripple. Hence, it is desirable to have a simple expression for the relation between these various quantities. This may be derived from the fundamental condition that was demonstrated earlier, namely,

$$(2m+1)\frac{K'}{K} = \frac{K_1'}{K_1}.$$

By defining $q_1 = e^{-\pi K'_1/K_1}$ as the modular constant for functions of modulus k_1, we get the following alternative expression for the above condition,

$$q_1 = q^{(2m+1)}. \tag{35}$$

The objective now is to derive the desired relation through use of the modular constants q and q_1. The first step is to determine q in terms of k. One of the many approximations that can be used is based on the observation that:

$$\operatorname{dn}(0, k) = \sqrt{k'}\frac{\vartheta_3(0, q)}{\vartheta_0(0, q)} = 1.$$

However, a better approximation is obtained by forming,

$$2\epsilon = \frac{1-\sqrt{k'}}{1+\sqrt{k'}}.$$

Upon introducing the theta-function series, dividing the numerator by the denominator, and reverting the resulting series,

$$q = \epsilon + 2\epsilon^5 + 15\epsilon^9 + 150\epsilon^{13} + \cdots \tag{36}$$

The inverse approximation is needed also. This is found by noting that,

$$\operatorname{sn}(K, k) = \frac{1}{\sqrt{k}}\cdot\frac{\vartheta_1(1/2, q)}{\vartheta_0(1/2, q)} = 1.$$

Hence,

$$k = 4\sqrt{q}\left[\frac{1+q^2+q^6+\cdots}{1+2q+2q^4+\cdots}\right]^2. \tag{37}$$

It is recognized that these approximations are equally valid for the quantities denoted by the subscript "1."

The relation which we are seeking is based on the definition of the discrimination parameter,

$$k_1^2 = \frac{e^{2\alpha_p}-1}{e^{2\alpha_a}-1}.$$

In any ordinary filter design, α_a is large, α_p is small, so that k_1 is very small and the approximation $k_1^2 = 16q_1$ is sufficiently accurate. This leads to:

$$e^{2\alpha_a} = 1 + \frac{e^{2\alpha_p}-1}{16q_1}.$$

Now, if unity is neglected in comparison with the other quantity, and (35) is used, we obtain the result,

$$\alpha_a \doteq 10[\log(e^{2\alpha_p}-1) + (2m+1)\log 1/q] - 12\text{ db} \tag{38}$$

where the logarithms are to the base 10 and α_a is in decibels but α_p is in nepers.[3] This important equation is plotted in Fig. 9 for $\alpha_p = 0.25$ db. This is in a sense a universal chart since α_a is changed by a fixed amount as α_p is changed. The change is roughly proportional on a logarithmic basis. Table I illustrates this point.

TABLE I

α_p	change α_a by
0.1 db	−4.0 db
0.2	−1.0
0.3	+0.8
0.4	+2.1
0.5	+3.1

The interdependence of the parameters is evident. For example, the discrimination decreases as the selectivity increases for a given number of filter sections and a fixed value of the (nondissipative) pass band ripple; also, for a specified selectivity and number of sections, the discrimination is reduced much more rapidly than the pass band ripple. It should be emphasized that as soon as three of the quantities, α_a, α_p, m, k (or q) are chosen, the fourth is determined, and the insertion loss characteristic of the filter is completely specified.

The frequencies at which the (nondissipative) pass band loss is zero, and the reciprocals of the frequencies at which the attenuation band loss is infinite are:

$$\Omega_s = 2q^{1/4}\left[\frac{\sin\frac{s\pi}{2m+1} - q^2\sin\frac{3s\pi}{2m+1} + q^6\sin\frac{5s\pi}{2m+1} + \cdots}{1 - 2q\cos\frac{2s\pi}{2m+1} + 2q^4\cos\frac{4s\pi}{2m+1} + \cdots}\right] \tag{39}$$

where $s = 0, 1, \cdots, m$. This is (25) written in terms of the theta-function series. Similarly, the frequencies at which the pass band loss is equal to the maximum, α_p, and the reciprocals of the frequencies at which the attenuation band loss is equal to the minimum α_a are:

[3] Usually α_p would be specified in decibels. This may be used directly in (38) provided $e^{2\alpha_p}$ is replaced by $10^{\alpha_p/10}$.

$$\Omega_t = 2q^{1/4}\left[\frac{\sin\frac{(2t+1)\pi}{2(2m+1)} - q^2\sin\frac{3(2t+1)\pi}{2(2m+1)} + q^6\sin\frac{5(2t+1)\pi}{2(2m+1)} + \cdots}{1 - 2q\cos\frac{(2t+1)\pi}{(2m+1)} + 2q^4\cos\frac{2(2t+1)\pi}{(2m+1)} + \cdots}\right] \tag{40}$$

where $t=0, 1, \cdots, m$.

The determination of the network impedances is expedited through the use of approximate formulas. The quantity that is to be calculated is the value of the real root a_0 of the insertion function. In practical cases the modulus k_1 in (28) is so small that it may be considered equal to zero without introducing appreciable error. Then in place of (29) we have:

$$\sinh\,(2m+1)\,\frac{v_0\pi}{2K} = \frac{1}{\sqrt{e^{2\alpha_p}-1}}\,.$$

By expressing the square of the hyperbolic sine in terms of exponential functions and arbitrarily assigning a positive value to v_0, we get:

$$\frac{v_0\pi}{2K} = \frac{1}{2(2m+1)}\log_e\coth\frac{\alpha_p}{2}\,.$$

The application of the standard series expansions yields the result,

$$\Lambda = \frac{v_0\pi}{2K} = \frac{1}{2(2m+1)}\left[\log_e\frac{(2)}{(\alpha_p)} + \frac{\alpha_p{}^2}{12} + \cdots\right] \tag{41}$$

where α_p is in nepers. The real root a_0 as specified by (31), may be written as,

$$a_0 = j\,\frac{\vartheta_1(jv_0/2K, q)}{\vartheta_0(jv_0/2K, q)}\,.$$

Upon introducing the theta-function series, and arbitrarily assigning a positive sign to the result, we obtain:

$$a_0 = 2q^{1/4}\left[\frac{\sinh\Lambda - q^2\sinh 3\Lambda + q^6\sinh 5\Lambda + \cdots}{1 - 2q\cosh 2\Lambda + 2q^4\cosh 4\Lambda + \cdots}\right]. \tag{42}$$

Many other exact and approximate formulas can be derived. However, those which have been described are sufficient for the design of Tchebycheff parameter filters.

Design Charts

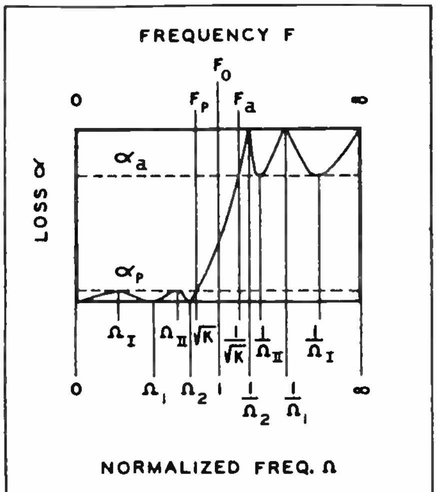
The design formulas for the various types of filters can be expressed most conveniently in terms of a normalized low-pass filter. The relations that are needed for transforming low-pass, high-pass, band-pass, and band-elimination characteristics to the normalized low-pass characteristic are given in Fig. 4 (next page). It is important to notice that the insertion loss characteristics of the band-pass and band-elimination filters are symmetrical about the midband frequency, f_m, when plotted against a logarithmic frequency scale.

The determination of the roots of the insertion power ratio for one to four section normalized low-pass filters is given as a step-by-step procedure[4] in Figs. 5–8 (pp. 465–466). This procedure assumes that the number, m, of filter sections and the selectivity parameter, k, have been selected. This selection can be made with the aid of Fig. 9 (p. 466), which depicts the relation between these quantities and the minimum attenuation band loss for a specified pass band ripple. In a typical design problem, the pass band is specified within more or less well defined limits. Allowance should be made for the fact that dissipation in the components will narrow the effective pass band. The frequency band which is to be attenuated a prescribed amount is also more or less well defined. These specifications are sufficient for choosing a trial value for the parameter k with which to enter the chart, Fig. 9. It is clear that, in general, a compromise must be made among the four available parameters.

The design process continues by next computing the component values for the normalized low-pass filter. Although the lattice configuration has been used for purposes of analysis whenever attention has been directed at a specific structure, it is not a desirable configuration to use in practice. In the first place, it can be inserted only between lines that are balanced to ground or transformers must be included to convert unbalanced circuits to balanced; in the second place, stop band losses are attained by bridge balance which is difficult to maintain in regular production and throughout the life of the filter. Consequently, the remaining design formulas are directed at ladder configurations. They are derived from the corresponding lattice filters which have the branch impedances, Z_1 and Z_2. As is well known, the open circuit impedance is $(Z_1+Z_2)/2$. The ladder network must, of course, exhibit this same open circuit impedance. Now, Table I of Darlington's paper contains a straightforward procedure for calculating the component values of a ladder network from the open circuit (or short circuit) impedance. The procedure is particularly simple in the case of converting a lattice to a ladder. Because of electrical symmetry, calculations can be made from each end of the network, thereby reducing the complexity of the equations. Further simplification is possible since the fact that $Z_1=Z_2$ at the frequencies of infinite loss provides relations between the various constants. Nevertheless, there is a considerable amount of algebraic manipulation required to reduce the results to a compact form. Figs. 10–13 (pp. 466–467) contain the ladder network co-

[4] The check on the computations (step 11) is an approximate relation. For relatively large values of k_1, that is, relatively small discrimination (say, less than 20 db), the two sides of the equation may differ from each other by as much as 1/2 per cent.

LOW PASS

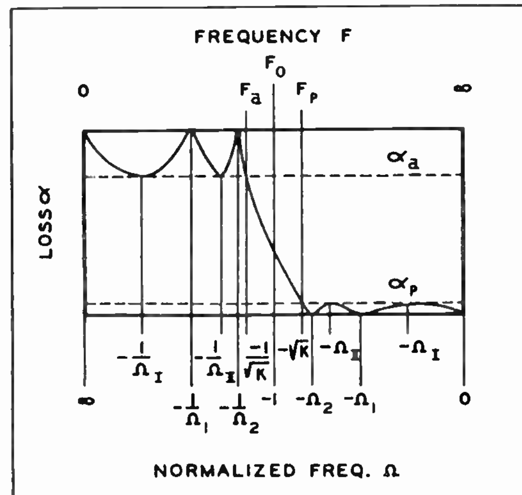


$$F_0=\sqrt{F_p F_a}$$

$$\Omega = F/F_0$$

$$K = F_p/F_a$$

HIGH PASS

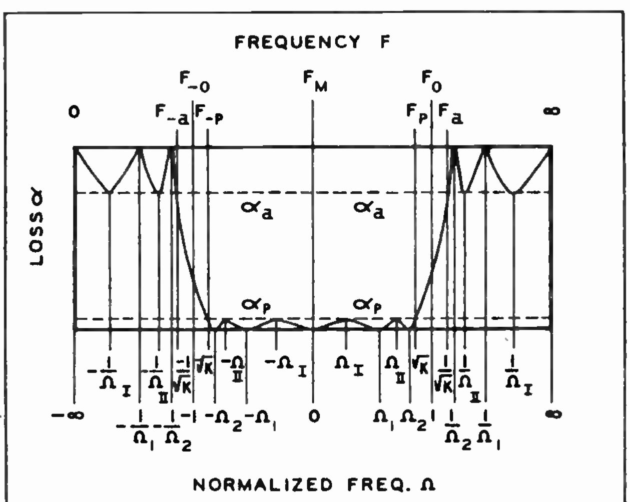


$$F_0=\sqrt{F_a F_p}$$

$$\Omega = -F_0/F$$

$$K = F_a/F_p$$

BAND PASS



$$F_M=\sqrt{F_p F_{-p}} = \sqrt{F_a F_{-a}}$$

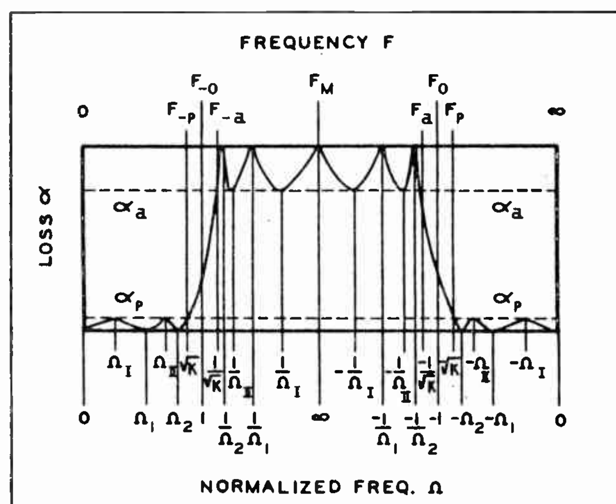
$$\Delta = \frac{1}{F_M}\sqrt{\left(F_p - F_{-p}\right)\left(F_a - F_{-a}\right)}$$

$$\Omega = \frac{F/F_M - F_M/F}{\Delta}$$

$$F = F_M\left(\frac{\Delta\Omega}{2} + \sqrt{\left(\frac{\Delta\Omega}{2}\right)^2 + 1}\right)$$

$$K = \frac{F_p - F_{-p}}{F_a - F_{-a}}$$

BAND ELIMINATION



$$F_M=\sqrt{F_a F_{-a}} = \sqrt{F_p F_{-p}}$$

$$\Delta = \frac{1}{F_M}\sqrt{\left(F_a - F_{-a}\right)\left(F_p - F_{-p}\right)}$$

$$\Omega = \frac{\Delta}{F_M/F - F/F_M}$$

$$F = F_M\left(\frac{-\Delta}{2\Omega} + \sqrt{\left(\frac{\Delta}{2\Omega}\right)^2 + 1}\right)$$

$$K = \frac{F_a - F_{-a}}{F_p - F_{-p}}$$

Fig. 4—Definition of design parameters.

1) $k' = \sqrt{1 - k^2}$

2) $\epsilon = \dfrac{1}{2}\left[\dfrac{1 - \sqrt{k'}}{1 + \sqrt{k'}}\right]$

3) $q = \epsilon + 2\epsilon^5 + 15\epsilon^9$

4) $\alpha_a = 10 \log (10^{\alpha_p/10} - 1) + 30 \log \dfrac{1}{q} - 12$

5) $\Lambda = 0.475808 + 0.383764 \log \dfrac{1}{\alpha_p} + 0.000184\, \alpha_p^2$

6) $a_0 = 2(q)^{1/4}\left[\dfrac{\sinh \Lambda - q^2 \sinh 3\Lambda + q^6 \sinh 5\Lambda}{1 - 2q \cosh 2\Lambda + 2q^4 \cosh 4\Lambda}\right]$

7) $\Omega_1 = \sqrt{3}(q)^{1/4}\left[\dfrac{1 - q^6}{1 + q - q^4}\right]$

8) $a_1 = \dfrac{a_0}{1 + a_0^2\Omega_1^2}\sqrt{(1 - k\Omega_1^2)(1 - \Omega_1^2/k)}$

9) $b_1 = \dfrac{\Omega_1}{1 + a_0^2\Omega_1^2}\sqrt{(1 + ka_0^2)(1 + a_0^2/k)}$

10) $p_1^2 = a_1^2 + b_1^2$

11) $a_0 p_1^2 = \dfrac{2(q)^{3/4}}{\sqrt{10^{\alpha_p/10} - 1}}$ (for check)

12) $\Omega_I = (q)^{1/4}\left[\dfrac{1 - 2q^2 + q^6}{1 - q - q^4}\right]$

Fig. 5—Design formulas for one-section filter.

1) $k' = \sqrt{1 - k^2}$

2) $\epsilon = \dfrac{1}{2}\left[\dfrac{1 - \sqrt{k'}}{1 + \sqrt{k'}}\right]$

3) $q = \epsilon + 2\epsilon^5 + 15\epsilon^9$

4) $\alpha_a = 10 \log (10^{\alpha_p/10} - 1) + 50 \log \dfrac{1}{q} - 12$

5) $\Lambda = 0.285485 + 0.230258 \log \dfrac{1}{\alpha_p} + 0.000110\, \alpha_p$

6) $a_0 = 2(q)^{1/4}\left[\dfrac{\sinh \Lambda - q^2 \sinh 3\Lambda + q^6 \sinh 5\Lambda}{1 - 2q \cosh 2\Lambda + 2q^4 \cosh 4\Lambda}\right]$

7) $\Omega_1 = 2(q)^{1/4}\left[\dfrac{\sin \pi/5 - q^2 \sin 2\pi/5}{1 - 2q \cos 2\pi/5 - 2q^4 \cos \pi/5}\right]$

$\Omega_2 = 2(q)^{1/4}\left[\dfrac{\sin 2\pi/5 + q^2 \sin \pi/5}{1 + 2q \cos \pi/5 + 2q^4 \cos 2\pi/5}\right]$

8) $a_s = \dfrac{a_0}{1 + a_0^2\Omega_s^2}\sqrt{(1 - k\Omega_s^2)(1 - \Omega_s^2/k)}, \quad (s = 1, 2)$

9) $b_s = \dfrac{\Omega_s}{1 + a_0^2\Omega_s^2}\sqrt{(1 + ka_0^2)(1 + a_0^2/k)}$

10) $p_s^2 = a_s^2 + b_s^2$

11) $a_0 p_1^2 p_2^2 = \dfrac{2(q)^{5/4}}{\sqrt{10^{\alpha_p/10} - 1}}$ (for check)

12) $\Omega_I = 2(q)^{1/4}\left[\dfrac{\sin \pi/10 - q^2 \sin 3\pi/10 + q^6}{1 - 2q \cos \pi/5 + 2q^4 \cos 2\pi/5}\right]$

$\Omega_{II} = 2(q)^{1/4}\left[\dfrac{\sin 3\pi/10 - q^2 \sin \pi/10 - q^6}{1 + 2q \cos 2\pi/5 - 2q^4 \cos \pi/5}\right]$

Fig. 6—Design formulas for two-section filter.

1) $k' = \sqrt{1 - k^2}$

2) $\epsilon = \dfrac{1}{2}\left[\dfrac{1 - \sqrt{k'}}{1 + \sqrt{k'}}\right]$

3) $q = \epsilon + 2\epsilon^5 + 15\epsilon^9$

4) $\alpha_a = 10 \log (10^{\alpha_p/10} - 1) \div 70 \log \dfrac{1}{q} - 12$

5) $\Lambda = 0.203918 + 0.164470 \log \dfrac{1}{\alpha_p} + 0.000079\, \alpha_p^2$

6) $a_0 = 2(q)^{1/4}\left[\dfrac{\sinh \Lambda - q^2 \sinh 3\Lambda + q^6 \sinh 5\Lambda}{1 - 2q \cosh 2\Lambda + 2q^4 \cosh 4\Lambda}\right]$

7) $\Omega_1 = 2(q)^{1/4}\left[\dfrac{\sin \pi/7 - q^2 \sin 3\pi/7 + q^6 \sin 2\pi/7}{1 - 2q \cos 2\pi/7 - 2q^4 \cos 3\pi/7}\right]$

$\Omega_2 = 2(q)^{1/4}\left[\dfrac{\sin 2\pi/7 - q^2 \sin \pi/7 - q^6 \sin 3\pi/7}{1 + 2q \cos 3\pi/7 - 2q^4 \cos \pi/7}\right]$

$\Omega_3 = 2(q)^{1/4}\left[\dfrac{\sin 3\pi/7 + q^2 \sin 2\pi/7 + q^6 \sin \pi/7}{1 + 2q \cos \pi/7 + 2q^4 \cos 2\pi/7}\right]$

8) $a_s = \dfrac{a_0}{1 + a_0^2\Omega_s^2}\sqrt{(1 - k\Omega_s^2)(1 - \Omega_s^2/k)}, \quad (s = 1, 2, 3)$

9) $b_s = \dfrac{\Omega_s}{1 + a_0^2\Omega_s^2}\sqrt{(1 + ka_0^2)(1 + a_0^2/k)}$

10) $p_s^2 = a_s^2 + b_s^2$

11) $a_0 p_1^2 p_2^2 p_3^2 = \dfrac{2(q)^{7/4}}{\sqrt{10^{\alpha_p/10} - 1}}$ (for check)

12) $\Omega_I = 2(q)^{1/4}\left[\dfrac{\sin \pi/14 - q^2 \sin 3\pi/14 + q^6 \sin 5\pi/14}{1 - 2q \cos \pi/7 + 2q^4 \cos 2\pi/7}\right]$

$\Omega_{II} = 2(q)^{1/4}\left[\dfrac{\sin 3\pi/14 - q^2 \sin 5\pi/14 - q^6 \sin \pi/14}{1 - 2q \cos 3\pi/7 - 2q^4 \cos \pi/7}\right]$

$\Omega_{III} = 2(q)^{1/4}\left[\dfrac{\sin 5\pi/14 + q^2 \sin \pi/14 - q^6 \sin 3\pi/14}{1 + 2q \cos 2\pi/7 - 2q^4 \cos 3\pi/7}\right]$

Fig. 7—Design formulas for three-section filter.

efficients (component values for the normalized low-pass filter) for one to four section filters.[5] The first three figures are merely duplications of Table II included in Darlington's paper.

It must be recognized that some of the network coefficients may turn out to be negative numbers. This happens rarely in a practical design but may occur if the selectivity and discrimination are high for the selected number of sections. Unfortunately, this unhappy situation will not be known until the design is fairly well along. The only condition that seems to be easy to apply to the design procedure outlined here is the first of Darlington's sufficient conditions. This can be checked as soon as the β_j are known. These numbers are the squares of the reciprocals of the finite poles of the open circuit driving point impedance. Then the sufficient condition requires that $\Omega_s < \beta_j$ for all s and j. If this condition is satisfied but further computation reveals a negative network coefficient, it means that Darlington's second condition has been violated. In many cases, this difficulty can be overcome by reallocating the shunt

[5] The quantity g_4 in step 1, Fig. 12 is superfluous and is not required in the computations.

1) $k' = \sqrt{1-k^2}$

2) $\epsilon = \frac{1}{2}\left[\frac{1-\sqrt{k'}}{1+\sqrt{k'}}\right]$

3) $q = \epsilon + 2\epsilon^5 + 15\epsilon^9$

4) $\alpha_a = 10\log(10^{\alpha_p/10} - 1) + 90\log\frac{1}{q} - 12$

5) $\Lambda = 0.158603 + 0.127921\log\frac{1}{\alpha_p} + 0.000061\,\alpha_p$

6) $a_0 = 2(q)^{1/4}\left[\frac{\sinh\Lambda - q^2\sinh 3\Lambda + q^6\sinh 5\Lambda}{1 - 2q\cosh 2\Lambda + 2q^4\cosh 4\Lambda}\right]$

7) $\Omega_1 = 2(q)^{1/4}\left[\frac{\sin\pi/9 - q^2\sin 3\pi/9 + q^6\sin 4\pi/9}{1 - 2q\cos 2\pi/9 + 2q^4\cos 4\pi/9}\right]$

$\Omega_2 = 2(q)^{1/4}\left[\frac{\sin 2\pi/9 - q^2\sin 3\pi/9 - q^6\sin\pi/9}{1 - 2q\cos 4\pi/9 - 2q^4\cos\pi/9}\right]$

$\Omega_3 = \sqrt{3}(q)^{1/4}\left[\frac{1-q^6}{1+q-q^4}\right]$

$\Omega_4 = 2(q)^{1/4}\left[\frac{\sin 4\pi/9 + q^2\sin 3\pi/9 + q^6\sin 2\pi/9}{1 + 2q\cos\pi/9 + 2q^4\cos 2\pi/9}\right]$

8) $a_s = \frac{a_0}{1 + a_0^2\Omega_s^2}\sqrt{(1-k\Omega_s^2)(1-\Omega_s^2/k)}, \quad (s = 1, 2, 3, 4)$

9) $b_s = \frac{\Omega_s}{1 + a_0^2\Omega_s^2}\sqrt{(1+ka_0^2)(1+a_0^2/k)}$

10) $p_s^2 = a_s^2 + b_s^2$

11) $a_0 p_1^2 p_2^2 p_3^2 p_4^2 = \frac{2(q)^{9/4}}{\sqrt{10^{\alpha_p/10} - 1}}$ (for check)

12) $\Omega_I = 2(q)^{1/4}\left[\frac{\sin\pi/18 - q^2\sin 3\pi/18 + q^6\sin 5\pi/18}{1 - 2q\cos\pi/9 + 2q^4\cos 2\pi/9}\right]$

$\Omega_{II} = (q)^{1/4}\left[\frac{1 - 2q^2 + q^6}{1 - q - q^4}\right]$

$\Omega_{III} = 2(q)^{1/4}\left[\frac{\sin 5\pi/18 - q^2\sin 3\pi/18 - q^6\sin 7\pi/18}{1 + 2q\cos 4\pi/9 - 2q^4\cos\pi/9}\right]$

$\Omega_{IV} = 2(q)^{1/4}\left[\frac{\sin 7\pi/18 + q^2\sin 3\pi/18 - q^6\sin\pi/18}{1 + 2q\cos 2\pi/9 + 2q^4\cos 4\pi/9}\right]$

Fig. 8—Design formulas for four-section filter.

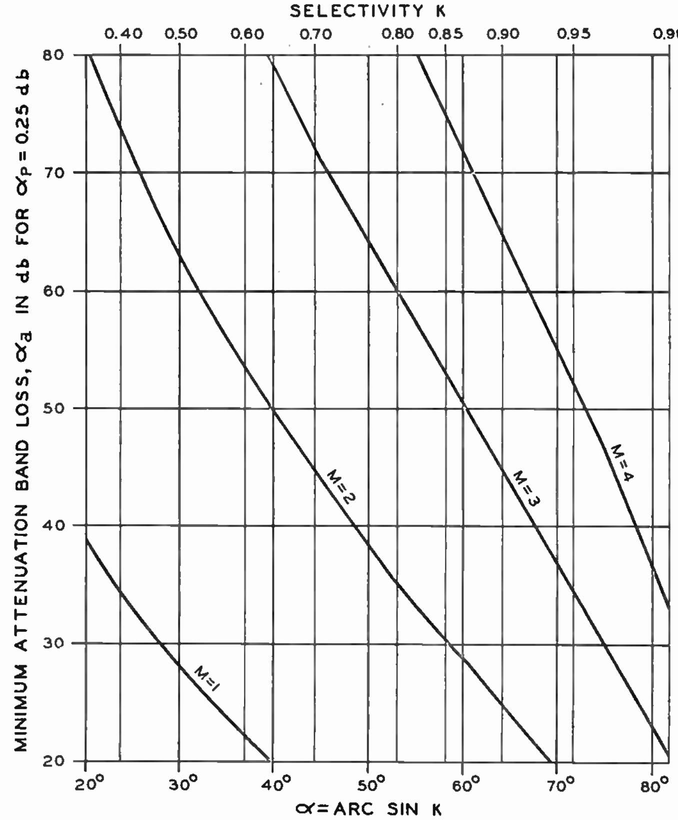
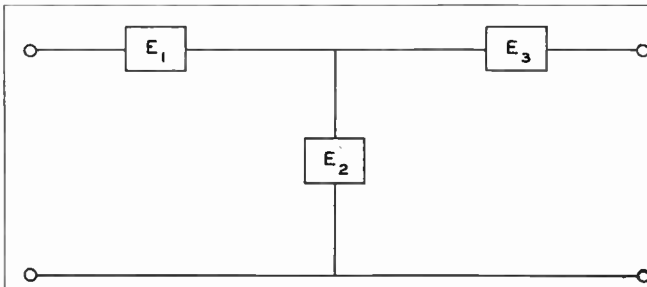


Fig. 9—Design chart for estimating loss of one-to-four section filters.



$E_1 = 1/a_0$

$E_2 = 4a_1/p_1^2$

$E_3 = 1/a_0$

Fig. 10—Network coefficients for one-section ladder.

tuned circuits so that those at each end of the filter contribute the loss peaks located at the highest frequencies. This should meet the sufficient condition that these resonances be equal to or greater than the roots of the open circuit impedance. The mechanics for this are as follows: in Fig. 12, interchange Ω_2 and Ω_3; in Fig. 13, interchange Ω_2 and Ω_4.

The final design chart is Fig. 14 which gives the component values for each type of filter in terms of the network coefficients obtained from the previous figures. It is understood that the four element branches in the band-pass and band-elimination filters can be replaced by any of the three circuits equivalent to them.

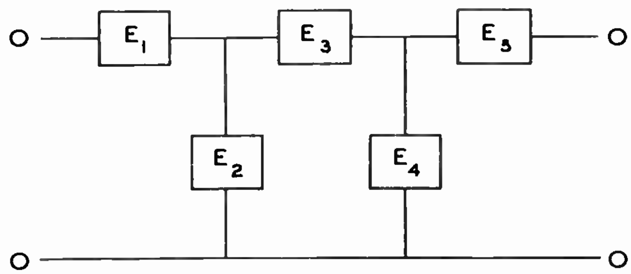
Acknowledgment

The design charts included in this paper were prepared many years ago, in fact, prior to the general publication of Darlington's thesis. The text is based on notes prepared by Dr. Darlington. As indicated in the introduction, it is intended that this paper provide a complete exposition of one aspect of the insertion loss design technique. Consequently, there may be overlap with previous publications. Since no search has been made of the literature, I hereby make my sincere apologies to those authors whose work may be duplicated here. I am indebted to Dr. Darlington not only for the material which he has provided but more importantly for the patience and skill he has shown in giving me some understanding of his classic work.

Appendix

It is possible to deduce various properties of the trigonometric sine, $z = \sin w$, from the properties of the integral,

$$w = \int_0^z \frac{dz}{\sqrt{1-z^2}}.$$



1)	$g_1 = \frac{1}{a_0 p_2^2}$ $g_2 = g_1(p_2^2 + 2a_0 a_2)$ $g_3 = \frac{1}{2a_1}$ $g_4 = g_3 p_1^2$	5)	$T_1 = d_{11} + d_{12}$
2)	$\beta_1 = g_1(a_0 + 2a_2)$	6)	$E_1 = g_3 - g_4\Omega_1^2$ $E_2 = \frac{d_{11}s_0}{T_1}$ $E_3 = \frac{2T_1}{s_0}$ $E_4 = \frac{d_{12}s_0}{T_1}$ $E_5 = g_3 - g_4\Omega_2^2$
3)	$d_{11} = \beta_1 - \Omega_1^2$ $d_{12} = \beta_1 - \Omega_2^2$		
4)	$s_0 = \frac{2}{g_4}$	7) (check)	$E_1 + E_3 + E_5 = 2g_2$ $E_2 + E_4 = s_0$

Fig. 11—Network coefficients for two-section ladder.

1)	$g_1 = \frac{1}{a_0 p_2^2}$ $g_2 = g_1(p_2^2 + 2a_0 a_2)$ $g_3 = \frac{1}{2(a_1 p_3^2 + a_3 p_1^2)}$ $g_4 = g_3(p_1^2 + p_3^2 + 4a_1 a_3)$ $g_5 = g_3 p_1^2 p_3^2$	5)	$T_1 = d_{11}d_{23} + d_{13}d_{21}$ $T_2 = 2d_{11}d_{13}d_{21} + (\Omega_1^2 - \Omega_2^2)T_1$ $T_3 = 2d_{11}d_{13}d_{23} + (\Omega_3^2 - \Omega_2^2)T_1$
2)	$\beta_1 = g_1(a_0 + 2a_2)$ $\beta_2 = 2g_3(a_1 + a_3)$	6)	$E_1 = \frac{g_1 - g_2\Omega_1^2}{d_{11}}$ $E_2 = \frac{d_{11}d_{21}s_1}{T_2}$ $E_3 = \frac{2d_{13}T_2}{s_1 T_1}$ $E_4 = \frac{d_{12}d_{22}s_0 T_1^2}{T_2 T_3}$ $E_5 = \frac{2d_{11}T_3}{s_1 T_1}$ $E_6 = \frac{d_{13}d_{23}s_1}{T_3}$ $E_7 = \frac{g_1 - g_2\Omega_3^2}{d_{13}}$
3)	$d_{11} = \beta_1 - \Omega_1^2$; $d_{21} = \beta_2 - \Omega_1^2$ $d_{12} = \beta_1 - \Omega_2^2$; $d_{22} = \beta_2 - \Omega_2^2$ $d_{13} = \beta_1 - \Omega_3^2$; $d_{23} = \beta_2 - \Omega_3^2$		
4)	$s_0 = \frac{2}{q_5}$ $s_1 = s_0(\beta_1 - \beta_2)$	7) (check)	$E_1 + E_3 + E_5 + E_7 = 2g_2$ $E_2 + E_4 + E_6 = s_0$

Fig. 12—Network coefficients for three-section ladder.

1)	$g_1 = \frac{1}{2(a_1 p_3^2 + a_3 p_1^2)}$ $g_2 = g_1(p_1^2 + p_3^2 + 4a_1 a_3)$ $g_3 = g_1 p_1^2 p_3^2$ $g_4 = \frac{1}{2a_0 p_2^2 p_4^2}$ $g_5 = g_4[a_0(p_2^2 + p_4^2 + 4a_2 a_4) + 2(a_2 p_4^2 + a_4 p_2^2)]$ $g_6 = 2g_4(a_0 + 2a_2 + a2_4)$		$T_5 = d_{13}T_3 + (\Omega_4^2 - \Omega_1^2)d_{11}T_1$ $T_6 = d_{12}T_3 + (\Omega_1^2 - \Omega_4^2)d_{14}T_2$ $T_7 = 2d_{11}d_{21}d_{31}T_1 + (\Omega_1^2 - \Omega_2^2)T_5$ $T_8 = 2d_{14}d_{24}d_{34}T_2 + (\Omega_4^2 - \Omega_3^2)T_6$ $T_9 = T_3 T_4 + (\Omega_4^2 - \Omega_1^2)(\Omega_3^2 - \Omega_2^2)T_1 T_2$
2)	$\beta_1 = 2g_1(a_1 + a_3)$ $\beta_2 = g_5 + \sqrt{g_5^2 - g_6}$ $\beta_3 = g_5 - \sqrt{g_5^2 - g_6}$	6)	$E_1 = \frac{g_1 - g_2\Omega_1^2 + g_3\Omega_1^4}{d_{11}}$ $E_2 = \frac{d_{11}d_{21}d_{31}s_2}{T_7}$ $E_3 = \frac{2T_1 T_7}{s_2 T_5}$ $E_4 = \frac{d_{12}d_{22}d_{32}s_0 T_5^2}{T_7 T_9}$ $E_5 = \frac{2T_3 T_9}{s_0 T_5 T_6}$ $E_6 = \frac{d_{13}d_{23}d_{33}s_0 T_6^2}{T_8 T_9}$ $E_7 = \frac{2T_2 T_8}{s_2 T_6}$ $E_8 = \frac{d_{14}d_{24}d_{34}s_2}{T_8}$ $E_9 = \frac{g_1 - g_2\Omega_4^2 + g_3\Omega_4^4}{d_{14}}$
3)	$d_{11} = \beta_1 - \Omega_1^2$; $d_{21} = \beta_2 - \Omega_1^2$; $d_{31} = \beta_3 - \Omega_1^2$ $d_{12} = \beta_1 - \Omega_2^2$; $d_{22} = \beta_2 - \Omega_2^2$; $d_{32} = \beta_3 - \Omega_2^2$ $d_{13} = \beta_1 - \Omega_3^2$; $d_{23} = \beta_2 - \Omega_3^2$; $d_{33} = \beta_3 - \Omega_3^2$ $d_{14} = \beta_1 - \Omega_4^2$; $d_{24} = \beta_2 - \Omega_4^2$; $d_{34} = \beta_3 - \Omega_4^2$		
4)	$s_0 = \frac{2}{g_3}$ $s_1 = (\beta_2 - \beta_1)(\beta_1 - \beta_3)$ $s_2 = s_0 s_1$		
5)	$T_1 = d_{13}d_{14} + s_1$ $T_2 = d_{11}d_{12} + s_1$ $T_3 = d_{11}d_{24}d_{34} + d_{14}d_{21}d_{31}$ $T_4 = d_{12}d_{23}d_{33} + d_{13}d_{22}d_{32}$	7) (check)	$E_1 + E_3 + E_5 + E_7 + E_9 = \frac{2}{a_0} + 4\left(\frac{a_2}{p_2^2} + \frac{a_4}{p_4^2}\right)$ $E_2 + E_4 + E_6 + E_8 = s$

Fig. 13—Network coefficients for four-section ladder.

LOW PASS

$$L_{2s+1} = \frac{R}{\omega_0} \cdot E_{2s+1} \qquad (s = 0, \cdots, M) \qquad C_{2s+1} = \frac{1}{\omega_0 R} \cdot E_{2s+1}$$

$$C_{2s} = \frac{1}{\omega_0 R} \cdot E_{2s} \qquad (s = 1, \cdots, M) \qquad L_{2s} = \frac{R}{\omega_0} \cdot E_{2s}$$

$$L_{2s} = \frac{R}{\omega_0} \cdot \frac{\Omega_s^2}{E_{2s}} \qquad (s = 1, \cdots, M) \qquad C_{2s} = \frac{1}{\omega_0 R} \cdot \frac{\Omega_s^2}{E_{2s}}$$

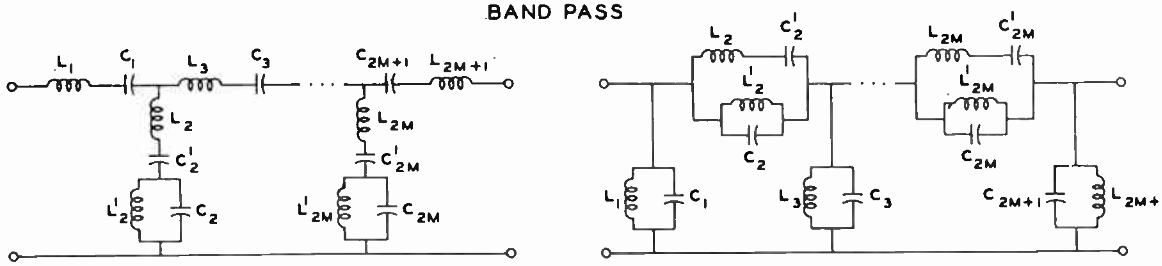
HIGH PASS

$$C_{2s+1} = \frac{1}{\omega_0 R} \cdot \frac{1}{E_{2s+1}} \qquad (s = 0, \cdots, M) \qquad L_{2s+1} = \frac{R}{\omega_0} \cdot \frac{1}{E_{2s+1}}$$

$$L_{2s} = \frac{R}{\omega_0} \cdot \frac{1}{E_{2s}} \qquad (s = 1, \cdots, M) \qquad C_{2s} = \frac{1}{\omega_0 R} \cdot \frac{1}{E_{2s}}$$

$$C_{2s} = \frac{1}{\omega_0 R} \cdot \frac{E_{2s}}{\Omega_s^2} \qquad (s = 1, \cdots, M) \qquad L_{2s} = \frac{R}{\omega_0} \cdot \frac{E_{2s}}{\Omega_s^2}$$

BAND PASS



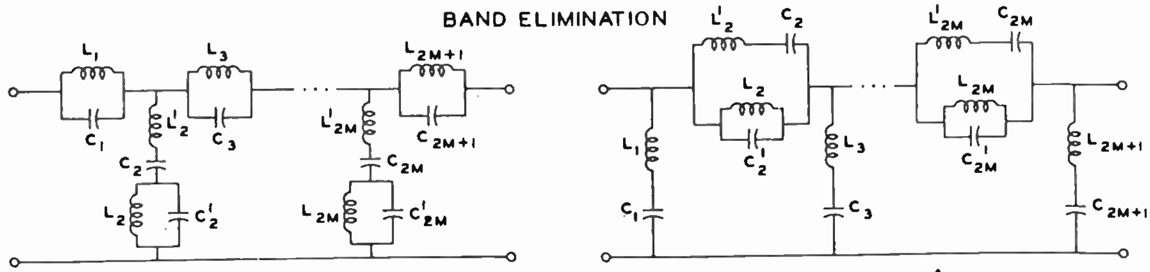
$$L_{2s+1} = \frac{R}{\Delta\omega_M} \cdot E_{2s+1} \qquad (s = 0, \cdots M,) \qquad C_{2s+1} = \frac{1}{\Delta\omega_M R} \cdot E_{2s+1}$$

$$C_{2s} = \frac{1}{\Delta\omega_M R} \cdot E_{2s} \qquad (s = 1, \cdots M,) \qquad L_{2s} = \frac{R}{\Delta\omega_M} \cdot E_{2s}$$

$$L_{2s} = \frac{R}{\Delta\omega_M} \cdot \frac{\Omega_s^2}{E_{2s}} \qquad (s = 1, \cdots M,) \qquad C_{2s} = \frac{1}{\Delta\omega_M R} \cdot \frac{\Omega_s^2}{E_{2s}}$$

$$L_{2s+1} C_{2s+1} = L_{2s} C'_{2s} = L'_{2s} C_{2s} = 1/\omega_M^2$$

BAND ELIMINATION



$$L_{2s+1} = \frac{\Delta R}{\omega_M} \cdot E_{2s+1} \qquad (s = 0, \cdots M,) \qquad C_{2s+1} = \frac{\Delta}{\omega_M R} \cdot E_{2s+1}$$

$$C_{2s} = \frac{\Delta}{\omega_M R} \cdot E_{2s} \qquad (s = 1, \cdots M,) \qquad L_{2s} = \frac{\Delta R}{\omega_M} \cdot E_{2s}$$

$$L_{2s} = \frac{\Delta R}{\omega_M} \cdot \frac{\Omega_s^2}{E_{2s}} \qquad (s = 1, \cdots M,) \qquad C_{2s} = \frac{\Delta}{\omega_M R} \cdot \frac{\Omega_s^2}{E_{2s}}$$

$$L_{2s+1} C_{2s+1} = L_{2s} C'_{2s} = L'_{2s} C_{2s} = 1/\omega_M^2$$

Fig. 14—Component values for ladder filters.

In a similar way, properties of the Jacobian elliptic functions can be deduced from properties of the Legendrian elliptic integral of the first kind. An elliptic integral of the first kind can be written in the following form:

$$u = F(\phi, k) = \int_0^\phi \frac{d\theta}{\sqrt{1 - k^2 \sin^2 \theta}}, \quad 0 \leqq k \leqq 1. \tag{43}$$

The quantity k is the "modulus," while θ is the variable of integration, and ϕ is the variable upper limit of integration, called the "amplitude" of the elliptic integral. This integral defines u as a function of ϕ when k is given a definite value. Some of the properties of the elliptic integral can be deduced by making use of the concept that a definite integral may be represented graphically as an area. According to (43), this area is bounded by the curve $1/\sqrt{1-k^2 \sin^2 \theta}$, the axis of θ, and the vertical lines erected at $\theta=0$ and $\theta=\phi$. This exhibits the fact that u is a function of the upper limit ϕ since, for a given k, the area under the curve depends only on the particular value of ϕ which is selected. As an illustration, the function $1/\sqrt{1-k^2 \sin^2 \theta}$ is plotted in Fig. 15. It is ob-

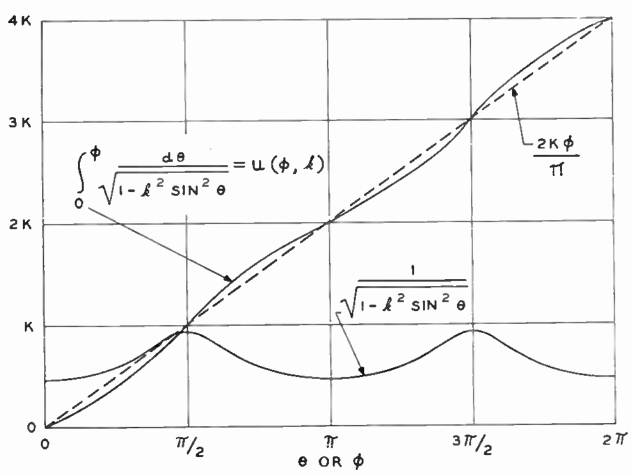


Fig. 15—Elliptic integral of the first kind.

served that this is a periodic function of θ admitting the period π, which has its minimum value, unity, when $\theta=0$, π, 2π, $\cdots$, and its maximum value $1/\sqrt{1-k^2}$ when $\theta=\pi/2$, $3\pi/2$, $\cdots$. Thus, not only does the integrand of (43) have the period π but it is symmetrical with respect to the ordinate $\pi/2$. Because of this symmetry, the area bounded by $\phi=0$ and $\phi=\pi/2$ is equal to the area bounded by $\phi=\pi/2$ and $\phi=\pi$, and likewise for each such strip bounded by $\phi=n\pi/2$ and $\phi=(n+1)\pi/2$. The value of u corresponding to this area is called the *complete* elliptic integral of the first kind, and is defined by:

$$K = \int_0^{\pi/2} \frac{d\theta}{\sqrt{1 - k^2 \sin^2 \theta}}. \tag{44}$$

As a consequence of the periodicity and symmetry of the integrand, the elliptic integral has the following properties:

$$u(n\pi + a, k) = 2nK + u(a, k)$$
$$u(\pi/2 + a, k) = 2K - u(\pi/2 - a, k).$$

This means that the value of the elliptic integral for a specified modulus can be found for any real value of the amplitude from a table giving the values of the integral in the interval 0 to $\pi/2$.

The dependence of the value of the elliptic integral on the value of the modular angle $\alpha = \text{arc sin } k$ is shown in Fig. 16. Since $0 \leqq k \leqq 1$, the limiting values of α are

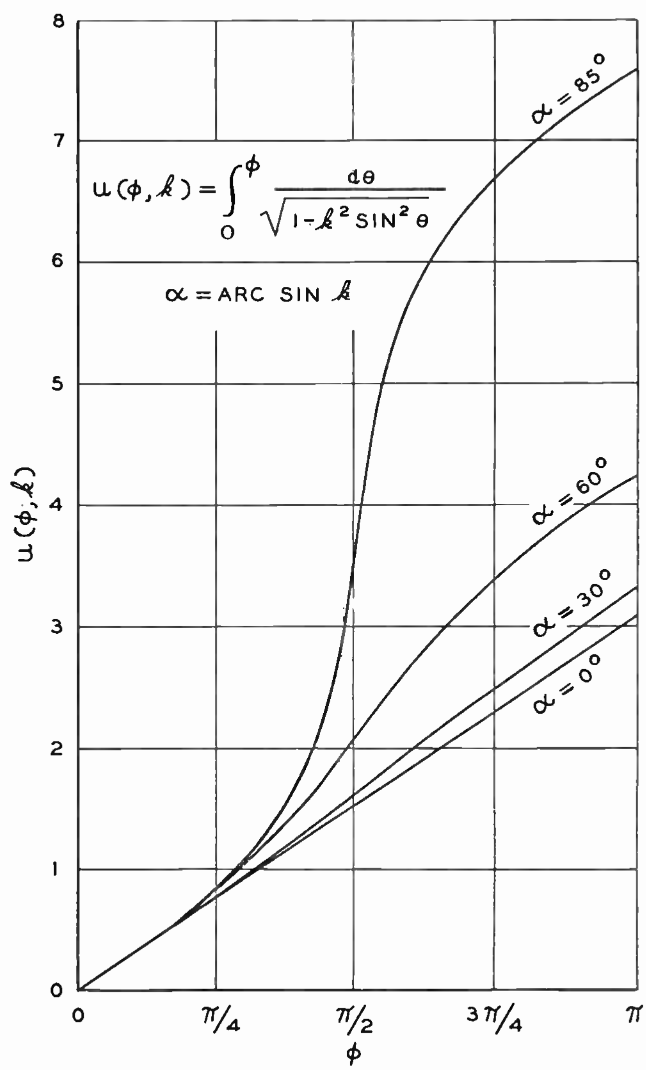


Fig. 16—Elliptic integral as a function of the modulus.

$0°(k=0)$ and $90°(k=1)$. For $\alpha=0°$, from (43),

$$u(\phi, 0) = \int_0^\phi d\theta = \phi.$$

A plot of $u(\phi, 0)$ against ϕ is the straight line designated by $\alpha=0°$. It represents the smallest value of the integral for a particular value of ϕ. For $\alpha = 90°$,

$$u(\phi, 1) = \int_0^\phi \frac{d\theta}{\cos\theta} = \text{Log}\tan\left(\frac{\pi}{4} + \frac{\phi}{2}\right).$$

This function is discontinuous at $\phi = \pi/2$ but up to that point it represents the greatest value of the elliptic integral for a particular value of ϕ. When the modular angle is greater than 0° and less than 90°, the curve for the elliptic integral lies in the region between the curves corresponding to the above extreme values.

An examination of Fig. 16 shows that there is a one-to-one correspondence between real values of ϕ and real values of u. Therefore, ϕ can be regarded as a function of u for a given k. The usual notation employed to express this relation is:

$$\phi = am(u, k) = \text{amplitude of } u \text{ for modulus } k.$$

Of greater interest to us are the functions defined:

$$\sin\phi = \text{sin amplitude } (u, k) = \text{sn } (u, k),$$

$$\cos\phi = \text{cos amplitude } (u, k) = \text{cn } (u, k).$$

$$\Delta\phi = \sqrt{1 - k^2 \sin^2\phi} = \text{dn } (u, k). \tag{45}$$

They are trigonometric functions of ϕ, the amplitude of (u, k). However, it is customary to regard them as functions of (u, k) itself.

Graphical representations of these functions for real u are determinable in a simple way from the plots of the elliptic integral as a function of the amplitude. As an illustration, we shall determine the set of elliptic functions of modular angle 60°. With the aid of Fig. 15, the results in Table II are evident.

TABLE II

u	ϕ	sn (u, k)	cn (u, k)	dn (u, k)
0	0	0	1	1
K	$\pi/2$	1	0	$k' = \frac{1}{2}$
$2K$	π	0	−1	1
$3K$	$3\pi/2$	−1	0	$k' = \frac{1}{2}$
$4K$	2π	0	1	1

Here, $k' = \sqrt{1 - k^2}$ is known as the complementary modulus. These results are displayed in Fig. 17. It is recognized that the details of these curves depend on the value of the modulus. The effect of changing the modulus in the case of the sn function is shown in Fig. 18. Not only does the period of the function increase with the value of the modulus, but the exact shape of the curve of the function changes with the modulus. This is brought out by using u/K as the abscissa.

At this point we may summarize some of the obvious properties of the elliptic functions. Of great importance is the fact that they are periodic functions. In particular, we note that:

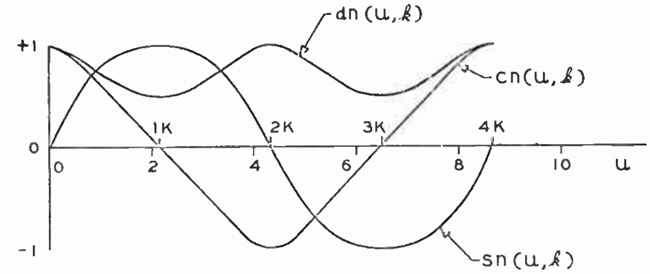


Fig. 17—The elliptic functions for real values of the argument.

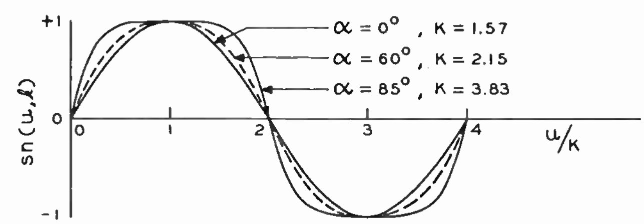


Fig. 18—The sn function as a function of the modulus.

$$\text{sn } (u, k) \text{ has the real period } 4K,$$
$$\text{cn } (u, k) \text{ has the real period } 4K,$$
$$\text{dn } (u, k) \text{ has the real period } 2K. \tag{46}$$

As a consequence of the symmetry and periodicity of these functions:

$$\text{sn } (-u, k) = -\text{sn } (u, k)$$
$$\text{cn } (-u, k) = \text{cn } (u, k)$$
$$\text{dn } (-u, k) = \text{dn } (u, k)$$

$$\text{sn } (K + u, k) = \text{sn } (K - u, k)$$
$$\text{cn } (K + u, k) = -\text{cn } (K - u, k)$$
$$\text{dn } (K + u, k) = \text{dn } (K - u, k)$$

$$\text{sn } (u + 2K, k) = -\text{sn } (u, k)$$
$$\text{cn } (u + 2K, k) = -\text{cn } (u, k)$$
$$\text{dn } (u + 2K, k) = \text{dn } (u, k)$$

$$\text{sn } (u + 4K, k) = \text{sn } (u, k)$$
$$\text{cn } (u + 4K, k) = \text{cn } (u, k)$$
$$\text{dn } (u + 4K, k) = \text{dn } (u, k). \tag{47}$$

Thus far we have been dealing with real values of the argument only and so it is a simple matter to proceed directly from a plot of the elliptic integral to an exhibit of the behavior of the elliptic functions. This, however, is not sufficient for our needs. We are interested also in the fact that these are doubly periodic functions, having an imaginary as well as a real period. The following definition of a doubly periodic function is taken from Whittaker and Watson's "Modern Analysis": "Let w_1 and w_2 be any two numbers (real or complex) whose ratio is not purely real. A function which satisfies the equations

$$f(z + 2w_1) = f(z),$$
$$f(z + 2w_2) = f(z),$$

for all values of z for which $f(z)$ exists is called a doubly periodic function of z with periods $2w_1$ and $2w_2$." The Jacobian elliptic functions form a particular class of such functions. This property may be demonstrated in a plausible way by calling upon the familiar circular and hyperbolic functions. The approach is through elliptic functions of extreme values of the modulus. For definiteness, we shall confine our attention to the sn function. Starting with the extreme value, $k=0$, we have from (43) and (45):

$$\text{sn}\,(z, 0) = \sin z \qquad (z = u + jv).$$

For real values of the argument ($z=u$) this is the familiar sine curve

$$\text{sn}\,(u, 0) = \sin u.$$

For pure imaginary values of the argument ($z=jv$), we find, with the aid of well-known relations between circular and hyperbolic functions

$$\text{sn}\,(jv, 0) = j \sinh v.$$

This function increases positively in value as v increases positively, and it increases negatively in value as v increases negatively; it is nonperiodic.

For complex values of the argument, we apply the addition theorem for the circular sine and obtain:

$$\text{sn}\,(z, 0) = \sin\,(u + jv) = \sin u \cosh v + j \cos u \sinh v.$$

This function may be studied by making a plot of the sn $(z, 0)$ plane on top of the z plane. In Fig. 19, the z plane is drawn as a grid with values of u and v indicated around the periphery. Values of the function sn $(u, 0)$ are spotted on the diagram along the u axis, *i.e.*, for $v=0$. It is evident that there is a repetition of values of sn $(u, 0)$ along the real axis in each interval 2π units in length. Next, values of sn $(jv, 0)$ are introduced at points along the v axis, *i.e.*, for $u=0$. It appears that sn $(jv, 0)$ will extend to infinitely great imaginary values with v in the positive and negative vertical directions. This brings out an essential point. If, for example, the argument z follows a path parallel to the real axis, the value of sn $(z, 0)$ is known for every value of z along this path provided the value of the function is known in the interval 0 to 2π; on the other hand, if z follows a path parallel to the imaginary axis, each value of z gives rise to a new value of sn $(z, 0)$. Consequently, a tabulation of sn $(z, 0)$ for values of z in the area extending from 0 to 2π along the real axis and from $-\infty$ to $+\infty$ along the imaginary axis gives the value of the function for any value of z. This fact is expressed by:

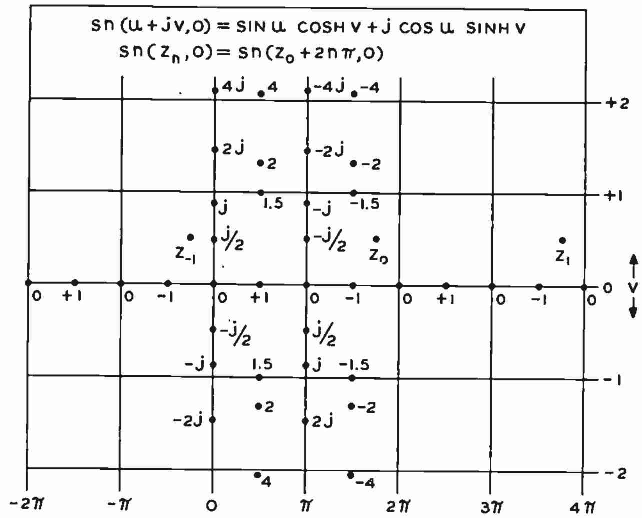


Fig. 19—The sn function of modulus zero for complex values of the argument.

$$\text{sn}\,(z + 2n\pi, 0) = \text{sn}\,(z, 0) \quad n = \pm 1, \pm 2, \cdots.$$

The second extreme value of the modulus of the elliptic functions is $k=1$. In this case, it is found from (43) and (45) that:

$$\text{sn}\,(z, 1) = \tanh z.$$

For real values of the argument ($z=u$), this function is 0 for $u=0$ and approaches $+1$ as u increases positively, and approaches -1 as u increases negatively; it attains its maximum value, unity, only when u becomes infinitely great; it is nonperiodic. If we imagine a sine curve stretched out along the real axis so that the point $\pi/2$ becomes situated at infinity, we would obtain a similar sort of curve. Stated in another way—we start with the elliptic sine (for real argument) of modulus zero and obtain a function which has the period 2π; then as the modulus increases, the value of the period increases until, for $k=1$, we arrive at a function which is nonperiodic, or has an infinite period.

For imaginary values of the argument ($z=jv$),

$$\text{sn}\,(jv, 1) = j \tan v.$$

This is the familiar tangent function. It becomes infinitely great at odd multiples of $\pi/2$. Furthermore, the values of the function are repeated in each interval of v which is π units in length extending from $(2n-1)\pi/2$ to $(2n+1)\pi/2$. Hence, the elliptic sine of modulus unity, for imaginary argument, has the period $j\pi$.

Here, too, if we may imagine the tangent curve stretched out along the imaginary axis in such a way that the point $\pi/2$ is pulled out to infinity, the result is somewhat similar to the curve of the hyperbolic sine. Thus, we may say that the elliptic sine (for imaginary argument) of modulus zero is a nonperiodic function; as the modulus increases, the value of the period decreases from an infinitely great value, through imaginary values until, for $k=1$, the function admits the period $j\pi$.

For complex values of the argument, the elliptic sine of modulus unity is given by:

$$\mathrm{sn}\,(z,1) = \tanh\,(u+jv) = \frac{\tanh u + j\tan v}{1 + j\tanh u \tan v}.$$

In general, this is a complex quantity. We may proceed as before and spot values of the function on a z-plane grid. This is done in Fig. 20. It is to be noted that values

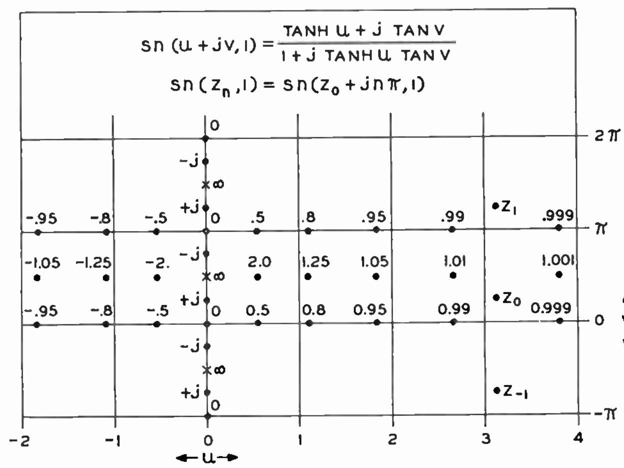


Fig. 20—The sn function of modulus unity for complex values of the argument.

of the function are repeated along the imaginary axis in each interval π units in length. From this it follows that, if the values of sn $(z, 1)$ are known for values of z in the area extending from 0 to $j\pi$ along the imaginary axis and from $-\infty$ to $+\infty$ along the real axis, the value of the function is known for any value of z. This fact is expressed by:

$$\mathrm{sn}\,(z + jn\pi, 1) = \mathrm{sn}\,(z, 1), \quad n = \pm 1, \pm 2, \cdots.$$

This study of the elliptic sine for extreme values of the modulus may be summarized in the following way: the elliptic sine of modulus zero has the real period 2π and an infinite imaginary period; the elliptic sine of modulus unity has the imaginary period $j\pi$ and an infinite real period; as the modulus increases from zero to unity, the real period increases from 2π to an infinite value, and the imaginary period decreases from an infinite value to $j\pi$. Viewing the situation in this light, it seems reasonable to draw the conclusion that elliptic functions of modulus greater than zero and less than unity are doubly periodic functions, having one real and one imaginary period.

The real periods have been found previously and are specified in (46). The imaginary periods may be determined in a similar manner by examining the functions for pure imaginary arguments. We start with the elliptic integral of the first kind, but in anticipation of the result, we denote its value by jv, or,

$$jv = \int_0^{\phi} \frac{d\theta}{\sqrt{1 - k^2 \sin^2\theta}}.$$

The elliptic functions are defined, in accordance with (45), by:

$$\mathrm{sn}\,(jv, k) = \sin\phi$$
$$\mathrm{cn}\,(jv, k) = \cos\phi$$
$$\mathrm{dn}\,(jv, k) = \sqrt{1 - k^2\sin^2\phi}.$$

Now, the variable of integration and the variable limit are subjected to the transformations,

$$\sin\theta = j\tan\theta',$$
$$\sin\phi = j\tan\psi.$$

Upon making these substitutions, we obtain:

$$v = \int_0^{\psi} \frac{d\theta'}{\sqrt{1 - (k')^2\sin^2\theta'}},$$

where $k' = \sqrt{1-k^2}$ is the complementary modulus. The corresponding elliptic functions are,

$$\mathrm{sn}\,(v, k') = \sin\psi,$$
$$\mathrm{cn}\,(v, k') = \cos\psi,$$
$$\mathrm{dn}\,(v, k') = \sqrt{1 - (k')^2\sin^2\psi} \qquad (48)$$

By combining the above relations, elliptic functions for imaginary argument may be expressed in terms of the elliptic functions for real argument in the following way:

$$\mathrm{sn}\,(jv, k) = j\,\frac{\mathrm{sn}\,(v, k')}{\mathrm{cn}\,(v, k')},$$
$$\mathrm{cn}\,(jv, k) = \frac{1}{\mathrm{cn}\,(v, k')},$$
$$\mathrm{dn}\,(jv, k) = \frac{\mathrm{dn}\,(v, k')}{\mathrm{cn}\,(v, k')}. \qquad (49)$$

We recall that the complete elliptic integral K plays an important part in the theory concerned with functions of modulus k. Quite analogous to (44) is the definition of the complementary complete integral:

$$K' = \int_0^{\pi/2} \frac{d\theta}{\sqrt{1 - (k')^2\sin^2\theta}}. \qquad (50)$$

This, of course, may be interpreted as the area under the curve $1/\sqrt{1-(k')^2\sin^2\theta}$ between 0 and $\pi/2$, which is just like the definition of K except that k has been replaced by $k' = \sqrt{1-k^2}$. That is, K' bears the same relation to functions of modulus k' as K does to those of modulus k. Thus, by (46), K' is a quarter period of sn (v, k') and cn (v, k'), and a half period of dn (v, k'). One can visualize the course of functions of imaginary argument by combining the relations (49) and Fig. 17, where K is now replaced by K', and u by v. For example, the values of sn (jv, k) are pure imaginary, and the curve $[-j \text{ sn } (jv, k)]$ has the value zero at $v=0$, and increases in the positive sense until at $v=K'$ it is infinitely great. Between K' and $2K'$ it is negative, and decreases in magnitude to zero at $v=2K'$. The values in the interval $2K'$ to $4K'$ are a repetition of those in the interval 0 to $2K'$. The other two functions can be traced in the same way. We arrive at the conclusion that:

the sn function has the imaginary period $j\,2K'$,

the cn function has the imaginary period $j\,4K'$,

the dn function has the imaginary period $j\,4K'$. (51)

(While it is true that $j4K'$ is a period of the cn function, the primitive period is $2K+j2K'$.)

There are many important theorems concerning doubly periodic functions which enable one to prove the results used in the design of Tchebycheff parameter filters. These are mentioned in Darlington's paper. They are quite beyond the scope of this appendix, and we shall close with a discussion of one important property. For definiteness, we consider the elliptic sine. If the argument is denoted by $z=u+jv$, then:

$$\text{sn}\,(z + 4mK,\, k) = \text{sn}\,(z,\, k).$$

$$\text{sn}\,(z + j2nK',\, k) = \text{sn}\,(z,\, k).$$

These relations may be combined as:

$$\text{sn}\,(z + 4mK + j2nK',\, k) = \text{sn}\,(z,\, k), \quad (m, n = 0, \pm 1, \pm 2, \cdots). \tag{52}$$

This means that the function has the same value at z and at $(z+4mK+j2nK')$. If the complex z plane is divided into parallelograms, as shown in Fig. 21, by

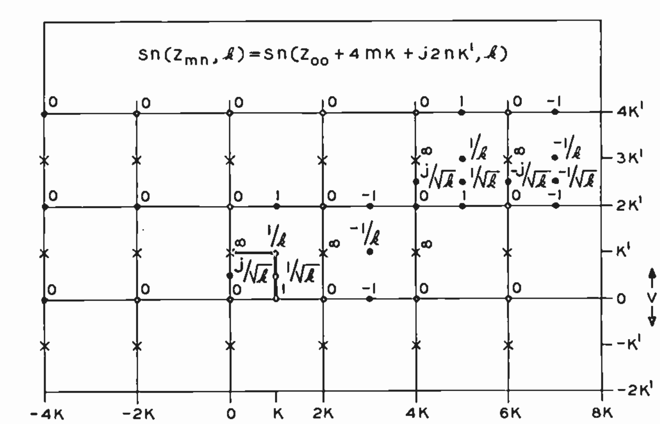


Fig. 21—The sn function of modulus k for complex values of the argument.

means of the vertical lines $u=4mK$ and the horizontal lines $v=2nK'$, we obtain the "period-parallelograms"; the particular one with vertices at (0, 0), $(4K, 0)$, $(4K, 2K')$, $(0, 2K')$ is called the "fundamental period-parallelogram." If the function is known for each value of z within and along two adjacent sides of a parallelogram, it is known for any value of z. Several special values are indicated in Fig. 21. They are computed with the aid of the formula:

$$\text{sn}\,(x \pm y) = \frac{\text{sn}\,x\,\text{cn}\,y\,\text{dn}\,y \pm \text{cn}\,x\,\text{sn}\,y\,\text{dn}\,x}{1 - k^2\,\text{sn}^2\,x\,\text{sn}^2\,y} \tag{53}$$

where all the functions are of modulus k. This is one of the many useful elliptic function formulas included in Peirce's "Table."[6]

The transformation of variable that is used in the main part of this paper is $\Omega=\sqrt{k}$ sn (z, k). It appears from the preceding discussion that this substitution transforms the whole Ω plane into one period-parallelogram of the z plane. Now we shall determine the values of z that correspond to the *positive* real frequency axis, using Fig. 21 as a guide. Starting at $z=0$, we follow the real axis to $z=u=K$. Since $\Omega=0$ for $u=0$, and $\Omega=\sqrt{k}$ for $u=K$, the pass band frequency interval corresponds to real values of z between 0 and K, inclusive. If we continue along the real axis, we obtain either a repetition of these values of Ω, or their negatives. So at $z=K$, we set out along the path $z=K+jv$. The corresponding values of frequency are computed by means of the summation formula (53) and the relations (49) between functions of imaginary and real argument. We find:

$$\Omega = \sqrt{k}\,\text{sn}\,(K + jv,\, k) = \frac{\sqrt{k}\,\text{dn}\,(v,\, k')}{\text{cn}^2\,(v,\, k') + k^2\,\text{sn}^2\,(v,\, k')} \cdot$$

But $\text{cn}^2\,v=1-\text{sn}^2\,v$, and $1-(k')^2\,\text{sn}^2\,v=\text{dn}^2\,v$, (modulus k'), which simplifies the equation to:

$$\Omega = \frac{\sqrt{k}}{\text{dn}\,(v,\, k')} \cdot$$

With the special values tabulated earlier we see that $\Omega=\sqrt{k}$ for $v=0$, and $\Omega=1/\sqrt{k}$ for $v=K'$. Thus, the transition band frequency interval corresponds to the interval $z=K+jv$, with v between 0 and K', inclusive. Continuation along this path yields nothing new. Hence, at $z=K+jK'$, we follow the path $z=u+jK'$. Upon performing the same sort of calculations as above, we get:

$$\Omega = \sqrt{k}\,\text{sn}\,(u + jK',\, k) = \frac{1}{\sqrt{k}\,\text{sn}\,(u,\, k)} \cdot$$

When $u=K$, $\Omega=1/\sqrt{k}$, and Ω is infinite for $u=0$. Therefore, the attenuation band frequency interval corresponds to values of $z=u+jK'$, with u between 0 and K, inclusive. This demonstrates that the positive real frequency axis corresponds to three sides of a parallelogram that is contained within a period-parallelogram of the z plane.

[6] B. O. Peirce, "A Short Table of Integrals," Ginn & Co., New York, N. Y., 3rd ed., pp. 84–86; 1929.

A New Semiconductor Photocell Using Lateral Photoeffect*

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Summary—The effect of illumination of a semiconductor junction is, as is well-known, a photovoltage between the two sides of the junction. In this article it will be shown that a nonuniform illumination gives a lateral photovoltage parallel to the junction in addition to the (transverse) photovoltage mentioned above.

A photocell will be described that uses the lateral effect and can detect the position of a light spot to less than 100 Å. By utilizing an associated lens or aperture, one can measure an angular motion smaller than 0.1 second of arc. The output voltage of the cell is a linear function of the position of the light spot, with zero output for the light spot in the center, reversing in sign when the light spot changes from one side to the other of the center position. The linearity is better than 1.5 per cent over a distance of 0.030 inch. The equivalent noise resistance of the cell is equal to its output resistance, approximately 100 ohms. The sensitivity of the cell is approximately 200 microamperes per lumen and its frequency response is about the same as that of junction transistors.

The response curve can be shifted by the application of a voltage between the base contacts. This is an electronic equivalent of a mechanical translation of the cell. It is also possible to do the equivalent of "chopping" the light by applying a modulating voltage to the alloyed dot.

Introduction

The effect of illumination of a semiconductor junction is, as has long been known, a photovoltage between the two sides of the junction. If the illumination is nonuniform, an additional effect arises which has not been recognized earlier and which will be the subject of this article, namely the development of a photovoltage parallel to the junction in addition to the photovoltage between the two sides of the junction mentioned above. To distinguish between the two effects, the latter photovoltage will be called *transverse* (with respect to the junction) whereas the new effect will be called *lateral* photovoltage (lateral with respect to the junction).

This lateral photovoltage has been utilized in a new type of photocell whose most interesting characteristic is a *photosensitivity that varies from a positive to a negative value over its surface*. This means that a point source of light, imaged on the photocell by a lens, will produce a signal that varies with the angle between the direction to the light and the symmetry axis of the photocell-lens combination. When the direction to the light coincides with the symmetry axis of the cell, the signal is zero. When the light direction deviates from the symmetry axis in one direction a signal of one polarity is obtained and when it deviates in the opposite direction a signal of opposite polarity is obtained. The cell can therefore measure the direction to a light source by a null method and consequently with the high accuracy of such methods.

The photocell is very simple and may be constructed in a manner well known from transistor technology. It consists typically of a germanium wafer on which a junction has been applied by alloying an indium dot onto the wafer. Two ohmic contacts, one at each end of the junction, are used for picking up the lateral photovoltage. It may be remarked that no contact is needed to the indium dot, which may be left electrically floating.

A further valuable feature of the cell is that mechanical rotation of the cell to aim its optical axis in different directions can be replaced by a method of *electronic sweeping*.

A desirable detail of the electronic sweeping is a method of *electronically chopping the light signal*, which has merits of its own in some applications.

The Lateral Photoeffect

When hole-electron pairs are injected locally under a semiconductor junction, a lateral photovoltage is developed in addition to the well-known transverse photovoltage across the junction. This lateral photovoltage is related to the feed-in feed-out effect described by Moore and Webster.[1] The lateral photoeffect is the explanation of some of the difficulties that have been encountered in measuring lifetime with the Valdes[2] method in the presence of surface layers.

Consider a junction, as shown in Fig. 1, between an n-type region and another more heavily doped p-type region (herein denoted p^+). The arguments to follow apply as well to a p-n^+ junction and also, with some modifications shown later, to other types of junctions.

Referring to Fig. 1 assume a beam of light, which injects hole-electron pairs at the point A. The consequences of the injection may be thought of as a sequence of events as follows. From the equilibrium situation before injection the junction goes over to a new steady-state condition in which most of the injected holes are found in the p^+ region, most of the injected electrons in the n region. This is accompanied by a shift in Fermi levels constituting the well-known transverse photovoltage indicated as V_a in Fig. 1.

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[1] A. R. Moore and W. M. Webster, "The effective surface recombination of a germanium surface with a floating barrier," Proc. IRE, vol. 43, pp. 427–435; April, 1955.

[2] L. B. Valdes, "Measurement of minority carrier lifetime in germanium," Proc. IRE, vol. 40, pp. 1420–1423; November, 1952.

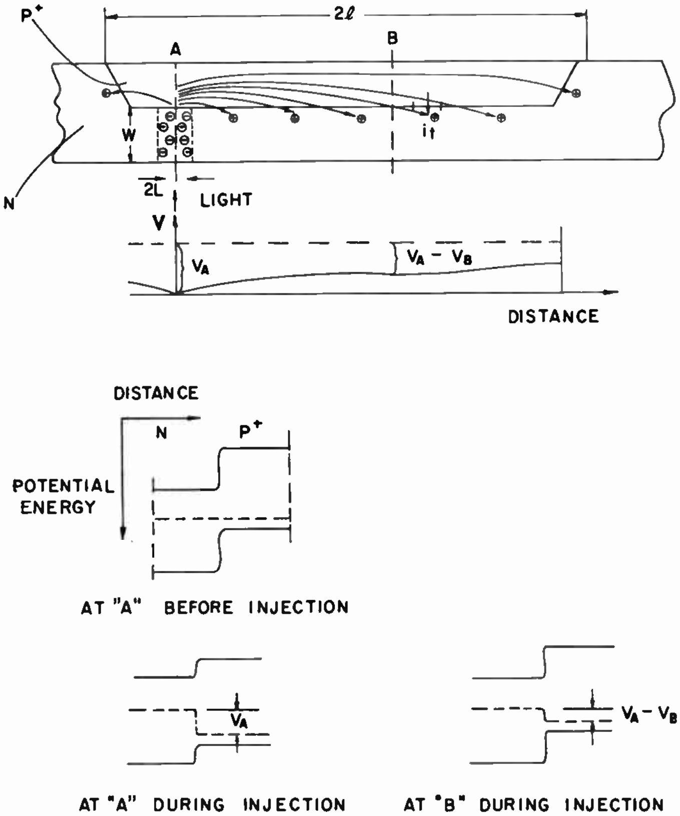

Fig. 1—Illustration of charge movement, potential distribution, and band picture for the lateral photoeffect.

However, if the conductivity of the p^+ region is much larger than in the n region, as would be the case for an alloyed construction, the p^+ region may be thought of as an equipotential region and the holes will instantaneously redistribute themselves uniformly over the region. At any other point, then, a deviation from equilibrium will appear resulting in a transfer of holes back into the n region. These reinjected holes are minority carriers in the n region. A lateral field is therefore set up so as to move majority carriers from the point of illumination to the point of reinjection, thereby accomplishing charge neutralization. The potential V in Fig. 1 characterizes this lateral field.

From the foregoing it is immediately apparent that an n^+ region on a p base also gives a lateral photovoltage. Also an n^+ region on an n type base behaves in quite an analogous manner. In this case, the electrons are swept up by the n^+ and after redistribution are reinjected into the n region. The holes left behind at the point of illumination are minority carriers and therefore a field is set up to move the electrons back to the point of illumination. The lateral photovoltage in this case, then, has opposite polarity compared to the p^+ on n case.

It is easy to show that n^- and p^- layers, *i.e.*, layers less heavily doped than the base layer, also show the lateral photoeffect. In other words, the lateral photoeffect under nonhomogeneous illumination (as well as the transverse photoeffect) is a completely general characteristic of any junction, if a junction is defined as a transition between regions of different conductivities. An approximate sketch of the field is plotted in Fig. 2.

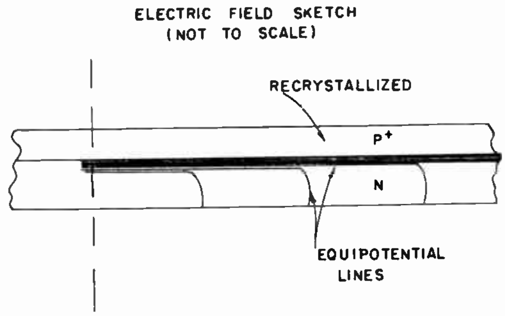

Fig. 2—Equipotential map of the high-resistivity region at injection. (Not to scale)

Fig. 3 shows the direction of the lateral photovoltage for different doping combinations.

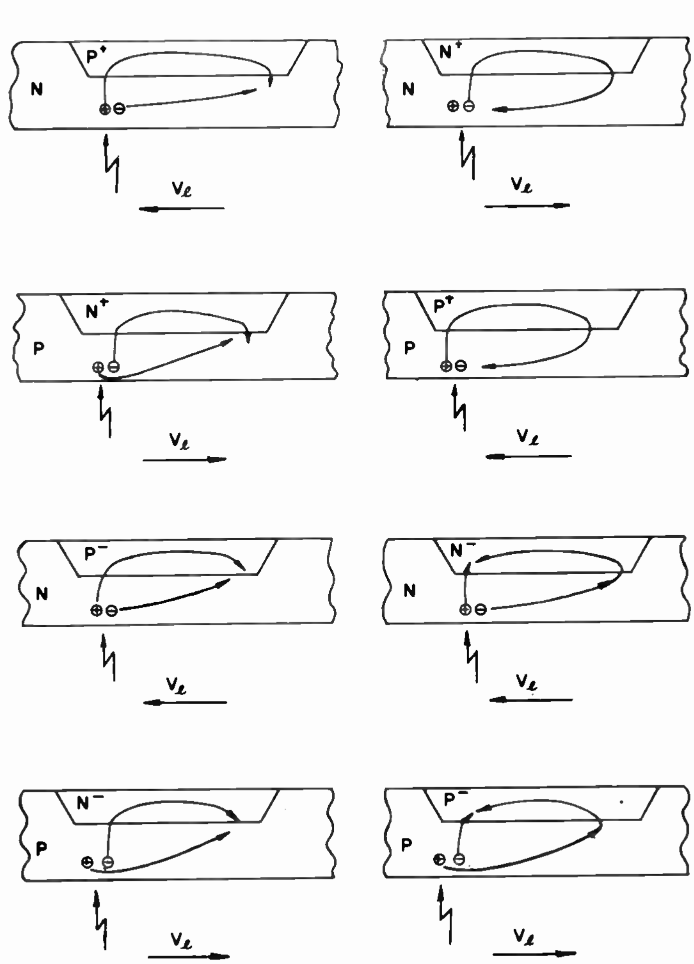

Fig. 3—The lateral photovoltage for different doping combinations.

If a voltage is applied between the dot and the base, the resulting current flow may be thought of as a superposition of lateral and transverse photocurrents. The same is, of course, true if carriers are injected by methods other than light injection. However, for transverse currents so large that the voltage drop in the n region becomes appreciable the lateral photoeffect decreases which fact is the basis for the electronic chopping to be described later.

Calculation of the Lateral Field

An approximate expression for the lateral field will now be derived. The analysis will cover the two cases p^+ on n and n^+ on n separately as the field is entirely different in the two cases.

A) P^+ on N

Assume quasi-infinite dimensions perpendicular to the plane of Fig. 1. Assume that the illumination creates a potential difference, V_A, (transverse photovoltage) between the p^+ and the n region at the point of illumination, A, here taken at the left end of the junction. It is also assumed that the diffusion length for minority carriers, L, is small compared to the lateral dimensions of the junction, $2l$, but large compared to the width of the n region, w.

$$w \ll L \ll 2l.$$

This means that most of the light-injected carriers will reach the p^+ region close to the point A. At any other point not too close to A the transverse potential difference will be $V_A - V$, where V is the potential drop in the lateral direction caused by the return current in the n region

$$V = -\int_0^x \rho i_l dx$$

where

ρ = resistivity

i_l = lateral return current density.

The transverse current density may be calculated from conventional junction theory.[3]

$$i_t = i_s[\exp \lambda(V_A - V) - 1]$$

where

$$\lambda = q/kT$$

and i_s is given, *e.g.*, by Webster.[4] The rate of change of the electric field in the n region is given by the increments of reinjected current.

$$\frac{d^2V}{dx^2} = -\frac{\rho i_s}{w^2}[\exp \lambda(V_A - V) - 1]. \tag{1}$$

Multiplication with dV/dx and integration gives

$$\frac{dV}{dx} = \left\{\frac{\rho i_s}{\lambda w}[\exp \lambda(V_A - V) + \lambda V + C_1]\right\}^{1/2} \tag{2}$$

[3] W. Shockley, "Electrons and Holes in Semiconductors," D. van Nostrand Co., Inc., New York, N. Y., p. 316; 1950.

[4] W. M. Webster, "Saturation current in alloy junctions," Proc. IRE, vol. 43, pp. 277–280; March, 1955. See (6), p. 277.

where C_1 is an integration constant. At $x=0$, $V=D$ and dV/dx may be derived from

$$\left(\frac{dV}{dx}\right)_{x=0} = \rho(i_l)_{x=0}$$

where ai_l is the total photocurrent, I

$$I = b\int_0^{2l} i_t dx \qquad a = bw.$$

Then

$$C_1 = \frac{\lambda w \rho I^2}{i_s a^2} - \exp \lambda V_A. \tag{3}$$

From measured values of I and V_A, typical values for an illumination of approximately $\frac{1}{2}$ lumen being

$$V_A = 70mV$$

$$I = 50\mu a.$$

It follows that with good accuracy (2) and (3) may be written

$$\frac{dV}{dx} \approx \left(\frac{\rho i_s}{\lambda w} C_1\right)^{1/2} \tag{4}$$

$$C_1 \approx \frac{\lambda w \rho I^2}{i_s a^2}. \tag{5}$$

Further integration of (4) gives simply

$$V \approx \frac{\rho}{a} I x \tag{6}$$

B) N^+ on N

With the same assumptions as in the previous section the lateral field for the case of an n^+ region on an n-type base may be derived. This junction is nonrectifying and the transverse return current is given by Ohm's law

$$i_t = \frac{V_A - V}{\rho\delta} \tag{7}$$

where δ is the distance in which the transverse field changes over to lateral and w is the width of the base region as before.

The rate of change of the electric field in the N region

$$\frac{d^2V}{dx^2} = \frac{1}{2w}\rho i_t$$

$$= \frac{V_A - V}{2w\delta}. \tag{8}$$

Multiplication with dV/dx and integration gives

$$\frac{dV}{dx} = \left\{\frac{1}{2w\delta}[(V_A - V)^2 + C_2^2]\right\}^{1/2}. \tag{9}$$

At $x=0$, $V=0$ and as before $(dV/dx)_{x=0} = \rho i_l$

$$i_l = \frac{I}{a}.$$

Then

$$C_2^2 = \frac{2w\rho^2 I^2 \delta}{a^2} - V_A^2.$$

Further integration gives

$$(2\delta w)^{1/2} \sinh^{-1} \frac{(V_A - V)}{C_2} = -(x + x_0)$$

or

$$\frac{V_A - V}{C_2} = -\sinh \frac{x + x_0}{(2w\delta)^{1/2}}$$

where x_0 is an integration constant defined by the boundary conditions $x=0$, $V=0$

$$V = C_2 \sinh \frac{x}{(2w\delta)^{1/2}}. \qquad (10)$$

The Lateral Field Photocell

The use of the lateral photoeffect in some experimental photocells will now be described. The construction is shown in Fig. 4. The cell consists of a germanium wafer, of 1–2 ohm cm resistivity, of transistor dimensions, *i.e.*, approximately 0.005-inch thick and ¼ inch square. In a p-n junction photocell, an indium dot is alloyed on the wafer. The indium dot is 0.045-inch diameter and alloyed to a depth of 0.002-inch. Two base contacts are applied symmetrically. Some experimental units are shown in Fig. 5.

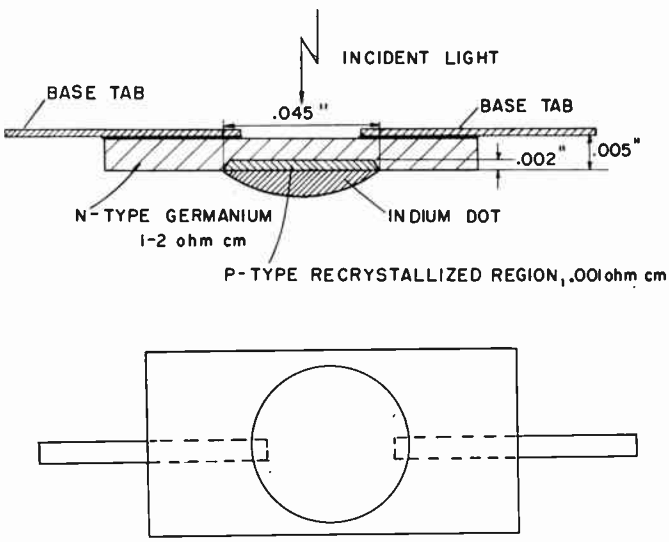

Fig. 4—Photocell using lateral photoeffect.

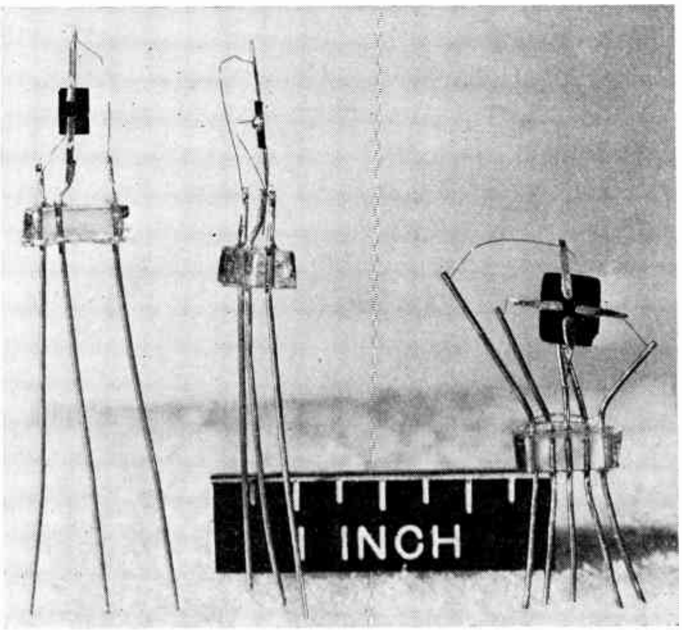

Fig. 5—Experimental samples.

Consider first a beam of light incident on the center of the cell directly under the dot. Two lateral photovoltages will be set up, of opposite polarity but equal magnitude, so that the net voltage between the base contacts is zero. When the beam of light is moved towards one side of the cell, a net output voltage equal to the difference between the two lateral photovoltages will appear between the base contacts. The output voltage may be calculated from (6). Let us assume illumination at a point x_1. Then the output voltage is the sum of two terms

$$V_{\text{out}} = \frac{\rho}{a}\frac{Ix_1}{2l}(-x_1) + \frac{\rho}{a}\frac{I(2l - x_1)}{2l}(2l - x_1)$$

or

$$= 2\frac{\rho}{a} I(l - x_1). \qquad (11)$$

This equation represents a straight line going through zero at the point $x_1 = l$ which corresponds to the center point of the photocell. Fig. 6 shows an experimental curve.

A similar derivation may be carried out for the n^+ on n photocell giving

$$V_{\text{out}} = C \sinh \frac{(l - x_1)}{(2w\delta)^{1/2}}. \qquad (12)$$

Fig. 7 shows an experimental curve with a theoretical curve fitted to it.

The principle may be extended to two coordinates by introducing a second set of base contacts as shown in Fig. 8. In this cell any deviation of the light beam off dead center will be registered by a pair of interbase voltages whose amplitudes and signs uniquely specify the location of the light spot.

Electronic Sweeping

The influence of a superimposed bias current from an external battery between the base contacts may be derived from (11). Assume the battery voltage V_b with the resistance R_b and the cell resistance R_e. Then

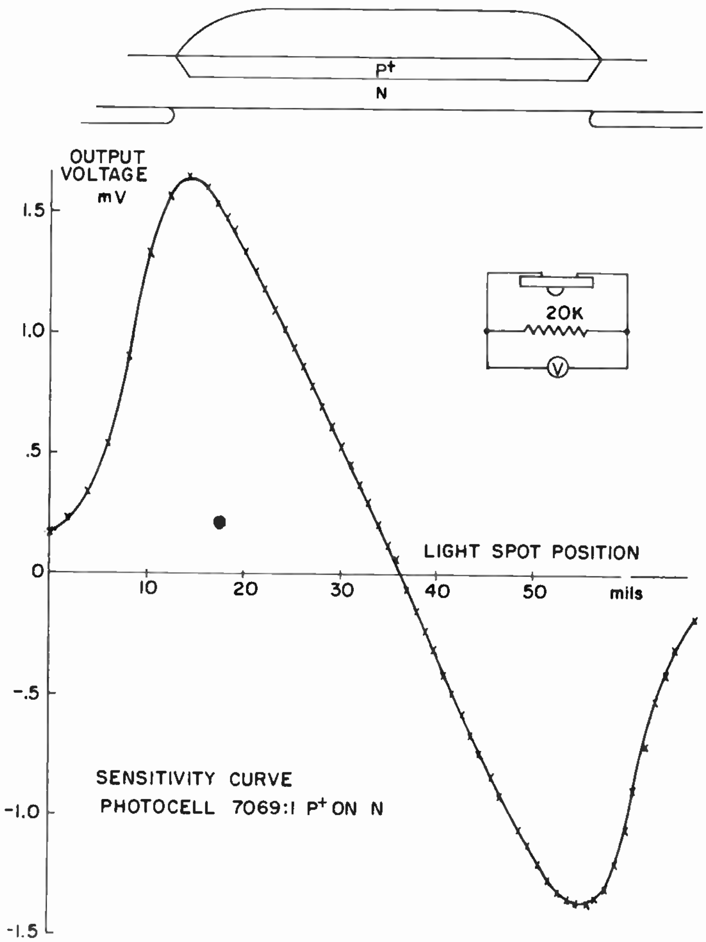

Fig. 6—Response curves for p^+ on n photocell.

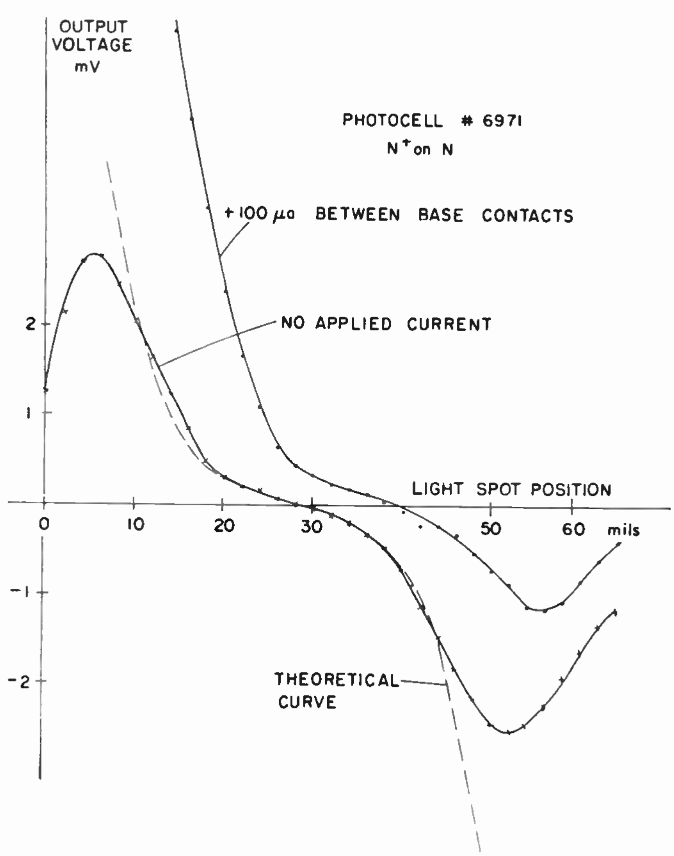

Fig. 7—Response curves for p^+ on n photocell.

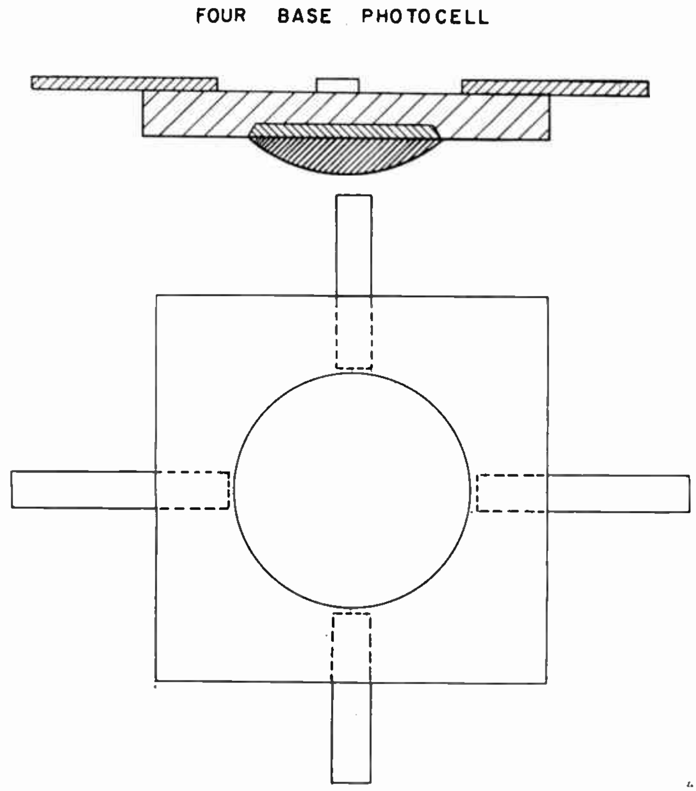

Fig. 8—Four-base photocell.

$$V_{\text{out}} = 2\,\frac{\rho}{a}\,I(l - x_1) + \frac{V_b R_c}{R_c + R_b}$$

$$= 2\,\frac{\rho}{a}\,I(l + C_5 V_b - x_1) \qquad (13)$$

$$C_5 = \frac{R_c}{R_c + R_b} \cdot \frac{a}{2\rho I}\,.$$

This represents a shift of the sensitivity curve along the x axis as V_b is varied. Fig. 9 shows experimental results. As seen in Fig. 9, the curve shifts to one side for the bias current in one direction, and towards the other side for the bias current in the other direction. If a light chopper is inserted in the path of the light, the desired signal is converted to ac and can be separated from the sweep current by a (high-pass or band-pass) filter.

This shifting of the sensitivity curve is comparable to a mechanical turning of the whole photocell as is apparent from Fig. 10.

The linearity of the sweep, *i.e.*, the shift of the curve per unit sweep current is quite good. This would be evident from a family of curves such as shown in Fig. 9. However, it is more convenient to measure the shift of one point rather than retaking the whole curve each time. Also, for convenience, it was simpler to measure the shift vertically (measurement of current) rather than horizontally. Fig. 11 shows the results of such a measurement. The linearity is beyond the accuracy of measurement, the largest deviation from the straight line for any one point being of the order of 1 per cent of the maximum value.

This method of electronic sweeping may also be applied to a conventional phototube provided with two

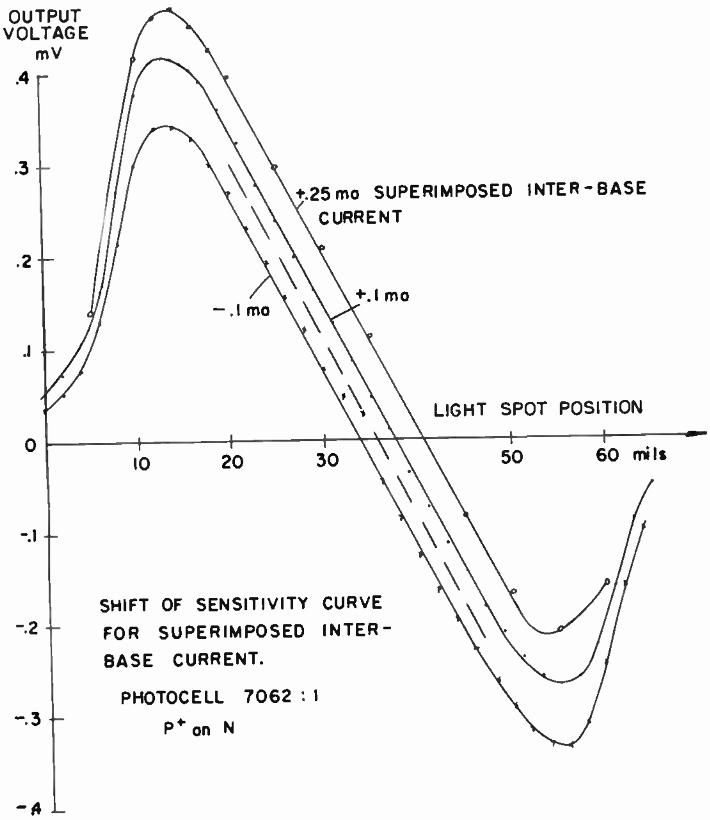

Fig. 9—Response curves for p^+ on n photocell. Superimposed interbase current parameter.

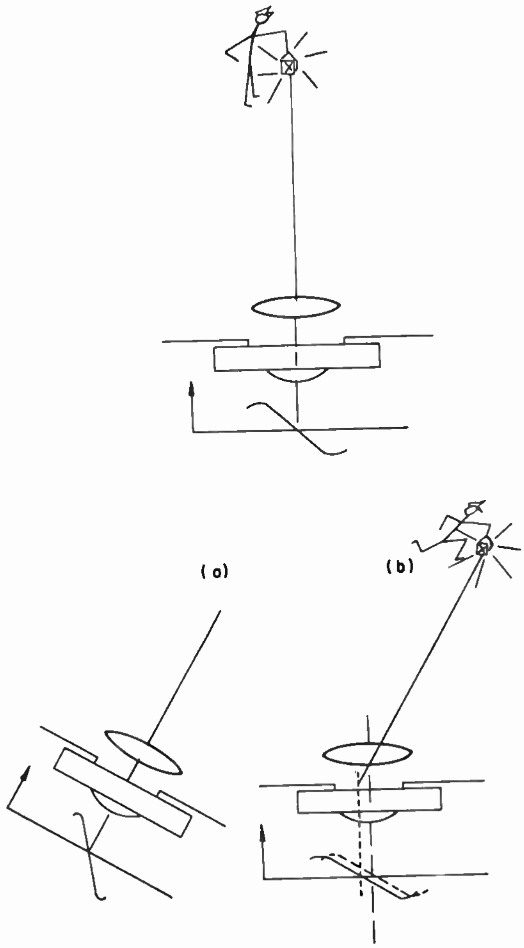

Fig. 10—Illustration of principle for electronic sweep. Electronic sweep, b, is equivalent to a turning of the photocell, a.

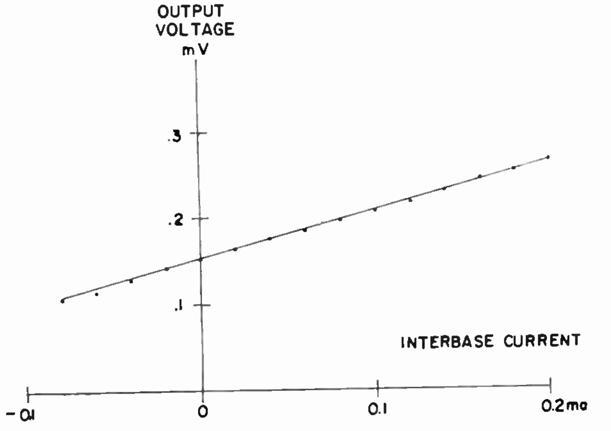

Fig. 11—Linearity of electronic sweep.

anodes and sweep electrodes to shift the current from one anode to the other.

Electronic Light Chopping

The use of a sweeping current described in the previous section provides an electronic rather than a mechanical method of scanning. In many photocell applications the advantages of amplification of an ac rather than a dc signal is obtained by mechanically interrupting the light signal. It would be convenient if this could also be done by electronic means. This may be done in the present photocell by applying an ac biasing voltage to the junction electrode.

Fig. 12 shows the influence on the sensitivity curve of a p^+ on n photocell of a forward current between the dot

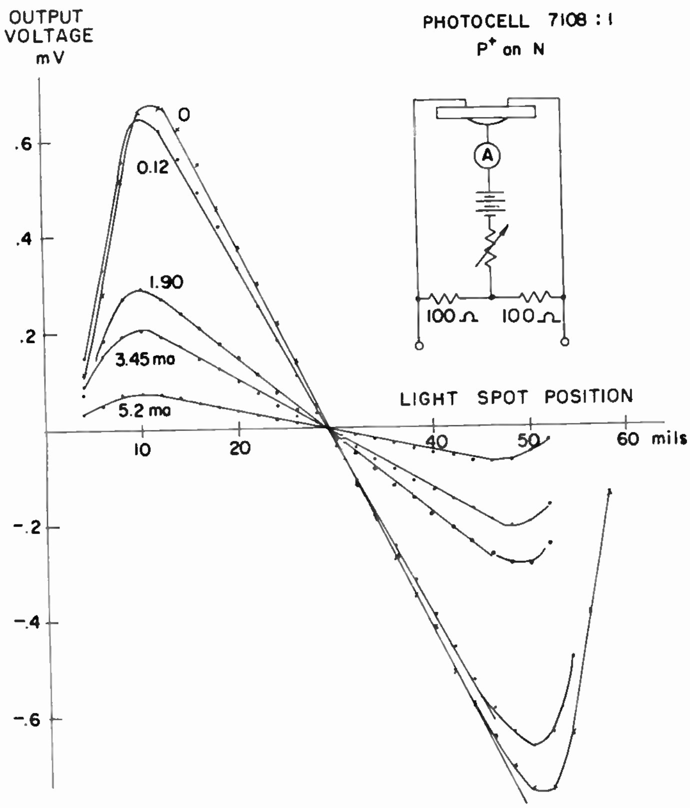

Fig. 12—Influence of forward current through junction for p^+ on n photocell.

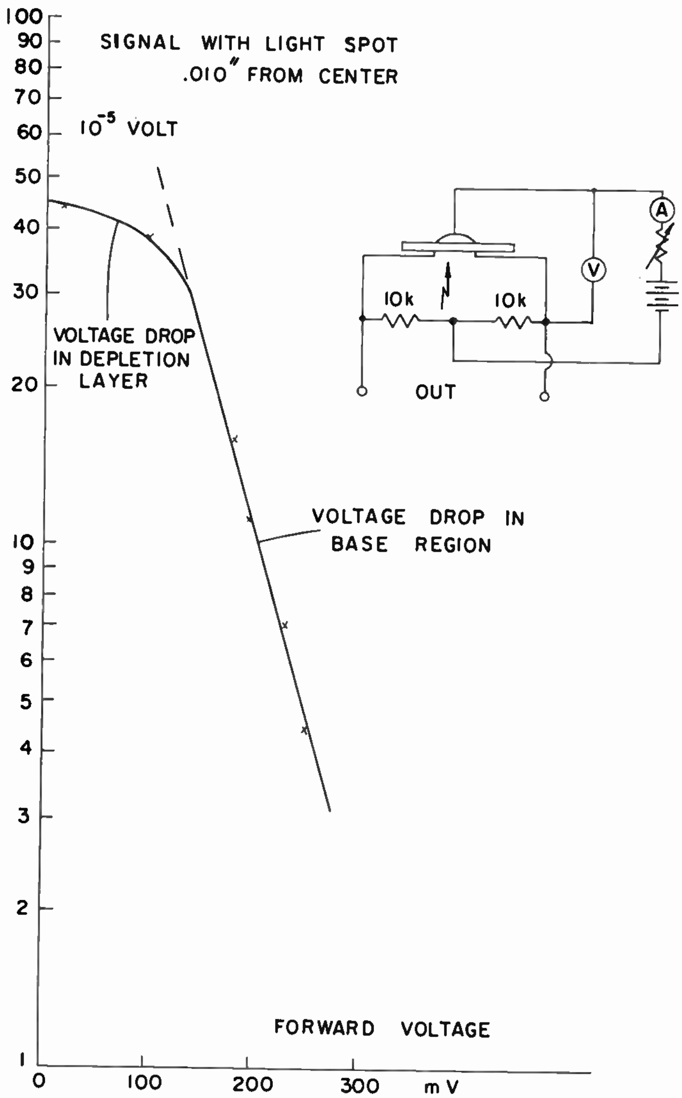

Fig. 13—Influence of forward current through junction for p^+ on n photocell. For small current the voltage drop in the depletion layer dominates, for large current the voltage drop in the base region.

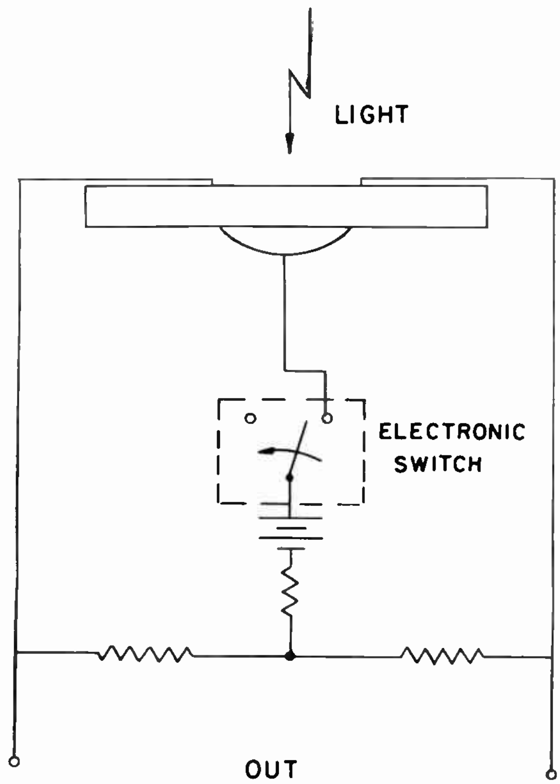

Fig. 14—Principle of electronic chopping.

and the two base contacts. A forward current of approximately 5 ma reduces the sensitivity to approximately 10 per cent of its original value. This decrease in sensitivity is caused by the potential drop in the base region, created by the forward current, and prevents the carriers injected by the light beam from entering the dot.

Fig. 13 shows the signal as a function of forward voltage.

In a practical arrangement, then, the dot may be switched between two states, disconnected (floating) and connected to the positive end of a battery as shown in Fig. 14. This may be accomplished by a simple switching transistor oscillator working as an astable multivibrator.

This electronic chopping method may of course also be applied to conventional photocells, again with the advantage that moving parts can be avoided. However, noise, leakage current, etc., are chopped at the same time and therefore cannot be separated out. This may be a disadvantage in some cases.

Linearity and Slope of the Response Curve

Fig. 15 shows a careful measurement of the response curve. The maximum deviations from a straight line amount to approximately 1 per cent which, however, coincides with the accuracy of the measurement. To obtain a better estimate, two similar cells were mounted side by side and connected with the output voltages opposing each other. The net output is then the difference between the two showing nonsystematic deviations from linearity. It is to be noted that this method does not in itself reveal systematic deviations from linearity such as the bending off near the maxima. If the two cells deviate with the same amount, no output is obtained. To minimize systematic deviations, two cells with somewhat different interbase distance were compared. The result is shown in Fig. 16, which shows a maximum deviation of 1.5 per cent of the maximum value of the signal over a distance of 0.030 inch.

It may be mentioned that the deviation from the straight line response that occurs just before the maximum is caused by the width of the light spot which was approximately 0.005 inch.

The slope of the response curve of Fig. 6 is 6 per cent of the maximum value of the signal per 0.001 inch. For

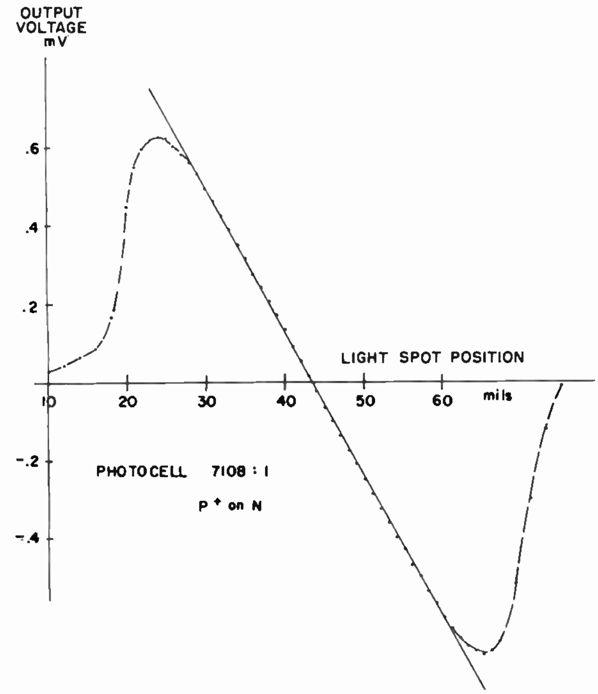

Fig. 15—Experimental response curve. Photocell p^+ on n.

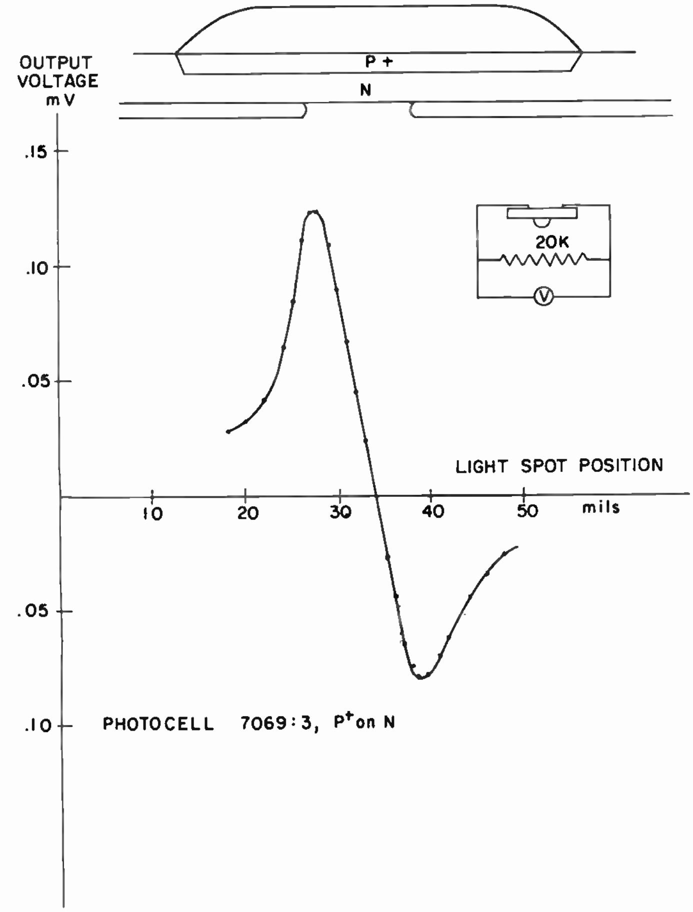

Fig. 17—Response curve with close interbase spacing.

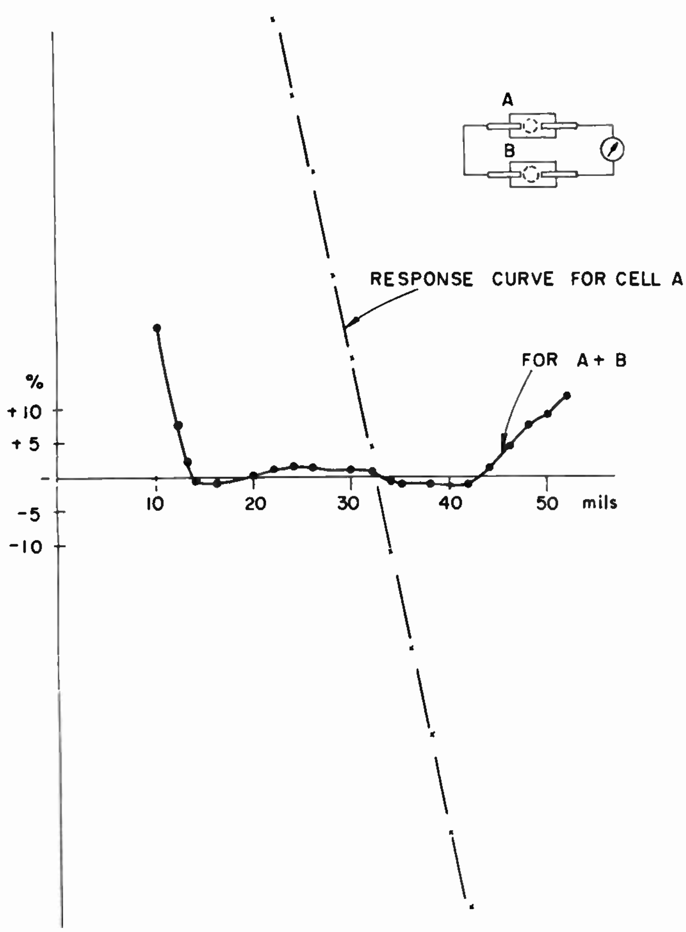

Fig. 16—Differential output from two opposing photocells.

many applications a larger slope is desirable and Fig. 17 shows a response curve for a cell whose base contacts are separated by only 0.012 inch. The slope for this cell is 20 per cent per 0.001 inch and still larger slopes are possible with smaller interbase distances. The slope also depends on the size of the light spot. While with an infinitely small light spot, the maxima would fall at the ends of the base contacts, a light spot of a width of half the interbase distance would move the maxima half way in towards the center and thereby increase the slope. On the other hand a spot size larger than the interbase distance distorts the response curve. With an optimum spot size a slope of 43 per cent of maximum value per 0.001 inch has been measured.

Spot Movement Sensitivity

The minimum amount of light spot movement that can be detected with the cell can be derived from the response curve and the noise level. Assume that the instrument used to measure the interbase current is no better than a conventional panel type ammeter. Then it is reasonable to assume that to be detectable the change in current has to be at least 3 per cent of the current that gives full scale deflection. Higher sensitivity is then obtained with operation near the crossover point than near the maxima. The limit for crossover point operation is set by the noise.

With a low noise amplifier it is possible to measure reliably down to 10 μv on a response curve having a maximum value of 1.5 mv, and with a slope of 43 per cent per 0.001 inch, 3 per cent of 10 μv with this slope corresponds to detection of a light spot movement of only 100 Å. The limiting factor in this measurement was 60 cps pickup so that still better accuracy is possible. The limit set by the noise of the cell is approximately 1/100 Å.

With the lens at a distance of 1 inch from the photocell, 100 Å corresponds to an angle of $4 \cdot 10^{-5\circ}$, or 0.1 second of arc. This compares favorably with the accuracy that can be obtained with the eye and a good optical range finder, amounting to 10–15 seconds of arc.[5]

In a range finder two lines are brought to coincidence, and this can be done with an accuracy that considerably exceeds the resolving power of the eye alone which is about $1\frac{1}{2}$ minutes of arc under favorable circumstances.[6]

Frequency Response

The injection of hole-electron pairs, and their subsequent diffusion to the alloyed (collector) dot is exactly analogous to conditions in a transistor, and similar arguments on n-region (base) width, built-in fields, etc., as for transistors should hold. The rest of the cycle—the redistribution and reemission and subsequent space charge neutralization—is majority carrier conduction and may be assumed to have negligible time delay. It may therefore be concluded that the frequency response of the photocell can be expected to be equivalent to that of a transistor with comparable base width, *i.e.*, with a cutoff frequency of approximately 1 mc with a base width of 0.001 inch. The feasibility of making thin base layers should be even better in the photocell than in the transistor since only one alloyed dot is necessary.

Sensitivity and Signal-to-Noise Ratio

Many potential applications involve the detection of distant or weak lights or, consequently, very small signals. The lower limit for the detection is set by the noise of the photocell. For any practical light level and with no voltage applied to the photocell, the dominating noise source is the thermal noise of the base resistance. The base resistance may be calculated from the dimensions and is approximately 100 ohms with the dimensions and materials used. Then, in the conventional notation,

$$\bar{e}^2 = 4kTRdf$$

where R is the base resistance as defined above.

[5] E. B. Brown, "Optical Instruments," Chemical Publishing Co., Inc., Brooklyn N. Y. p. 376; 1945.

For the low-frequency equivalent circuit in Fig. 18, the noise generator will generate a current

$$\bar{i}^2 = 4kT \frac{1}{R} dF$$

which at room temperature and a bandwidth of 1 cps is

$$i \approx 10^{-11} \text{ amp.}$$

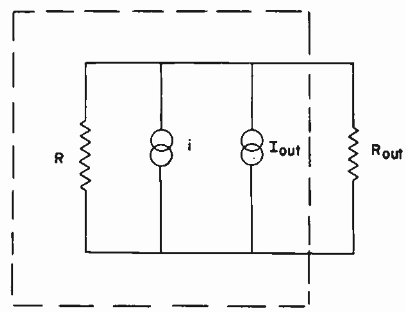

Fig. 18—Low-frequency equivalent circuit.

The signal intensity may be computed as follows: consider a light source emitting P watts of monochromatic light of wavelength, λ, at a distance x. The optical system in front of the photocell consists of a lens with an effective area, A, and with a transmission factor, M, including any protective enclosure around the cell. The reflection coefficient of the semiconductor surface may be taken as 0.37 (germanium). The collecting efficiency of either base tab in a symmetrical design is 0.5. The quantum efficiency is taken as 1. Assuming further that $w \ll D$ (one-dimensional diffusion) and

$$R_{out} = R\,(\text{match}) = \rho \frac{2l}{a}.$$

No loss of carriers through recombination during diffusion from injection point to the barrier is considered. Then the output current at the maximum response is

$$\begin{aligned} I_{out} &= \frac{\frac{2\rho}{a} I(l - x_1)}{\rho \frac{2l}{a}} \\ &= \frac{PAM(1 - 0.37)0.5 \cdot \lambda e}{4\pi x^2 hc} \left(1 - \frac{x_1}{l}\right) \\ &= 2 \cdot 10^4 \frac{PAM\lambda}{x^2} \end{aligned} \tag{14}$$

and the signal-to-noise ratio

$$\frac{S}{N} = \frac{2.7 \cdot 10^{15} PAM\lambda\sqrt{2R}}{x^2\sqrt{Tdf}} \left(1 - \frac{x_1}{l}\right). \tag{15}$$

Now let us consider a typical example and find the maximum allowable distance between light and photocell.

Assume

$P = 10$ watts
$A = 10$ cm^2
$M = 80$ per cent
$\lambda = 6000$ Å
$R = 100$ ohm ($\rho = 1$ ohm cm, $w = 0.003$ inch, $b = 0.040$ inch)
$T = 300$ K°
$df = 1$ cps
$S/N = 10$ (20 db).

For this example x is 1000 m.

Of practical ways to increase this distance a parabolic mirror behind the light source may be mentioned. This should give an improvement of 10^2 in distance. It may be worth noticing that the dimensions of the photocell are not critical and enter only through R in (15).

It may be noted here that the signal-to-noise ratio naturally goes to zero at the crossover point. This point can therefore be localized only by interpolating between two points, one on each side, where the signal-to-noise ratio is large enough to allow an accurate reading.

The data used above correspond to a sensitivity of the order of 200 μa/lumen after deduction of losses, which is a reasonable value for a germanium photocell.

ACKNOWLEDGMENT

The author has greatly benefited from discussions with H. Johnson and H. Kroemer on different aspects of the work and with W. H. Fonger on noise.

Design Considerations for Broad-Band Ferrite Coaxial Line Isolators*

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Summary—**An analysis of the microwave magnetic fields associated with a dominant TEM mode propagating in a coaxial line reveals only a linear sense of microwave *H*-vector polarization. As a consequence, this transmission line structure does not inherently lend itself for use with ferrites to obtain nonreciprocal propagation characteristics. A means for obtaining microwave *H*-vector circular polarization in coaxial line is described in this paper. This technique consists of partially filling a coaxial line cross section with a low-loss dielectric. This structure, in conjunction with certain broad-banding techniques, has been utilized in the development of an octave bandwidth coaxial line isolator. A treatment of the parameters, with associated experimental verification, is presented which affects the operation of the isolator. Also included in this paper are the design and experimental characteristics of this isolator. An anticipated mode configuration in the dielectric and ferrite loaded coaxial line is derived on the basis of the experimental results presented in this paper.**

INTRODUCTION

INITIAL WORK performed on nonreciprocal ferrite loaded waveguide structures utilized ferrites axially located in circular waveguide propagating the circularly polarized dominant TE_{11} mode.[1,2] Subsequently, published work demonstrated that a sense of microwave H-vector circular polarization exists in other microwave structures as rectangular waveguide[3] and a helical traveling-wave tube.[4] However, such a sense of circular polarization does not inherently exist in coaxial line in which microwave energy is propagating in the dominant TEM mode.

It is the purpose of this paper to describe some of the results of a method for generating a sense of microwave H-vector circular polarization in coaxial line. The circular polarization is achieved by a mode distortion technique which utilizes a coaxial line whose cross section is partially filled with a low-loss dielectric material. The quality of the generated sense of circular polarization, *i.e.*, the ellipticity of the microwave H-vector components, and the magnitude of the reverse and forward wave attenuation at ferrite gyromagnetic resonance are shown to be dependent on the dielectric constant (ϵ_m) of the mode distorting dielectric material. Also, such characteristics of the ferrite as saturation magnetization ($4\pi M_s$), dielectric constant (ϵ_f), etc., are observed to affect the above quantities.

The manner in which ϵ_m and certain ferrite parameters affect ferrite nonreciprocity in the resonance region is

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[1] C. L. Hogan, "The ferromagnetic Faraday effect at microwave frequencies and its applications," *Rev. Mod. Phys.*, vol. 25, pp. 253–263; January, 1953.

[2] H. N. Chait and N. G. Sakiotis, "Ferrites at microwaves," PROC. IRE, vol. 41, pp. 87–93; January, 1953.

[3] M. L. Kales, H. N. Chait, and N. G. Sakiotis, "A non-reciprocal microwave component," *J. Appl. Phys.*, vol. 24, pp. 816–817; June, 1953.

[4] J. S. Cook, R. Kompfner, and H. Suhl, "Nonreciprocal loss in traveling-wave tubes using ferrite attenuators," PROC. IRE, vol. 42, pp. 1188–1189; July, 1955.

treated in detail. Design considerations for the development of an extremely broad-band coaxial line ferrite isolator are discussed. Finally, an operative coaxial line isolator is described and data are presented on its attenuation and vswr characteristics.

Experimental

The measurement system used in this investigation permitted the measurement of ferrite element absorption and reflection losses, and the measurement of these same type losses which arise owing to the presence of the mode distorting dielectric material. The system shown in Fig. 1 is similar to those previously reported[2,5] with the exception that operation is affected in $\frac{7}{8}$ inch coaxial line with propagation in the dominant TEM mode. A transversely applied dc magnetic field is used to bias the ferrite to gyromagnetic resonance.

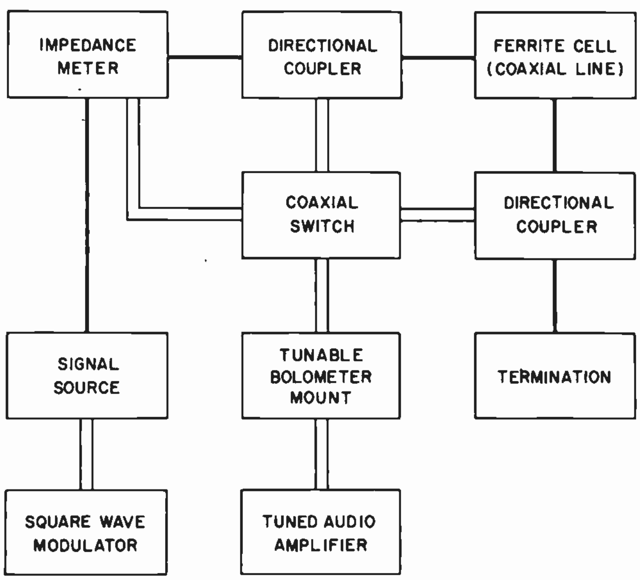


Fig. 1—Microwave system used for evaluation of ferrite coaxial line characteristics.

The system reliability is such that attenuation measurements are accurate to within ± 0.1 db. The magnetic field measurements are accurate to within ± 10 oersteds, as determined using a paramagnetic resonance fluxmeter developed at Sperry. Also, the field intensity along the length of the ferrites is uniform to within ± 1 per cent.

Theoretical

Circular polarization of the H-vector associated with microwave propagation has been found to exist in several types of microwave transmission lines. Included among these structures are circular waveguide propagating either the linearly or circularly polarized TE_{11} mode,[6] rectangular waveguide propagating the TE_{10} mode,[3] and a helical transmission line.[4] However, such is not the case for TEM mode propagation in coaxial line.

It can be shown that no sense of microwave H-vector circular polarization inherently exists in coaxial line for dominant mode propagation. This can be seen quite readily from an examination of the fields which exist in coaxial line. The microwave H-vector components are given by Moreno[7] as:

$$H_r = 0$$

$$H_\theta = \frac{I_0}{2\pi r}\left(e^{j\omega t - \gamma z} - \rho e^{j\omega t + \gamma z}\right)$$

$$H_z = 0. \tag{1}$$

In these equations I_0 is the conduction current amplitude along the inner conductor associated with the incident wave, and r is an arbitrary radius. The term ρ indicates the reflection coefficient, γ the propagation constant, μ and ϵ the permeability and dielectric constant of the dielectric medium respectively. As is shown, only one component of microwave H-vector (H_θ) exists. Hence, the requirements for circular polarization that equal amplitude components exist at a single point with a phase quadrature relationship are obviously not satisfied.

Since the basic requirement is not fulfilled for obtaining nonreciprocity in the ferrite loaded transmission line, *i.e.*, the ferrite does not see a sense of H-vector circular polarization at any position in unperturbed coaxial line, it is apparent that only reciprocal effects can be obtained. Thus, if nonreciprocal effects are to be achieved some means for introducing circular polarization in the coaxial line must be devised.

It is desirable in the case of coaxial line to obtain a sense of circular polarization either along the longitudinal (Z-) or transverse (r-) direction with respect to the direction of propagation. In order to generate a Z-sense of circular polarization it is required that an H_r of amplitude equal to H_θ be produced at a given point, and that a phase quadrature relationship between the two components be generated at the given point. Similarly, the generation of an r-sense of circular polarization requires that an H_z of amplitude equal to H_θ, and in phase quadrature with H_θ, be produced at a particular point. In this paper a technique for generating an r-sense of circular polarization only will be considered.

A technique which appeared feasible for achieving the required Z-component of the microwave H-vector con-

[5] B. J. Duncan and L. Swern, "Temperature behavior of ferrimagnetic resonance in ferrites located in waveguide," *J. Appl. Phys.*, vol. 27, pp. 209–215; March, 1956.

[6] A. G. Fox, S. E. Miller, and M. T. Weiss, "Behavior and applications of ferrites in the microwave region," *Bell Sys. Tech. J.*, vol. 34, pp. 5–103; January, 1955.

[7] T. Moreno, "Microwave Transmission Design Data," McGraw-Hill Book Co., Inc., New York, N. Y., p. 66; 1948.

sists of partially filling the cross section of the coaxial line with a low-loss dielectric material. Under these conditions the wave tends to propagate with different phase velocities in the two dielectric media. Thus, a Z-component of microwave magnetic field which is not in phase with the θ-component should be produced. By proper selection and shaping of dielectric materials it appeared feasible that an almost true sense of circular polarization could be produced.

Starting with the above considerations an extensive experimental program was conducted using various ferrites and low-loss dielectrics with dielectric constants covering a wide range. A portion of the results obtained are reported and discussed in the following section. Using the experimentally derived results a probable mode configuration has been derived for the partially dielectric filled coaxial line. Finally an octave bandwidth coaxial line isolator was developed which uses commercially available ferrites to obtain isolation over the 2.0 to 4.0 kmc frequency range.

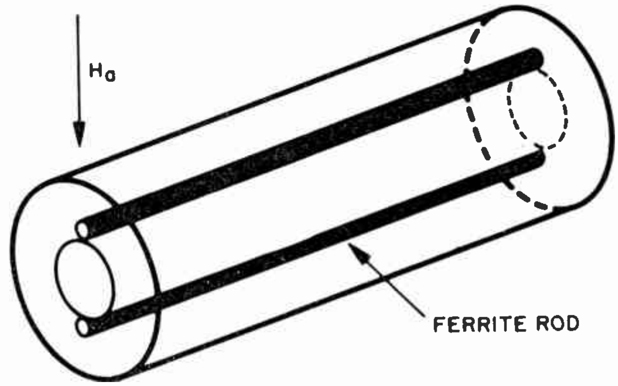
Experimental Results and Discussions

In order to demonstrate the origin of nonreciprocal effects in ferrite loaded coaxial line experimental results are presented on three coaxial line configurations shown in Fig. 2. As anticipated no sense of microwave H-vector circular polarization was found to exist in air-filled coaxial line propagating the dominant TEM mode. On the other hand circular polarization has indeed been found to exist in the two coaxial structures shown in Figs. 2(b) and 2(c). Each of the three cases shown in Fig. 2 will be treated in this paper. However, primary interest will be centered on the partially dielectric filled coaxial line configuration shown in Fig. 2(b).

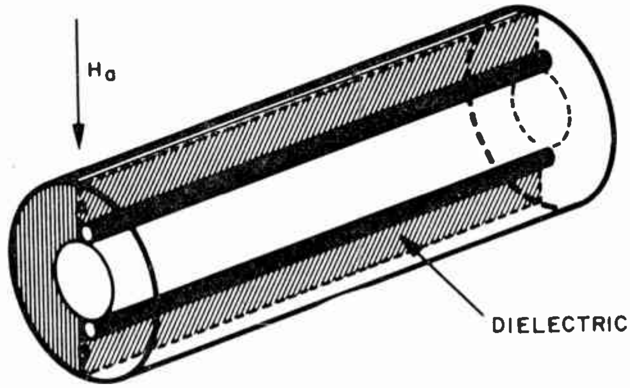
In the configuration shown in Fig. 2(b) the dielectric was made four-inches long for all studies with three-inch impedance matching sections on each end. The matching sections consisted of dielectric transitions tapering smoothly from the outer to inner coaxial line conductor. The ferrites used for all studies were nickel-zinc compositions which exhibit moderate values of saturation magnetization and dielectric constant.

In the first of these three cases, two transversely magnetized ferrites were located on opposite sides of the coaxial line center conductor [Fig. 2(a)]. Their θ-location was such that both ferrites were in a plane parallel to the direction of the magnetic biasing field. The results obtained on two 0.060-inch diameter by 3.0-inch length rods of a nickel-zinc ferrite are shown in Fig. 3 (next page). As indicated, there exists absolutely no nonreciprocity of the ferrite attenuation characteristics.

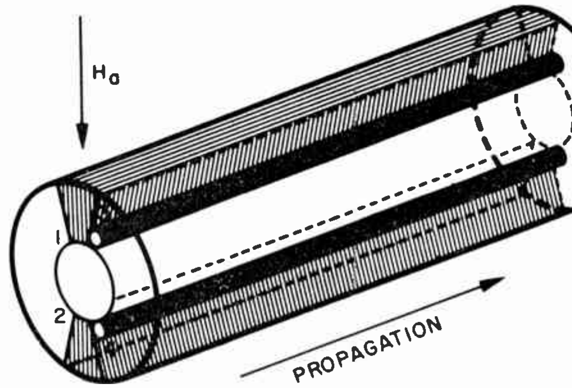
Next, the coaxial line cross section was one-half filled with a low loss-high dielectric constant-dielectric material [Fig. 2(b)]. The same two 0.060-inch diameter by 3-inch long ferrite rods were located against both the coaxial line center conductor, and the mode distorting



(a)



(b)



(c)

Fig. 2—Ferrite and dielectric configurations used for studies of ferrite nonreciprocity in coaxial line. (a) Ferrite coaxial line reciprocal structure. (b) Ferrite coaxial line nonreciprocal structure one sense of c.p. (c) Ferrite coaxial line nonreciprocal structure two senses of c.p.

dielectric material. The magnetic biasing field was applied parallel to the exposed face of the dielectric as shown in Fig. 2(b). The attenuation characteristics of this configuration are also recorded in Fig. 3. As noted, a high degree of nonreciprocity of the ferrite attenuation characteristics is obtainable at ferrite gyromagnetic resonance.

An investigation of the ellipticity of the microwave H-vector at various points in the dielectric loaded coaxial line was conducted using small ferrite rods. The

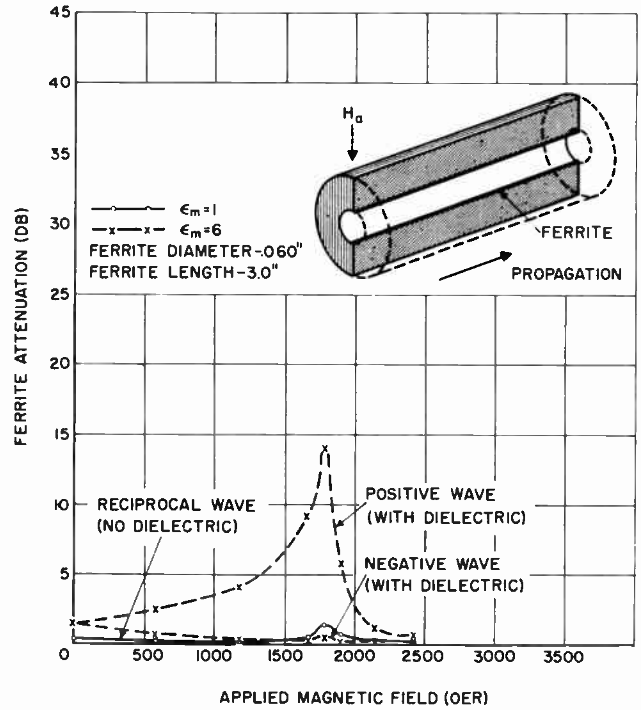


Fig. 3—Experimental verification of the existence of microwave H-vector circular polarization in coaxial line.

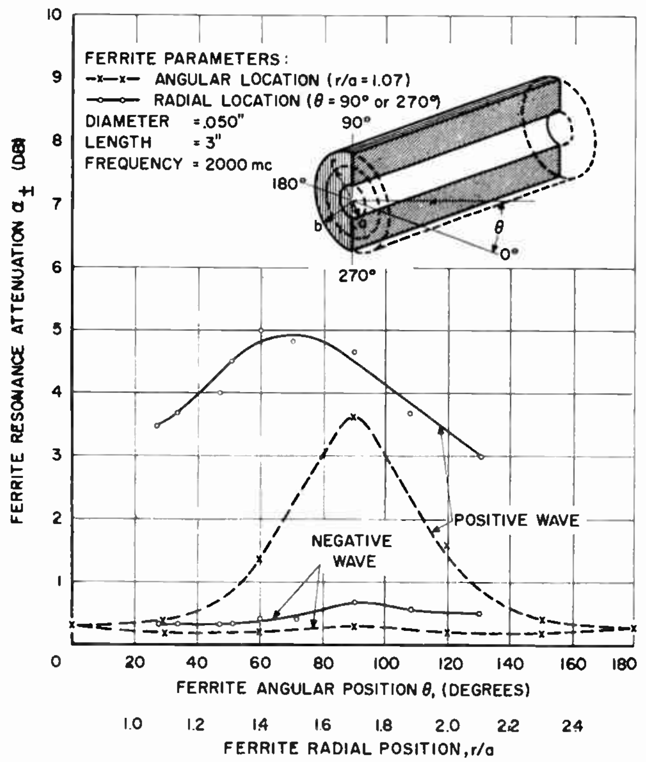


Fig. 4—Experimental data indicating the location of optimum microwave H-vector circular polarization in dielectric loaded coaxial line.

diameter of the rods was maintained extremely small so as to cause a minimum of field perturbation. These tests were performed in an effort to determine the position of minimum microwave H-vector ellipticity and hence the optimum degree of circular polarization. It was determined (Fig. 4) that the largest degree of circular polarization is obtained at the dielectric surfaces [Points A and B of Fig. 2(b)]. Furthermore, it was observed that the same senses of circular polarization exist at both A and B. However, it is significant to note that the maximum reverse loss was obtained at a position away from the center conductor and on the dielectric surface (Fig. 4).

In addition to the changes in attenuation with ferrite position both the ferromagnetic resonance linewidth (Δx) at the half-power points and the applied magnetic field required for resonance (H_r) were observed to change slightly as the ferrite was moved from the coaxial line inner to outer conductor along the dielectric surface. However, these changes are very small and can, in general, be considered negligible.

In addition to the above changes these same ferrite quantities were observed to be dependent on the ferrite θ-location (Fig. 4), and the location of the ferrite along the length of the dielectric. However, for dielectric lengths greater than a certain value, which is frequency dependent, the changes are small with location along the length of the dielectric. Also, except for changes in attenuation, the changes with θ location can generally be disregarded. The information on the dependence of attenuation on location will be used later in the tentative determination of the microwave magnetic field configuration in the dielectric loaded section of coaxial line.

The quality of the generated microwave H-vector circular polarization at the ferrites, and the magnitude of nonreciprocity of ferrite attenuation at resonance H_r, are dependent on several other quantities in addition to location. These include dielectric constant of the distorting medium ϵ_m, the ferrite scalar permeabilities $\mu_\pm$ to positive and negative circularly polarized waves (henceforth referred to as positive[8] and negative waves), ferrite dielectric constant ϵ_f, and the size and shape of the ferrite material. An attempt to determine the effects of these parameters theoretically presents an extremely difficult problem. However, a large portion of this information can be determined experimentally. The manner in which ferrite diameter (d_f) and ϵ_m affect the attenuation characteristics of the nickel-zinc ferrite at gyromagnetic resonance is shown in Figs. 5 and 6 for the configuration displayed in Fig. 2(b). These tests were performed in $\frac{7}{8}$-inch coaxial line at 2000 mc using high dielectric constant materials. As indicated, for ϵ_m equal to a constant, ferrite positive wave attenuation (α_+) increases with increasing d_f, while ferrite negative wave attenuation (α_-) remains small. Also a splitting of the resonance line was observed to occur as ferrite rod diameter is increased. These behaviors are similar to those observed in rectangular and circular waveguide when oversized ferrite samples at resonance are located

[8] The notation used here is the one where the positive wave is the wave which is rotating in the direction of the positive electric current which generates the dc magnetic biasing field.

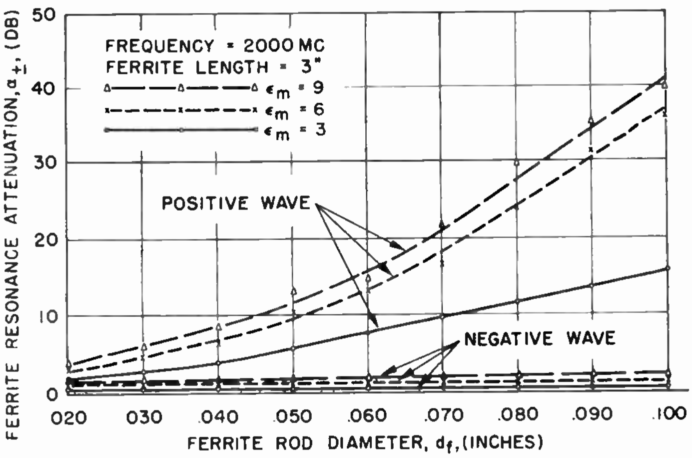


Fig. 5—Dependence of resonance attenuation characteristics on ferrite rod diameter for a nickel-zinc ferrite, with ϵ_m as parameter.

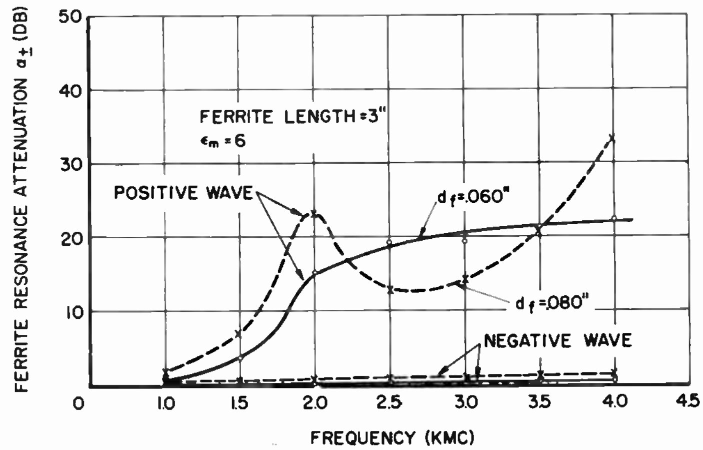


Fig. 7—Dependence of resonance attenuation characteristics on frequency for a nickel-zinc ferrite with ferrite rod diameter as parameter.

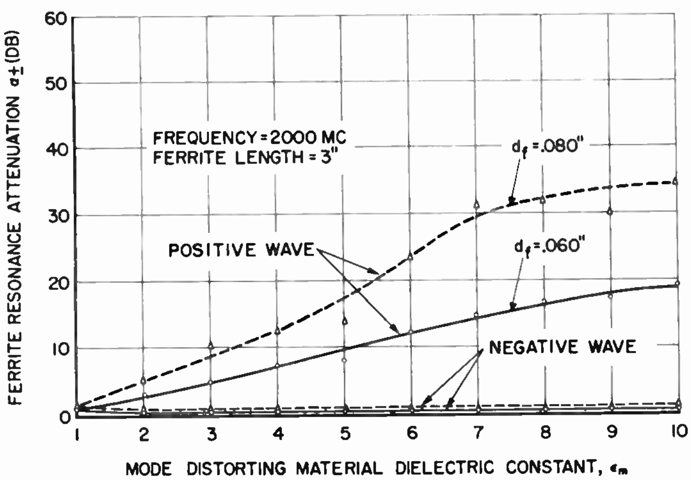


Fig. 6—Dependence of resonance attenuation characteristics on ϵ_m for a nickel-zinc ferrite with ferrite rod diameter as parameter.

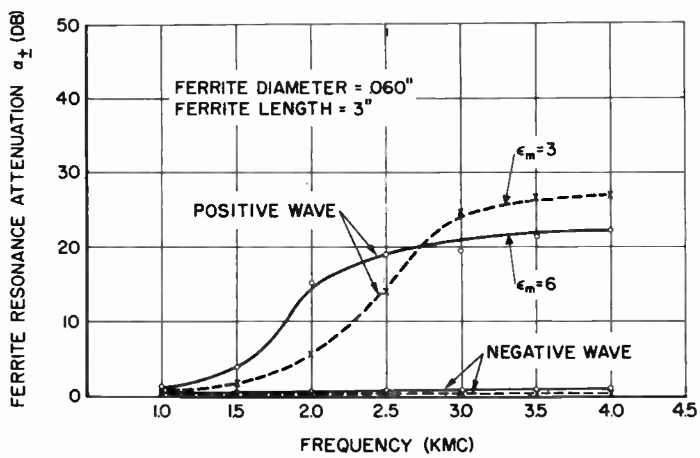


Fig. 8—Dependence of resonance attenuation characteristics on frequency for a nickel-zinc ferrite with ϵ_m as parameter.

in the waveguide structure; this has been attributed to a moding condition set up in the region of the ferrite when operation is near resonance.

An additional important effect is the increase in α_+ with increasing ϵ_m, with α_- at the positive wave resonance field remaining small. This can probably be attributed to a combination of dielectric loading,[9] and an improvement in the quality of the generated microwave H-vector circular polarization. However, a maximum α_+ is finally reached for ϵ_m greater than 10 and then decreases with increasing ϵ_m, while α_- increases. Also the resonance line splitting becomes more significant for high values of ϵ_m. This is again probably due to the simultaneous existence of one or more undesirable higher order modes in the vicinity of the ferrite.

The effects of ϵ_m and d_f on the frequency behavior of α_+ and α_- are depicted in Figs. 7 and 8. As might be anticipated, α_+ increased with frequency without an appreciable increase in α_- for the smaller values of d_f and ϵ_m when occurring simultaneously. However, when the values of d_f and ϵ_m were both increased α_+ first increased with increasing frequency and then eventually began to decrease; an appreciable increase in α_- was observed to occur in the frequency region of decreasing α_+. The frequency at which the peak of α_+ occurs is dependent on the values of d_f and ϵ_m used in the coaxial structure. This observed decrease in α_+, and increase in α_-, can most probably be attributed to the presence of some higher order mode propagating independently of the distorted TEM coaxial line mode.

The manner in which ϵ_m and d_f affect H_r and Δx is depicted in Figs. 9 and 10. It is very interesting to note the decrease in H_r both with increasing ϵ_m and increasing d_f, which is the inverse of the behavior of H_r with ferrites in rectangular waveguide.[10] The decrease in H_r with both d_f and ϵ_m in coaxial line is not presently understood by the authors. However, these effects can probably be

[9] M. T. Weiss, "Improved Rectangular Waveguide Resonance Isolators," presented at National Symposium on Microwave Techniques, University of Pennsylvania, Pittsburgh, Pa; February 2–3, 1956.

[10] Unpublished experimental results of work performed by the authors at Sperry Gyroscope Company.

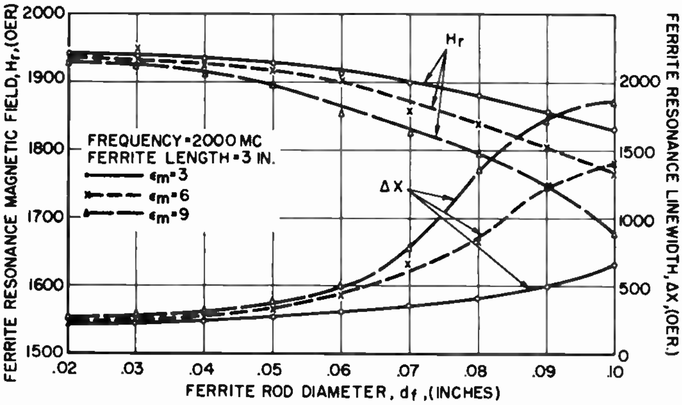


Fig. 9—Dependence of resonance field and linewidth on ferrite rod diameter for a nickel-zinc ferrite with ϵ_m as parameter.

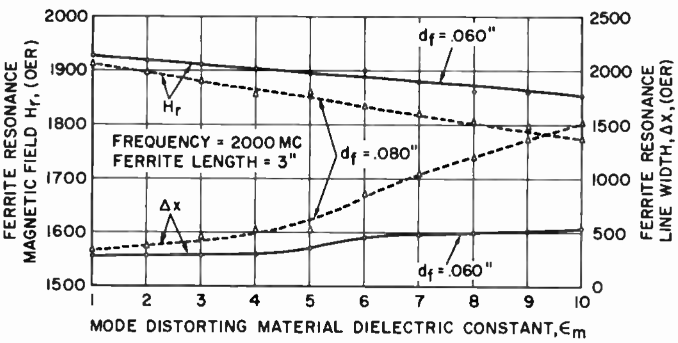


Fig. 10—Dependence of resonance field and linewidth on ϵ_m for a nickel-zinc ferrite, with ferrite rod diameter as parameter.

accounted for by a change in the effective ferrite demagnetizing factors with increasing d_f and ϵ_m.

An increase in Δx was observed to occur for increasing values of d_f and ϵ_m. As noted in Figs. 9 and 10 the rate of increase in Δx is largest for large values of d_f and ϵ_m. These results are as might be expected, particularly the increase in Δx with d_f. Similar increases have been observed with ferrites and dielectric material in rectangular waveguide.[11] They have generally been attributed to changes in effective ferrite demagnetizing factors.

The behavior with frequency of H_r and Δx with varying values of ϵ_m and d_f is demonstrated in Figs. 11 and 12. As anticipated, the applied magnetic field required to produce gyromagnetic resonance increases with increasing frequency for all values of d_f and ϵ_m tested. The increase in H_r probably not only reflects the inherent frequency sensitivity of the ferrite resonance but includes as well certain dielectric loading effects due to the presence of the mode distorting dielectric materials. The effect of frequency on the field required for resonance according to Kittel's equation is included for comparison purposes. This plot was obtained assuming that the ferrite is completely saturated, and that d_f is much less than a wavelength in the ferrite medium.

[11] *Ibid.*

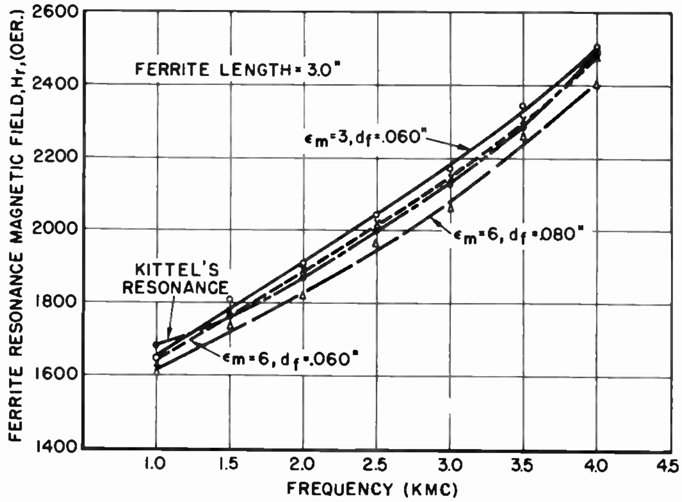


Fig. 11—Dependence of ferrite resonance field on frequency for a nickel-zinc ferrite with d_f and ϵ_m as parameters.

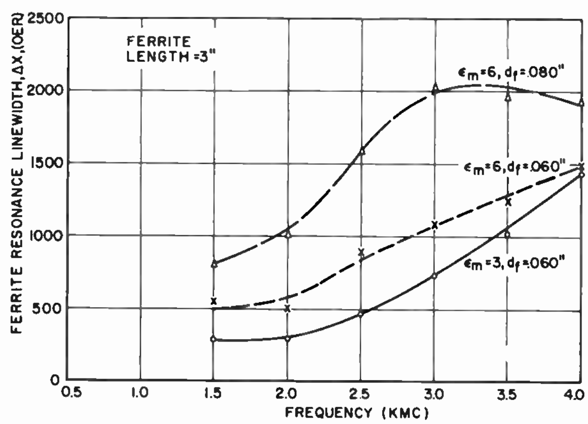
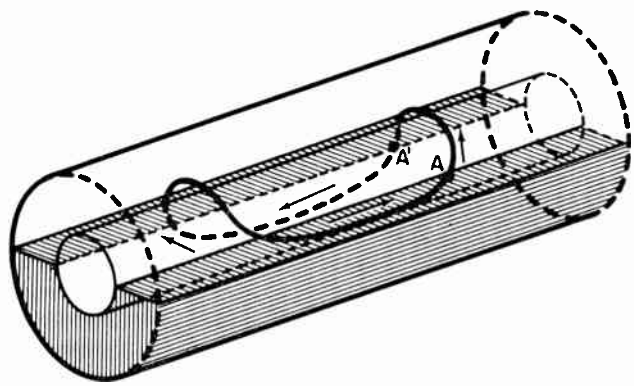


Fig. 12—Dependence of ferrite resonance linewidth on frequency for a nickel-zinc ferrite with d_f and ϵ_m as parameters.

Similarly, the increase in Δx with frequency is probably due to both a change in the effective ferrite demagnetizing factors with frequency across the ferrite cross section, and to complicated dielectric loading effects.

From the experimental results obtained on the structure depicted in Fig. 2(b) certain conclusions regarding a probable microwave magnetic field configuration can be reached. In particular, microwave H-vector circular polarization was found to exist on opposite sides of the coaxial line center conductor, and on the dielectric surfaces (Fig. 3). The quality of the circular polarization, as measured by the ellipticity of the microwave H-vector components, was observed to decrease with a displacement from the dielectric surface. Furthermore, the sense of circular polarization was found to be the same on opposite sides of the coaxial line center conductor, and along the dielectric surfaces (Fig. 4). On the basis of these results, one might conclude that the microwave magnetic field configuration is as shown in Fig. 13. Such a mode configuration would give results compatible



H-VECTOR AT POSITION A		H-VECTOR AT POSITION A'	
$t = t_0$	↑	$t = t_0$	↓
$t = t_0 + T/4$	→	$t = t_0 + T/4$	←
$t = t_0 + T/2$	↓	$t = t_0 + T/2$	↑
OBSERVER AT A OR A' SEES H-VECTOR POLARIZATION OF SAME SENSE NORMAL TO CENTER CONDUCTOR AND PARALLEL TO DIELECTRIC FACE			

Fig. 13—Probable microwave H-vector field configuration in a dielectric loaded section of coaxial line for incident TEM mode.

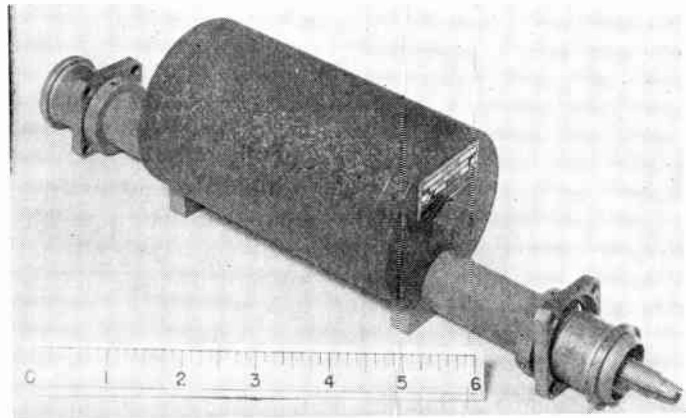


Fig. 14—An assembled octave bandwidth S-band coaxial line isolator.

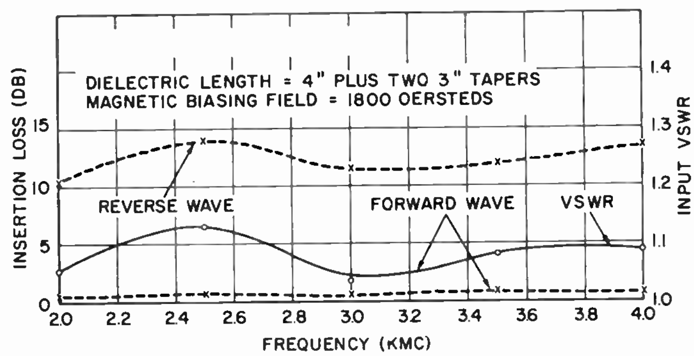


Fig. 15—Attenuation and vswr characteristics of an octave bandwidth coaxial line isolator.

with those derived experimentally and reported in this paper.

While no experimental data will be reported on the structure shown in Fig. 2(c) a brief discussion of the characteristics of this configuration is warranted. It has the characteristic that opposite senses of circular polarization can be made to exist for only one direction of propagation.

In the configuration of Fig. 2(c) the sense of circular polarization on surface (1) is identical to that on surface (2). Similarly, the sense of circular polarization at surface (3) is the same as that at surface (4). However, the sense of circular polarization at surfaces (1) and (2) is opposite to that at surfaces (3) and (4). Surfaces (1) and (2) may be compared electrically to one plane of circular polarization in rectangular waveguide whereas surfaces (3) and (4) could be compared to the opposite plane of circular polarization. If the arc subtended by the dielectric is small in each case then a single applied magnetic field can be used to bias ferrites located at each of the surfaces.

On the basis of the information reported thus far in this paper, together with the fact that operation is in coaxial line, the development of extremely broad-band nonreciprocal coaxial line components appeared feasible. Subsequently, the development of an experimental model of an octave bandwidth-coaxial line isolator was initiated using commercially available ferrites, with various values of saturation magnetization. The S-band offered a compromise region of operation in that ferrite differential propagation characteristics with small magnetic losses are still quite large in this region, and coaxial line structure losses can be maintained small without the introduction of separate undesirable higher order modes.

The configuration of the S-band coaxial line isolator is shown in Fig. 14. The over-all length of the device is only nine inches and its maximum cross sectional area at any cross section is three inches by three inches. These dimensions can probably be decreased appreciably through additional development.

The electrical characteristics of this component are depicted in Fig. 15. The reverse wave attenuation is observed to be greater than 10.5 db over the 2–4 kmc frequency range, and the forward loss is less than 0.8 db. The vswr over this same band is less than 1.15. Once again these characteristics can be improved with development.

It appears that the main source of forward wave attenuation in this isolator is the dielectric loss associated with both the ferrite and mode distorting dielectric. The use of low dielectric loss ferrites would considerably decrease the forward wave loss.

It should be noted that the use of ferrites with high saturation magnetizations in the isolator permits the construction of components of a minimum length. This is owing to the greater microwave activity per unit length of ferrite.

Also, the high saturation magnetization ferrites require larger magnetic biasing fields for resonance in the configuration shown. This aids to some extent in a reduction in forward wave loss in that low field losses in high saturation magnetization ferrites are generally reduced to a smaller value of resonance than in low saturation magnetization ferrites.

Conclusion

An air-filled coaxial line propagating the dominant TEM mode exhibits no sense of microwave H-vector circular polarization. Hence, it does not inherently lend itself to the use of ferrites to obtain nonreciprocal propagation characteristics. However, by the use of a mode distorting dielectric material a radially directed sense of microwave H-vector circular polarization can be generated in coaxial line. By proper selection of the dielectric structures either a single sense of circular polarization, or both senses of circular polarization, can be obtained for one direction of propagation.

Several important parameter effects are evident in a dielectric and ferrite loaded coaxial structure. Resonance attenuation, field, and linewidth are dependent on both the dielectric constant ϵ_m and the ferrite rod diameter d_f. The behavior of resonance field with mode distorting material ϵ_m is opposite to that of partially dielectric filled rectangular waveguide.[12]

Using dielectric loaded coaxial line it is feasible to develop nonreciprocal ferrite devices. In particular, an octave bandwidth S-band coaxial line isolator has been developed using commercial ferrites and a permanent magnetic biasing field. It is believed that even greater bandwidths can be obtained in the present component and that lower frequency components are feasible in the very near future.

Acknowledgment

The authors would like to acknowledge their indebtedness to Dr. R. E. Henning for his contribution to this work. Also, gratitude is expressed to Pasquale Iassogna and Alfred Guthenberg for assistance in obtaining the experimental data reported herein.

[12] Unpublished results obtained at Sperry, *loc. cit.*

Correction

Herbert P. Raabe, author of the paper "Measurement of Instantaneous Frequency with a Microwave Interferometer," which appeared on pages 30–38 of the January, 1957 issue of Proceedings, has brought the following error to the attention of the editors.

Eq. (17), on page 33 should be

$$U_s = U_{s0} \cos \frac{\psi d}{2} \cdot$$

Analysis of Nonreciprocal Effects in an N-Wire, Ferrite-Loaded Transmission Line*

H. BOYET† AND H. SEIDEL†, MEMBER, IRE

Summary—The N-wire rotationally symmetric transmission line surrounding a thin ferrite rod is analyzed here with the aim of determining the feasibility of this low-frequency nonreciprocal device. The structure is assumed to be infinitely long. The transmission line equations are solved in which the effect of mutual inductances and capacitances between wires and the perturbing effect of the ferrite are taken into account. An expression for the rotation per unit length of line is obtained as a function of the structure geometry, the magnetic condition of the ferrite, and the operating frequency. The result is valid for thin wires and thin ferrite rods. Two examples are given. The four-wire line (gyrator) is evaluated for a ferrite radius equal to 1/16 the structure radius and the result is a predicted rotation of 0.76°/cm at 1000 mc for a 500 gauss magnetization ferrite. Somewhat larger rotations than this were obtained experimentally by Rowen who worked with larger diameter ferrites where the perturbation theory developed here would not be expected to apply. The eight-wire line (circulator) is also evaluated on the basis of the theory and the analysis indicates that rotations of the order of 4°/cm should be obtained at 1000 mc for a 500 gauss ferrite at similar operating conditions.

Introduction

Devices based on the nonreciprocal properties of microwave propagation in ferrite-loaded waveguides have been discussed extensively in the literature.[1-9] Some of the new and useful devices which have been developed are the one-way transmission line (isolator),[1,3,7] the circulator for channeling energy,[3] nonreciprocal phase shifters,[3] ferrite tuned cavities,[8] ferrite directional couplers.[9] etc. The list of references below is simply representative and not in any way inclusive.

For example, nonreciprocal Faraday rotation of a linearly polarized rf wave may occur in a magnetized ferrite medium. If the rotation is adjusted to 45° and if properly oriented resistance vanes are employed, isolation may be achieved in a circular waveguide. Circulator action may be obtained if rectangular-to-round transitions and cross-polarization pick-offs are used with the circular guide. Gyrator action may be obtained in any ferrite loaded waveguide capable of giving a phase delay of $\pi+\theta$ for one direction of propagation and a phase delay of θ for the reverse direction and may be used to derive various nonreciprocal elements. These nonreciprocal waveguide devices, and others, are finding extensive use in microwave circuits and relay systems.

It is naturally interesting to inquire whether appreciable nonreciprocal ferrite effects can be attained with a multiple wire transmission line in the low frequency region (say 1 or 2 kmc and lower) where waveguides could not be used because of their impractical size.

Recent work by Suhl[10] of Bell Telephone Laboratories on single crystals of nickel ferrite has indicated that ferromagnetic resonance can be attained in the 1000–2000 mc range. Low saturation polycrystalline ferrite materials have also been developed[11,12] in which gyromagnetic resonance is observed at as low as 160 mc. These developments suggest the feasibility of constructing nonreciprocal ferrite circuit elements operating at low frequencies.

J. H. Rowen and A. M. Clogston of Bell Telephone Laboratories have recently suggested[13] compact TEM-like structures leading to low-frequency gyrators and circulators. The structures are comprised of four-wire and eight-wire transmission lines, respectively, either embedded in, or surrounding, a longitudinally magnetized ferrite rod. The difference in phase velocities of the modes can be adjusted, as desired, to effect total transfer of energy from one pair of wires to other pairs in either the four-wire or eight-wire cases. We will find that the solution for the characteristic modes of the N-wire rotationally symmetric system are such that for a given characteristic mode each wire is indistinguishable from its neighbors in amplitude of excitation but may differ uniformly in phase. The modes are therefore circularly polarized and, because of symmetry, they exist in both polarization senses. A linear polarization may be constructed as the sum of these two circular modes. The linear polarization will have a sinusoidal distribution of amplitude over the N wires. The polarization axis is defined as that axis passing through conductors having maximum amplitude of excitation. The system is de-

* Original manuscript received by the IRE, September 14, 1956; revised manuscript received, December 27, 1956.

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[1] C. L. Hogan, "The ferromagnetic Faraday effect at microwave frequencies and its applications," *Rev. Mod. Phys.*, vol. 25, p. 253–262; January, 1953.

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[3] A. G. Fox, S. E. Miller, and M. T. Weiss, "Behavior and applications of ferrites in the microwave region," *Bell. Syst. Tech. J.*, vol. 34, pp. 5–103; January, 1955.

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[5] B. Lax, K. J. Button, and L. M. Roth, "Ferrite phase shifters in rectangular waveguide," *J. Appl. Phys.*, vol. 25, pp. 1413–1419; November, 1954.

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[7] S. Weisbaum and H. Seidel, "The field displacement isolator," *Bell Syst. Tech. J.*, vol. 35, pp. 877–898; July, 1956.

[8] C. E. Fay, "Ferrite-tuned resonant cavities," Proc. IRE, vol. 44, pp. 1446–1449; October, 1956.

[9] A. D. Berk and E. Strumwasser, "Ferrite directional couplers," Proc. IRE, vol. 44, pp. 1439–1445; October, 1956.

[10] H. Suhl, *Phys. Rev.*, vol. 97, p. 555; 1955.

[11] H. Suhl, L. G. Van Uitert, and J. L. Davis, "Ferromagnetic resonance in magnesium-manganese aluminum ferrite between 160 and 1900 mc," *J. Appl. Phys.*, vol. 26, pp. 1180–1182; September, 1955.

[12] L. G. Van Uitert, *J. Appl. Phys.* vol. 26 pp. 1289–1290; 1955.

[13] Private communication.

generate in that an orthogonal axis may also be constructed for which the system is identically excited. This leads to two linear polarizations which, in any given transverse plane, maintain their orthogonality with respect to one another. These two linear polarizations provide the basis for Faraday rotation in exactly similar manner to that in waveguide ferrite devices.[3] The discrete wire system may, therefore, lead to gyrator or circulator properties as in the waveguide system.[3] In particular, the ferrite conditions may be adjusted in the four-wire rotationally symmetric line to produce 90° rotation, in which case we have the gyrator. For the eight-wire rotationally symmetric line, 45° rotation leads to circular action.[3]

The unloaded transmission line propagates TEM waves. It would be expected that any longitudinal components in the ferrite-loaded line will die off with decreasing frequency so that no cutoff frequency problem exists here either. The cross sectional area of the wire device can, therefore, be kept small in the low-frequency region in contrast to the impractically large cross sections associated with conventional waveguides below a few thousand mc.

In this paper we analyze, from a transmission line viewpoint, the N-wire rotationally symmetric structure surrounding a longitudinally magnetized thin pencil of ferrite. The structure is assumed to be infinite in length. The mutual capacitances and inductances inherent in the system, and the perturbing effect of the ferrite, are taken into account in analyzing the propagation properties of the structure. It is assumed that the ferrite diameter is small compared with the structure diameter, and that each wire diameter is small compared with the distance between nearest wire centers. When the appropriate transmission line equations are solved, an expression for the propagation constants of the modes that couple to the ferrite is obtained. This leads to an expression for the rotation of the polarization (per unit length of structure) in terms of the structure geometry, ferrite saturation magnetization, applied dc magnetic field, and operating frequency.

One result of the analysis is the existence of some modes which do not couple, in first order, to thin ferrite rods and propagate without rotation. There are always only two linearly independent modes that couple to the ferrite and undergo rotation in the N-wire structure.

Numerical results based on the analysis are presented in a later section for rotations in typical four-wire and eight-wire structures.

Analysis

Magnetic Field and Magnetic Induction at Center of Ferrite Rod

The geometry of the N-wire structure is shown in Fig. 1. The number of wires, N, may be even or odd but the structure possesses rotational symmetry. We are assuming the structure to have infinite length. No account is taken of reflections.

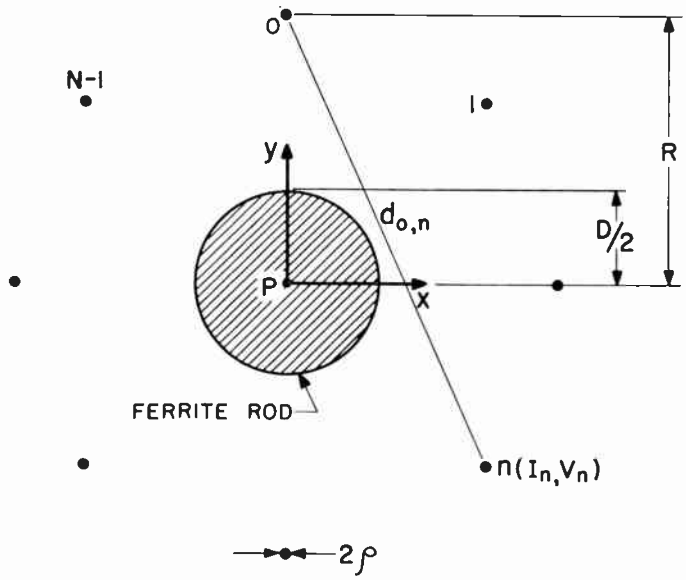

Fig. 1—Voltages and currents on N-wire rotationally symmetric structure. Ferrite rod is located at center of structure and $D \ll 2R$. Ferrite is magnetized longitudinally.

The list of symbols used is given below:

I_n, V_n—current and voltage on nth wire
N—total number of wires
R—radius of structure
D—diameter of ferrite rod
ρ—radius of wire
$d_{o,n}$—distance between centers of wires o and n
$\hat{x}$, $\hat{y}$—unit vectors in x and y directions (transverse plane)
z—longitudinal coordinate
P—center of structure
ω—angular frequency of wave
H_p—rf magnetic field at P in absence of ferrite
H_p^f—rf magnetic field at P with ferrite present
B_p^f—rf magnetic induction at P with ferrite present
m—rf magnetization inside ferrite
H_o—longitudinally applied dc magnetic field
M_s—saturation magnetization of ferrite
γ—gyromagnetic ratio of the electron ($2\pi \times 2.8 \times 10^6$ radians/oersted sec)
μ, k—diagonal and off-diagonal components of Polder tensor T
μ_o—permeability of free space (1.257×10^{-6} henry/meter)
ϵ_o—dielectric constant of free space (8.854×10^{-12} farads/meter)
c—speed of light in free space, $1/\sqrt{\mu_o \epsilon_o}$ (3×10^8 meters/sec)
L_{ni}, C_{ni}—mutual inductances and capacitances between wires n and i (per unit length of line)
β—propagation constant on line
Z_o—characteristic impedance of line
σ—rotation of polarization per unit length of line.

A voltage and current on one set of wires produces a magnetic field inside the ferrite which in turn sets up a component of magnetic induction at right angles to the magnetic field. This induces voltage and current on the other wires.

Because of the rotational symmetry of the structure we may write the voltage and current on the nth wire in the absence of ferrite in the form

$$\begin{aligned} V_n &= V_o e^{j(\omega t + 2\pi r n/N)} \\ I_n &= I_o e^{j(\omega t + 2\pi r n/N)} \end{aligned} \tag{1}$$

where V_o and I_o are current and voltage on the zeroth wire and are functions of z only. The number r indicates the possibility of different propagating TEM modes with $r=0, \pm 1, \pm 2, \cdots$. In (3), (11), and (12) we shall see that the magnetic field and induction are rotating circular polarizations at P and, therefore, voltages are induced which are consistent with the forms in (1), as expected.

The rf magnetic field at P in the absence of ferrite is simply the sum of contributions of fields from all wires and is given by (mks units)

$$\vec{H}_p = \frac{I_o e^{j\omega t}}{4\pi R}\left[\frac{e^{j2\pi(r+1)}-1}{e^{j(2\pi/N)(r+1)}-1}(\hat{x}+j\hat{y}) + \frac{e^{j2\pi(r-1)}-1}{e^{j(2\pi/N)(r-1)}-1}(\hat{x}-j\hat{y})\right], \tag{2}$$

valid for $r \neq \pm 1$.

We thus see that $\vec{H}_p = 0$ for all values of r except $r = \pm 1$. When $r = \pm 1$, the sum of all contributions leads to

$$\vec{H}_p = a(\hat{x} \mp j\hat{y}), \tag{3}$$

where

$$a = \frac{NI_o e^{j\omega t}}{4\pi R}. \tag{4}$$

These are the modes of interest, as they couple to the ferrite and undergo rotation. The total rf magnetic field at P, in the absence of ferrite, thus rotates in the transverse plane with angular frequency ω and constant amplitude a, the rotation being clockwise for one TEM mode and counterclockwise for the other TEM mode. All other TEM modes, r, give zero magnetic field at P and do not couple to the ferrite, to first order.

The total rf magnetic field inside the ferrite is given by

$$\vec{H}_p^f = \vec{H}_p - \frac{\vec{m}}{2} \tag{5}$$

where m is the rf magnetization inside the ferrite and $-(\vec{m}/2)$ is the demagnetizing field appropriate to a cylindrical geometry. The form of the driving rf field $\vec{H}_p$ (3) leads us to write for the magnetization

$$\frac{\vec{m}}{2} = b(\hat{x} \mp j\hat{y}) \tag{6}$$

where b is to be determined. The rf magnetic induction inside the ferrite cylinder is

$$\vec{B}_p^f = T\vec{H}_p^f \tag{7}$$

where

$$T = \begin{bmatrix} \mu & -jk & 0 \\ jk & \mu & 0 \\ 0 & 0 & 1 \end{bmatrix}\mu_o. \tag{8}$$

The elements μ and k are given by

$$\begin{aligned} \mu &= 1 + \frac{M_s\gamma\omega_0}{\omega_o^2-\omega^2} \\ k &= \frac{M_s\gamma\omega}{\omega_o^2-\omega^2} \\ \omega_o &= \gamma H_o. \end{aligned} \tag{9}$$

From (3), and (5)–(8), and from the fundamental relation $\vec{B}_p^f = \mu_o(\vec{H}_p^f + \vec{m})$ we find that b is related to a by

$$b = \frac{\mu \mp k - 1}{\mu \mp k + 1}\cdot a \tag{10}$$

for the two modes. The rf magnetic field and magnetic induction inside the ferrite are then given by[14]

$$\vec{H}_p^f = \frac{2a}{\mu \mp k + 1}(\hat{x} \mp j\hat{y}) \tag{11}$$

$$\vec{B}_p^f = 2a\mu_o\frac{\mu \mp k}{\mu \mp k + 1}(\hat{x} \mp j\hat{y}) \tag{12}$$

so that both field and induction are circularly polarized, though each mode has a different amplitude and they are polarized oppositely.

We are now in a position to calculate the effect of the ferrite on the propagation constants of the two modes on the loaded transmission line. First, however, we shall discuss the unloaded transmission line and calculate the unloaded propagation constant. This result will then be modified to include the perturbing effect of the ferrite.

Unperturbed Transmission Line Equations

The transmission line equations for the nth wire of the unloaded transmission line are[15]

[14] A. D. Berk and B. A. Lengyel, "Magnetic fields in small ferrite bodies," PROC. IRE, vol. 43, pp. 1587–1591; November, 1955.

[15] See, for example, W. R. Smythe, "Static and Dynamic Electricity," McGraw-Hill Book Co., Inc., New York, N. Y., p. 37; 1939. S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., Inc., New York, N. Y., p. 235; 1943.

$$\frac{dV_n}{dz} = -\sum_{i=0}^{N-1} L_{ni}\frac{dI_i}{dt}$$

$$\frac{dI_n}{dz} = -\sum_{i=0}^{N-1} C_{ni}\frac{d}{dt}(V_n - V_i). \tag{13}$$

It is not difficult to show, because of the rotational symmetry of the structure, that each wire obeys the same transmission line equation. We shall work, therefore, with the zeroth wire for convenience, and take as our level of zero potential the center of the structure, P. After using (1) in (13) we find

$$\begin{aligned} \frac{dV_o}{dz} &= -j\omega I_o A \\ \frac{dI_0}{dz} &= -j\omega V_o B \\ A &= \sum_{n=0}^{N-1} L_{o,n} e^{j2\pi rn/N} \\ B &= \sum_{n=0}^{N-1} C_{o,n}(1 - e^{j2\pi rn/N}). \end{aligned} \tag{14}$$

Assuming a traveling wave $V_o \propto e^{-j\beta z}$ in (14), we have

$$\begin{aligned} \beta &= \frac{\omega}{c} = \omega\sqrt{AB} \\ A &= \frac{1}{Bc^2} \\ Z_o &= \frac{V_o}{I_o} = \sqrt{\frac{A}{B}} = \frac{1}{cB} \\ \frac{dV_o}{dz} &= -\frac{j\omega Z_0}{c}\cdot I_o \\ \frac{dI_o}{dz} &= \frac{-j\omega}{cZ_o}\cdot V_o. \end{aligned} \tag{15}$$

Perturbed Transmission Line Equations and Propagation Constants

The ferrite rod contributes an additional flux

$$(\vec{B_P}^f - \mu_o\vec{H_P})_x \frac{D}{2}$$

through the area bounded by wire 0 and the center of the system P. We are assuming that the rod is thin and that the variation in field and flux across the rod cross section is small. The subscript x indicates component of flux in the x direction. The additional emf that this flux generates in wire 0 causes the first of (14) to be modified to

$$\frac{dV_o}{dz} = -j\omega I_o A - \frac{\partial}{\partial t}\left[(\vec{B_P}^f - \mu_o\vec{H_P})_x \frac{D}{2}\right] \tag{16}$$

or

$$\frac{dV_o}{dz} = -j\omega I_o[A + \delta A]$$

$$\delta A = \frac{(\vec{B_P}^f - \mu_0\vec{H_P})_x D}{2I_o} = \frac{\mu_o DN}{8\pi R}\cdot\frac{\mu \mp k - 1}{\mu \mp k + 1}. \tag{17}$$

The propagation constant on the ferrite loaded transmission line is given by the first of (15) with A modified to $A + \delta A$:

$$\beta = \omega\sqrt{(A + \delta A)B} = \frac{\omega}{c}\sqrt{1 + Bc^2\delta A}, \tag{18}$$

so that

$$\begin{aligned} \beta_+ &= \frac{\omega}{c}\sqrt{1 + Bc^2\cdot\frac{DN}{8\pi R}\cdot\frac{\mu - k - 1}{\mu - k + 1}\cdot\mu_0} \\ \beta_- &= \frac{\omega}{c}\sqrt{1 + Bc^2\cdot\frac{DN}{8\pi R}\cdot\frac{\mu + k - 1}{\mu + k + 1}\cdot\mu_0} \end{aligned} \tag{19}$$

for the two modes that couple to the ferrite.

The mutual capacitances between wires in (14) are given by

$$C_{o,n} = \frac{\pi\epsilon_o}{\ln\frac{d_{o,n}}{\rho}} \tag{20}$$

ρ being the radius of each wire and $d_{o,n}$ the distance between the zeroth and nth wire centers. Eq. (20) is valid for $\rho \ll d_{o,n}$. Then B is given by

$$\begin{aligned} B &= \pi\epsilon_o B^* \\ B^* &= \sum_{n=1}^{N-1}\frac{1 - e^{j(2\pi rn/N)}}{\ln\frac{d_{o,n}}{\rho}}. \end{aligned} \tag{21}$$

It is easily shown that B^* is the same for the modes $r = +1$ and $r = -1$.

Rotation Per Unit Length; Examples

The rotation of a linear polarization is determined by the propagation constants β_+ and β_- of the two circular modes $r = +1$ and $r = -1$ and is given by

$$\sigma = \tfrac{1}{2}(\beta_+ - \beta_-) \text{ radians/meter.} \tag{22}$$

From (19), (21), and (22) we find

$$\sigma = \frac{\omega}{2c}\left[\sqrt{1 + B^*\cdot\frac{D}{R}\cdot\frac{N}{8}\cdot\frac{\mu - k - 1}{\mu - k + 1}} - \sqrt{1 + B^*\cdot\frac{D}{R}\cdot\frac{N}{8}\cdot\frac{\mu + k - 1}{\mu + k + 1}}\right]. \tag{23}$$

If H_o is just small enough to saturate the sample so that $\omega_o \sim 0$, (9) gives $\mu \sim 1$ and $k \sim -(M_s\gamma/\omega)$ and the rotation is given by

$$\sigma = \frac{\omega}{2c}\left(\sqrt{1 + B^* \cdot \frac{D}{R} \cdot \frac{N}{8} \cdot \frac{M_s\gamma}{2\omega + M_s\gamma}} - \sqrt{1 - B^* \cdot \frac{D}{R} \cdot \frac{N}{8} \cdot \frac{M_s\gamma}{2\omega - M_s\gamma}}\right). \tag{24}$$

In the limiting case $D/R \to 0$, (23) gives, for arbitrary applied field H_o,

$$\sigma = \frac{B^*N}{32c} \cdot \frac{D}{R} \cdot \frac{M_s\gamma\omega^2}{\gamma^2\left(H_0 + \frac{M_s}{2}\right)^2 - \omega^2} \tag{25}$$

We note, in passing, the resonance in $\vec{H}_p{}'$, $\vec{B}_p{}'$, and σ at

$$\omega = \gamma\left(H_0 + \frac{M_s}{2}\right),$$

appropriate to a ferrite cylinder.

As a first example consider the gyrator ($N=4$). Taking as typical values $M_s = 500$ gauss,

$$H_o \sim 0, \quad \frac{D}{R} = \frac{1}{8} \quad \text{and} \quad \rho/R = \frac{1}{10}$$

at the frequency 1000 mc, we find (see Fig. 2)

$$\frac{d_{01}}{\rho} = \frac{d_{03}}{\rho} = 14.1,$$

$$\frac{d_{02}}{\rho} = 20,$$

and $B^* = 1.40$. Eq. (24) gives $\sigma = 0.76°$/cm. This compares with somewhat larger rotations obtained experimentally by Rowen for larger diameter ferrites where the perturbation theory developed here is not strictly applicable.

For the case of a circulator ($N=8$), with the same geometric and magnetic conditions at 1000 mc as above, we have

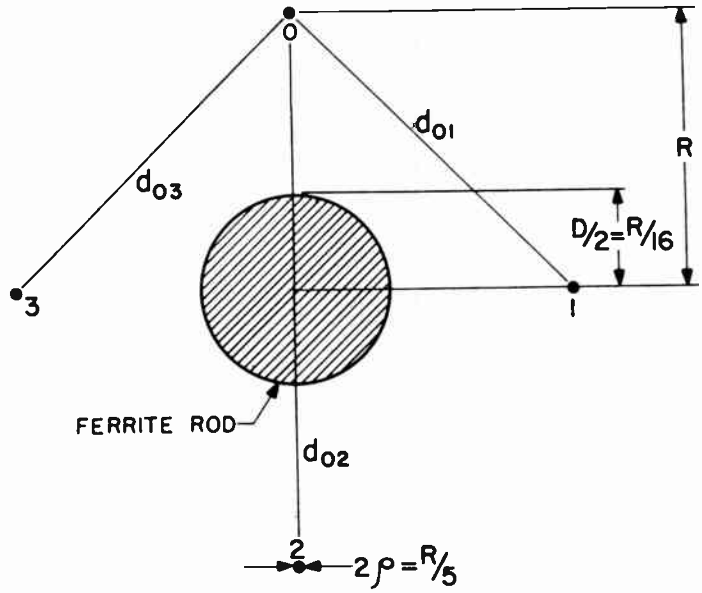

Fig. 2—Example of 4-wire gyrator geometry.

$$\frac{d_{01}}{\rho} = \frac{d_{07}}{\rho} = 7.66, \quad \frac{d_{02}}{\rho} = \frac{d_{06}}{\rho} = 14.1,$$

$$\frac{d_{03}}{\rho} = \frac{d_{05}}{\rho} = 18.5, \quad \frac{d_{04}}{\rho} = 20, \quad B^* = 2.90,$$

and from (24) $\sigma = 4.1°$/cm. The length of circulator required would be 11 cm.

As yet, there are no experimental data on the 8-wire line to confirm these analytical results for the circulator.

In conclusion, we feel that the analysis presented here indicates, for the ideal case of infinite length of line, that the rotations obtainable for 4-wire and 8-wire lines under typical operating conditions are of sufficient magnitude to indicate these devices to be practical low-frequency gyrators and circulators.

Acknowledgment

We should like to thank J. H. Rowen and A. M. Clogston of Bell Telephone Laboratories for suggesting the problem and for several stimulating discussions. The treatment presented here is a generalization of work done by Dr. Clogston on the 4-wire line and we are grateful to him for access to his unpublished notes.

Backward-Wave Oscillator Experiments at 100 to 200 Kilomegacycles*

ARTHUR KARP†, SENIOR MEMBER, IRE

Summary—Electronically tunable oscillations were obtained at wavelengths between 3 and 1.5 millimeters in a demountable backward-wave oscillator whose circuit structure was a ridged waveguide with transverse slots in the broad wall. The slot arrays were formed by tapes wound on suitably dimensioned, interchangeable frames, each giving a tuning range of about 15 per cent. For electron beam velocities between 650 and 2700 volts and current densities between 3 and 10 amperes per square centimeter, probable power outputs of a few tenths of a milliwatt were obtained.

One of the goals of the work was to gain insight into the "personality" of traveling-wave tubes at these high frequencies. Among the new concepts to be dealt with are "beam skin effect" and the very strong influence of circuit loss on starting current and power output.

Introduction

At MILLIMETER wavelengths the dimensions of the rf portion of an electron tube must be extremely small and as the wavelength decreases it becomes increasingly difficult to fabricate the tube, admit sufficient electron current, and dissipate the latter's energy. In tubes of the traveling-wave family, the situation is relatively favorable since the rf structure is distributed in at least one dimension. Space-harmonic designs can further reduce the handicap by permitting an increase in some dimensions. Still, even with these aids, wavelengths below 3 mm (frequencies above 100,000 megacycles) present quite a challenge. It is felt that the challenge has been met, at least for wavelengths down to 1.5 mm (200,000 mc), by the use of a particular slow-wave structure that is as readily fabricated for the shortest wavelengths as for the longer ones.

In the experimental work, it has been convenient to build the tubes as backward-wave oscillators (bwo),[1] since 1) the tube is its own signal source and needs no separate signal generator to be tested, 2) the match between the slow-wave structure and the external circuit is not too critical, and 3) the bwo itself can be useful as a source for other experimental work, especially when it tunes electronically over a range where there are as yet no other fundamental, coherent, cw sources available.

The Circuit

The circuit structure used is basically a ridged waveguide with slots in the broad wall (Fig. 1), as introduced in an earlier article.[2] It may also be considered as a relative of the capacitively-loaded "ladder line."[3]

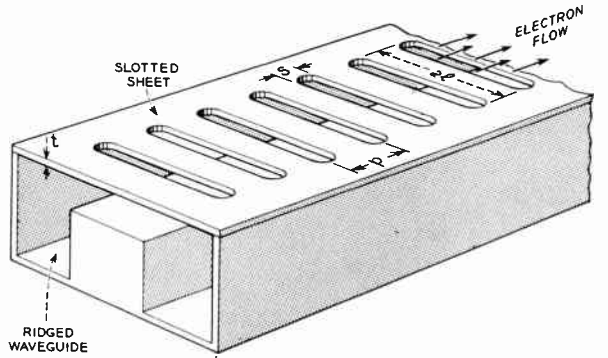

Fig. 1—Basic form of circuit structure.

Electrons made to travel parallel to the slotted surface, on either side, may interact with an rf electric field as they cross each slot. The phase shift from one slot to the next is given as a function of frequency by the dispersion curve, Fig. 2. The first backward wave may be considered to undergo a unit phase shift whose value is between π and 2π and equal to 2π minus the phase shift for the fundamental wave. The dispersion curve may be determined experimentally, using, for example, a probe to measure the guide-wavelength on a scale model, or analytically, using an equivalent circuit, such as that of Fig. 3, which has been found adequate for all practical purposes. The slots are series elements, represented by the transmission-line stubs, and the shunt capacity is that between the top of the ridge and the metal between slots. Neglecting resistance and coupling between slots,

$$\cos \theta_{\text{fund.}} = 1 - \tfrac{1}{2}\omega C z_0 \tan (\omega l/c) \qquad (1)$$

where the radian frequency $\omega = 2\pi f$ and c is the velocity of light.

The phase velocity equals $2\pi p f/\theta$, where p is the distance from one slot to the next. The group velocity equals $2\pi p(\partial\theta/\partial f)^{-1}$, where the derivative is the slope of the curve of Fig. 2. The circuit impedance, which varies inversely as the group velocity, then becomes proportional to the slope of the curve, $\partial\theta/\partial f$. The insertion loss

* Original manuscript received by the IRE, September 26, 1956; revised manuscript received, November 19, 1956.
† Bell Telephone Labs., Inc., Holmdel, N. J.

[1] R. Kompfner and N. T. Williams, "Backward-wave tubes," Proc. IRE, vol. 41, pp. 1602–1611; November, 1953.

P. Guénard, O. Doehler, B. Epsztein, and R. Warnecke, "Nouveaux Oscillateurs à large bande d'accord électronique," *Compt. Rend. de l'Acad. des Sci.*, vol. 235, p. 236; 1952.

[2] A. Karp, "Traveling-wave tube experiments at millimeter wavelengths with a new, easily built, space harmonic circuit," Proc. IRE, vol. 43, pp. 41–46; January, 1955.

[3] R. Warnecke, O. Doehler, and P. Guénard, "Sur les lignes à retard en forme de peigne," *Compt. Rend. de l'Acad. des Sci.*, vol. 231, pp. 1220–1222; 1950.

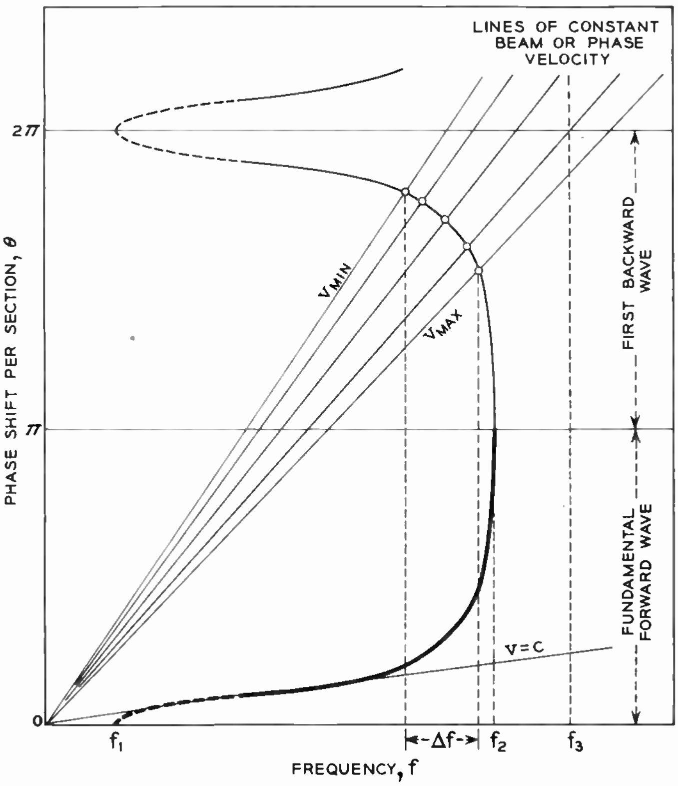

Fig. 2—Dispersion curve for the circuit of Figs. 1 and 4 for the first two spatial components. Lower cutoff frequency f_1=cutoff of ridged guide itself (not important if low); f_2=upper cutoff frequency; f_3=resonant frequency of slots$=c/4l$.

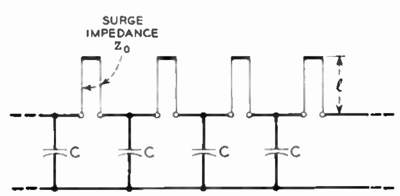

Fig. 3—An approximate equivalent of the circuit of Figs. 1 and 4.

is also inversely proportional to the group velocity,[4] so that the insertion loss in nepers per section (if small) becomes $(f/2Q)(\partial\theta/\partial f)$, where the Q is that of the slots considered as resonators.

The advantages of plotting the dispersion curve as in Fig. 2, where $\partial\theta/\partial f$ is readily observable, can thus be appreciated. A virtue of plotting the ordinate as phase shift, θ, instead of propagation constant, $\beta=\theta/p$, is to give convenient boundaries—multiples of π—between the regions representing the different space harmonics. From the Fourier analysis, the amplitude of the first backward wave, relative to the total amplitude of all the spatial components, decreases as θ varies from π to 2π, at a rate depending on the ratio s/p (see Fig. 1). The rate at which the beam-modulating fields decay in the direction normal to the slotted sheet increases with θ, so that the beam coupling decreases as θ increases.

In practice, one obtains backward-wave oscillations only over the frequency range, Δf, where the slope of the dispersion curve (Fig. 2) is of intermediate value. A limit is imposed at the high-frequency (high slope) end by the rapid increase of circuit loss, and at the low-frequency (low slope) end by the simultaneous falling off of the circuit impedance, the beam coupling, and the Fourier amplitude coefficient of the backward wave. To make the transitional-slope region as gradual as possible, and hence maximize Δf, one finds from plots of (1) that C and z_0 should be made as large as is practical. This indicates that the top of the ridge (whose optimum width is approximately equal to l) should be brought as close as feasible to the slotted wall and that the thickness, t, of the metal forming the slots should be as small compared to the slot opening, s, as thermal and loss considerations permit. Increasing the pitch, p, as for a high-beam-voltage design, also leads to a wider transitional-slope region, which can be considered to be due to the heretofore neglected phase shift in the guide itself and possibly to a decrease in coupling between slots. Making the waveguide size larger than the slot length, $2l$, as opposed to having vertical walls at the slot ends, extending both above and below the sheet, as in the original "ladder line,"[3,5] has in itself "broadened" the dispersion curve in the desired way, apparently by modifying the H-field configuration near the slot ends.

The mechanical realization of the rf structure at extremely short wavelengths is based on fabricating the slotted wall as a grid of tapes. The design that has been found the most convenient to date, which was developed in cooperation with W. H. Yocom, is shown in Fig. 4. The grid is wound on an auxiliary frame, which is separate from the waveguide portion. The tapes must be securely brazed to the frame, but the frame usually need only be clamped to the waveguide assembly. It has been found convenient to use the same waveguide assembly, whose dimensions are not critical, for several frequency bands, inserting different frames , whose only critical dimension is the channel width $2l$, as required. The depth of this channel, over which the tapes are wound, is not critical (provided it is not too shallow) since this region is nonpropagating in the pass band of the structure.

An adequate match into rectangular waveguide is obtained by reducing the channel width at the output end of the frame (see Fig. 7) so as to "taper" out the inductive reactance of the slots (z_0 tan $\omega l/c$) to a low value. The ridge is tapered out in any convenient portion of the output waveguide (see Fig. 5).

Frames and tapes used in the experiments were of molybdenum, high-temperature coated and brazed with pure gold. Early assemblies relying on sintering with gold and copper in combination resulted in tubes that

[4] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y., p. 95; 1950.

[5] *Ibid.*, Fig. 5.7, p. 90.

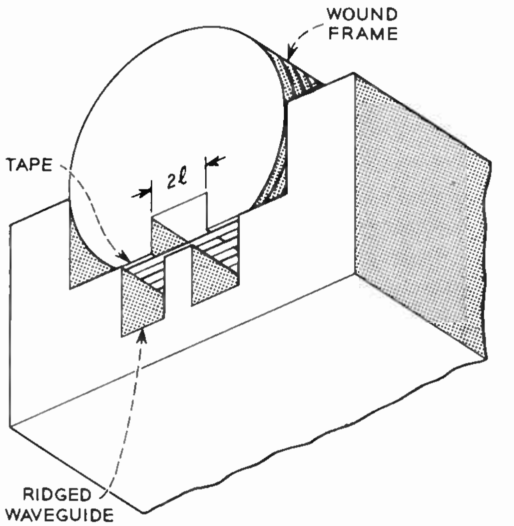

Fig. 4—Mechanical realization of the circuit of Fig. 1, as used in the experimental tubes.

oscillated very weakly or not at all. The explanation is that gold-copper alloys may have resistivities several times that of either pure metal and, as will be shown later, the conductivity of the metal around the slots is very critical at wavelengths below 3 mm. The tapes must of course lie very flat at all times and are wound under tension. Tape $\frac{1}{2} \times 3$ thousandths of an inch in cross section, wound at about 200 turns per inch, was used in these experiments, and there was no difficulty in obtaining the tape or in machining or winding the frames. In fact, if circuit fabrication were the only worry there would be no difficulty generating several times shorter wavelengths by these means. There are real difficulties, however, concerning cathode current density, beam focusing, heat dissipation, circuit loss, and signal detection.

The Experimental Tube

The experiments were performed with a demountable tube described by Figs. 5 and 6. The tube was located between the flat pole faces of a large electromagnet, with the axis of the electron beam (Fig. 6) coincident with the axis of symmetry of the pole pieces. The gun was designed for mechanical simplicity, with the magnetic field doing the focusing in a "brute force" manner. The cathode was a directly-heated hairpin of tungsten ribbon, heated sufficiently to emit several amperes per square centimeter either with or without thorium oxide coating. The coin-silver output waveguide was in RG 99/U size (0.061-inch $\times$ 0.122-inch id), which was the smallest on hand at the time, although rather oversize for the wavelengths obtained. The reflection caused by the right-angle bend in the ridged portion of the guide is small because of the low impedance of ridged guide.

Five wound molybdenum frames (over-all dimensions $\frac{3}{8}$ inch diameter $\times 1\frac{1}{4}$ inches long) were used, with dimensions, and wavelength band obtained, as in Table I.

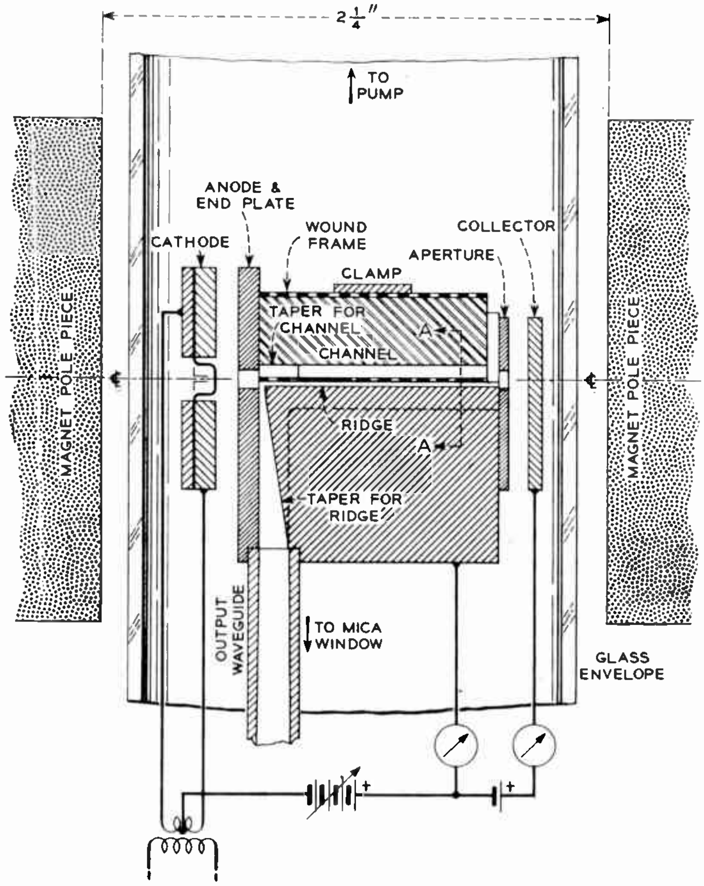

Fig. 5—Demountable backward-wave oscillator tube, 100–200 kmc (schematic).

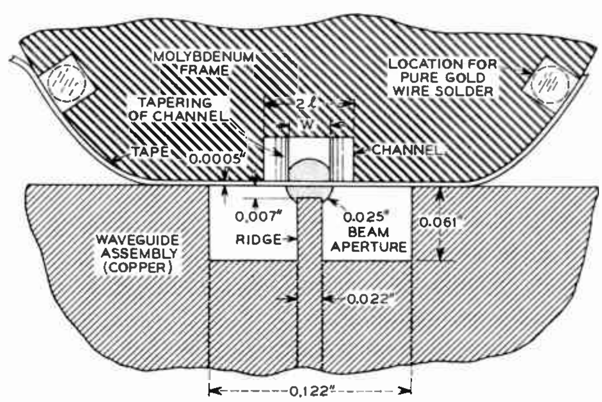

Fig. 6—Cross section A-A of Fig. 5 (cross section through beam, looking towards cathode).

TABLE I

Tape span, $2l$ (inches)	Taper throat, w (inches)	Turns per inch (tpi)	Wavelength band (mm)
0.043–0.044	0.024	188	2.9 –2.5+
0.036–0.040	0.022	200	2.54–2.17
$0.035^{+0.0015}_{-0.000}$	0.022	200	2.47–2.05
0.029 ± 0.001	0.017	200	2.00–1.71
$0.024^{+0.001}_{-0.000}$	0.012	200	1.7 –1.5⁻

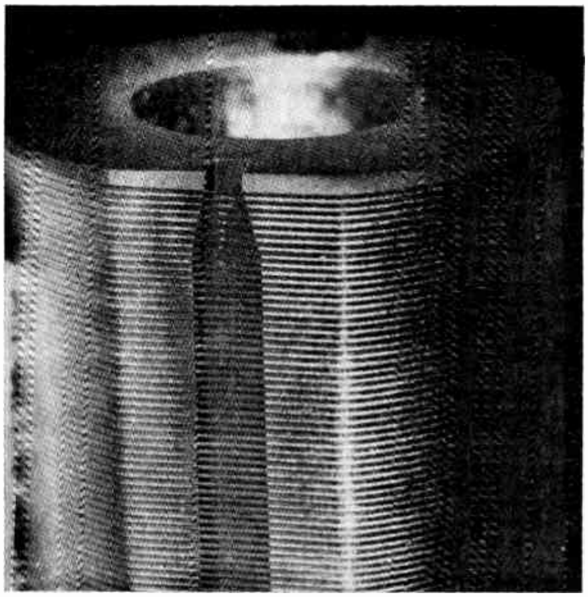

Fig. 7—Closeup of output end of typical frame for backward-wave oscillator. Apertures of this grid are resonant near 134,000 megacycles; provide oscillations in 2.5–2.9 mm band.

A closeup view of a typical frame is shown in Fig. 7.

Since the synchronous beam voltage is given by

$$V_0 = \left[(\text{phase velocity})/2\frac{e}{m}\right]^2$$
$$= [507\ (2\pi/\theta_{-1})(p/\lambda_0)]^2 \text{ volts,}$$

where $\pi<\theta_{-1}<2\pi$ and λ_0=free-space wavelength, the operating voltage must be increased if p is held constant (=1/200 inch) while λ_0 decreases. Corresponding to the five wavelength ranges for the five frames listed above, the voltage ranges over which backward-wave oscillations were tuned were, respectively: 650 to 920, 750 to 1300, 950 to 1300, 1300 to 1800, and 1700 to 2700 volts.

To collimate the electron stream sufficiently well to start and maintain oscillations, magnetic fields between 1500 and 3700 Gauss were required.

The trend was for the higher frequency frames to require higher fields, but it was appreciably masked by variations in the degree of perfection of alignment among the several assemblies.

(It is likely that these high fields are required more to "straighten out" the trajectories of electrons leaving the gun than to overcome space-charge forces along the length of the circuit and, as was observed by W. H. Yocom and C. F. Hempstead in sealed-off tubes at wavelengths greater than 3 mm, a proper gun design should permit a substantial reduction in field strength.) For the longer wavelengths, maximum tuning range and power output were obtained with the magnetic field and electron flow parallel to the plane of the tapes, as indicated by maximum collector current. For the shorter wavelengths, however, best performance was obtained when the electron flow was inclined at an angle to the plane of the tapes, as indicated by much-less-than-maximum collector current. An explanation for this effect is given later.

For each frame, the starting current is lowest near the center of the tuning range, and for the several frames, starting currents increased from the neighborhood of 1 ampere per square centimeter for the first to the neighborhood of 5 amperes per square centimeter for the fifth. Satisfactory output was obtained using about 3 amperes per square centimeter to about 10 amperes per square centimeter, respectively. Again, the trends were not smooth, but strongly affected by slight mechanical differences among the several assemblies. It is to be noted that the beam current is spoken of only in terms of its density, which appears to be the sensible thing to do at these wavelengths where a practical beam cross section is always very much greater than the cross section actually modulated by the rf fields (see Fig. 10). This situation, which is analogous to the "skin effect" in metallic conductors, will receive further comment later.

Using the above voltage and current-density data, the beam power density incident on the tapes under normal operation was between about 2 and 27 kilowatts per square centimeter. The beam power density at which a 0.003×0.0005-inch tape burns out was found to be in the neighborhood of 10 kw per square centimeter—depending, of course, on the span (0.024–0.044 inch), the tension, the angle of beam incidence, and the brazed joint.

Thus, the first three frames could be operated fully cw, while it was necessary to operate the last two with the average input power reduced by about $\frac{2}{3}$. Since backward-wave oscillators are most often used as wide-range frequency-sweep generators, modification of the beam-voltage wave form to provide a little reduction in average input power is readily accomplished. For example, to observe the curves in Fig. 9, the anode supply was a combination of −1000 volts dc and 4000 volts peak 60 cps ac, with oscilloscope sweep and blanking adjusted to display the detector signal during the 0.0025-second interval when the beam voltage sweeps from +1500 to +3000 volts.

The rf signal was detected with a silicon crystal in a mount, developed by W. M. Sharpless, which is in general use at the Holmdel Radio Laboratory in the 3.5 to 7-mm range. Although designed for these longer wavelengths, some mounts could be selected having adequate sensitivity in the 1.5 to 3-mm range. The crystals were backed by a micrometer-driven waveguide piston. As the piston is moved, the detected signal at a given wavelength exhibits periodic minima and maxima, so that the piston-micrometer combination can serve as an interferometer-type wavemeter. Since the piston travel covers many half-wavelengths, wavelengths may be measured with fair accuracy. As a preliminary check, one may pass the rf energy through guides necked down to various small "go" and "no go" cross sections (such as 0.060×0.030-inch or 0.040×0.020-inch) to establish upper and lower limits on the wavelengths being generated.

Typical oscillograms of detector output vs beam voltage are shown in Fig. 8 (frame 2) and Fig. 9 (opposite, frame 5).

Since frequency is roughly proportional to beam voltage, these are also curves of output vs frequency. In Fig. 9, there are two or three "piston dips" which move to the right as the piston advances toward the crystal, but they are shallower than those of Fig. 8 due to higher piston contact losses. In all the oscillograms, considerable "fine-structure" is seen, which is a sort of interference pattern caused by multiple reflections in the structure and output guide, which are both very many wavelengths long. In the curves of Fig. 9, the "fine structure" appears less extensive than it actually is due to the rapid sweep and limited bandwidth of the audio-transformer inserted to better match the detector crystal to the scope amplifier.

Reflections in the output guide may be caused by the right-angle bend, ridge taper, mica window, waveguide joints, and detector mismatch. An interference pattern results from each pair of reflectors, with a periodicity (in frequency) inversely proportional to the distance between them. Since these reflections are small and passive, the resulting contribution to the fine structure is shallow and can be reduced by inserting attenuation in the output guide. Reflections on the slow-wave structure may occur at the ends and at microscopic irregularities along its length, which contains many of the slowed wavelengths. A reflected wave traveling "downstream" (towards the collector) is attenuated rapidly by the high loss of the structure while a wave reflected "upstream" is amplified by the beam, so reflections on the circuit are not passive. In low-frequency bwo's the only pair of reflections that matters is due to the ends and it is feasible and sufficient to terminate the collector end of the circuit without reflection. At frequencies above 100 kmc, however, reflections at intermediate points are likely to be substantial and the loss, looking towards the collector, is already several tens of db per cm. Thus, at these high frequencies, pronounced "fine structure" would appear to be an inherent problem, with little to be gained by adding loss at the collector end.

Power Output

Lacking direct means to do so, the power output was not measured experimentally. Detector efficiencies at these frequencies were also unknown. However, it should be possible to obtain a close estimate of the power output by the following calculation, based on the work of Watkins and Grow,[6] which has been in good agreement with experiment at wavelengths of 5 mm and longer.

[6] R. W. Grow and D. A. Watkins, "Backward-wave oscillator efficiency," Proc. IRE, vol. 43, pp. 848–856; July, 1955.

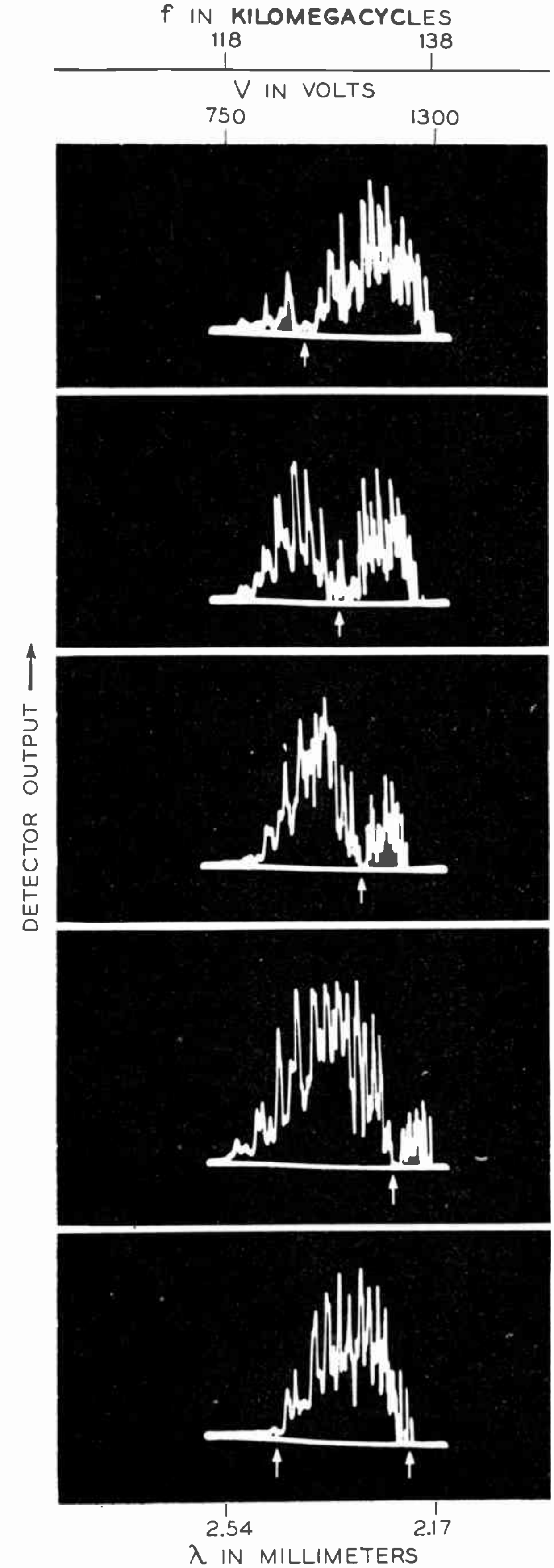

Fig. 8—Typical oscillograms of detector output vs beam voltage, for oscillator tested in February, 1955, for different positions of the piston behind the detector crystal. (Arrows show frequencies for which piston minimized detector response.)

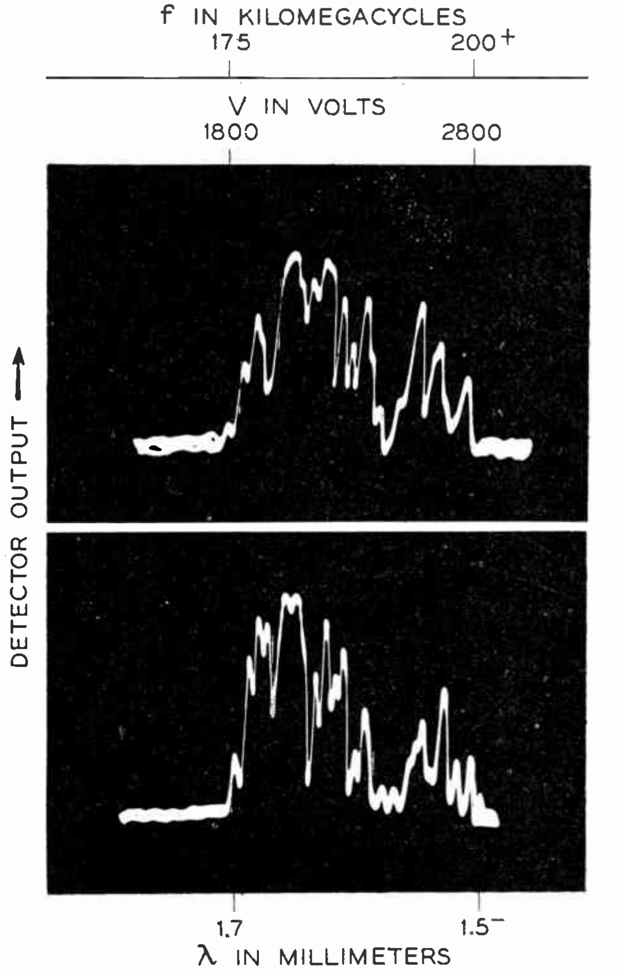

Fig. 9—Typical oscillograms of detector output vs beam voltage, for oscillator tested in September, 1955. Piston in different positions; rapid sweep.

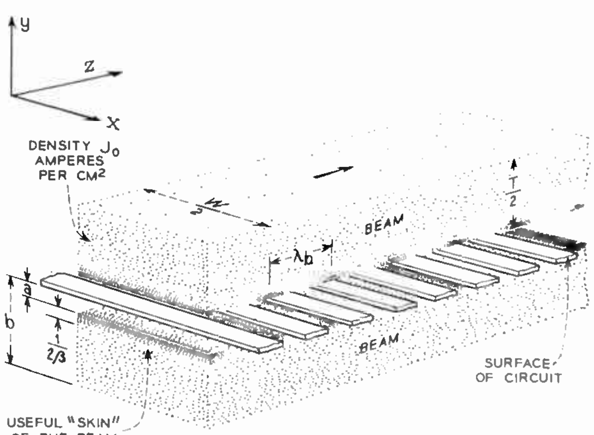

Fig. 10—General representation of beam and circuit of bwo at very short wavelengths.

Fig. 10 is a general representation of the geometry of beams and circuits at these very short wavelengths. For phase velocities small compared to c, the longitudinal rf field, E_z varies in the y direction as $\exp[-\beta y]$, where the propagation constant

$$\beta = \frac{\theta_{-1}}{p} = \frac{\omega}{v} = \frac{2\pi}{(v/c)\lambda_0} = \frac{2\pi}{\lambda_b},$$

if v is the phase velocity of the backward wave which is essentially the same as the beam velocity, U_0. Typically E_z decays to a negligible value in a distance small compared to the thickness of the beam. In a gain or power calculation it is actually E_z^2 that is of interest, so that the effective "skin depth" of the beam is

$$\frac{1}{2\beta} = \frac{v}{c}\,\frac{\lambda_0}{4\pi}.$$

(Note that this depth is approximately 1/12 of the beam wavelength, or of the order of 1/150 of the free-space wavelength!) The power *input* available for conversion to rf is then

$$\frac{J_0V_0W}{2\beta} = J_0V_0W\lambda_0\frac{v/c}{4\pi}\text{ watts.}$$

E_z^2 has been assumed invariant with x; if it is not, an effective reduced value of the width W can be used. Following Watkins and Grow, the efficiency is taken as ω_q/ω, where ω_q is the "modified plasma frequency." When the beam is relatively close to a conductor $(b \gg a)$ and thick compared to beam wavelength $(b \gg \lambda_b)$ ω_q essentially equals[7] the "unmodified" plasma frequency, ω_p, and there is no further need to know dimensions b and a.

Multiplying the input power by the efficiency, substituting

$$\omega_p = \left[\frac{(e/m)J_0^{1/2}}{\epsilon_0 U_0}\right],$$

$v = U_0 = [2(e/m)V_0]^{1/2}$ and $\omega = c/2\pi\lambda_0$, and inserting numerical values for the physical constants, one obtains after simplification,

$$\text{power output} \approx 86.5\ J_0^{3/2}V_0^{5/4}(\lambda_0^2 W), \tag{2}$$

using the units (milliwatts), (amperes per square centimeter)$^{3/2}$, (kilovolts)$^{5/4}$, and (cm)^3, respectively. Since W is usually scaled with λ_0, it is convenient to express W as $m\lambda_0$, and make the last term $(m\lambda_0^3)$, indicating a rapid decrease in available output with decreasing wavelength. The maximum operating value of J_0 is intimately associated with the chemistry of the cathode, since a converging gun and the resulting spiralling beam could not be tolerated.

It is fortunate that (2) can be expressed in terms of only the most basic quantities. The only assumptions made which affect the validity of the expression: efficiency $= \omega_q/\omega$ are that the current density is everywhere uniform, that the ohmic losses are negligible, and that the operating current is sufficiently greater than the starting current to let the beam "saturate" at the edge

[7] J. W. Sullivan, "A wide-band voltage-tunable oscillator," Proc. IRE, vol. 42, pp. 1658–65; November, 1954. See, for example, Fig. 13, p. 1664.

nearest the circuit. Further analysis by Watkins and Grow[6] indicates that the ohmic loss of the structure can be accounted for by reducing the theoretical output by about $\frac{1}{4}$ to $\frac{1}{2}$ of the total circuit loss.

Applying (2) with $m \approx \frac{1}{3}$ (including effect of variation of E_z with x), the following typical values result:

λ_0 (mm)	3	1.5
V_0 (kilovolts)	1	2.5
J_0 (amperes per square centimeter)	3	8
power output before circuit loss (mw)	3.5	6
estimated total circuit loss (db)	30	58
reduction[8] (db)	10	16
net output power (mw)	0.35	0.15

The circuit loss is evaluated by means of the relation given near the beginning of the paper, between the loss per resonator, the group velocity, and the Q of the little resonators. This Q is equal to $\pi/4$ times the surge impedance of the little slot divided by $\frac{1}{2}rl+R$, where r is the surface resistance per unit length of the slot edges carrying rf current and R is the rf resistance of the brazed joint. In the total loss data above, that in the first column corresponds to the high-group-velocity end of the band of that particular frame, and that in the second column corresponds to the low-group-velocity end.

Number of Electrons in a Bunch

As H. Heffner has pointed out, the "bunches" in the typical beam contain surprisingly few electrons when the wavelength is so minute. Using Fig. 10 again, it is seen that the electrons that can be drawn together to form one bunch can at most come from a volume

$$W \text{ wide} \times \frac{1}{2\beta} \text{ thick} \times \lambda_b \text{ long.}$$

This product simplifies to

$$\text{volume} = (U_0/c)^2 m\lambda_0^3/4\pi.$$

The electron density is J_0/U_0e, so that the number of electrons,

$$N = \text{electron density} \times \text{volume}$$

$$= \frac{J_0}{4\pi ce}\left(\frac{U_0}{c}\right)m\lambda_0^3$$

$$= 1.66(10)^7(U_0/c)J_0 m\lambda_0^3$$

with J_0 in amps/cm² and λ_0 in cm.

For the typical case: $\lambda_0 = 0.15$ cm, $m \approx \frac{1}{3}$, $J_0 \approx 5$ amperes per square centimeter, $V_0 = 2500$ volts ($U_0/c \approx 0.1$), and

$$\mathfrak{N} = 9300 \text{ electrons.}$$

It is difficult to say what conclusion, other than that of amazement, may be drawn from this result, or through what effects the existence of the relatively few electrons per bunch might make itself evident. Perhaps the signal is "noisier" than it would be otherwise, and very likely one is near the point where the interaction needs to be analyzed under quantum considerations.

On Calculating the Starting Current

It is always desirable to check experimental results and theory against one another, and the work here is no exception. The quantity of interest is the starting current (density) from which the "gain parameter" and "coupling impedance" can be derived. Unfortunately, however, there exists an essentially unknown parameter —the effective separation between the beam and the circuit (distance a in Fig. 10)—whose effect becomes very pronounced at these wavelengths.

While the value of a has no direct effect on the power output (provided the starting current remains a few times smaller than the operating current) and while it usually is small enough to have little effect at relatively long wavelengths, its effect on the starting current is enormous at the short wavelengths where it may equal or exceed $1/2\beta$. There is in reality no empty region between the beam and the circuit, as suggested by Fig. 10, but most of the electrons immediately next to the circuit are in the process of being intercepted because of their transverse velocities, which may be of thermal or other origin. These electrons, which are often called "fuzzy electrons," cannot participate in producing gain, and a is the *effective* thickness of the "fuzzy" region. Its value depends on the magnetic field and the cathode temperature, among other factors, and is usually estimated to be in the vicinity of 0.001 inch. Because E_z^2 varies as exp $(-2\beta y)$, a small uncertainty in a yields a very great uncertainty in the gain parameter. It is precisely the fact that a slight "adjustment" of the assumed value of a can make the theory agree with just about any measured value of starting current that makes a comparison under these conditions almost pointless.

It will be more fruitful to examine how the starting current depends on the various controllable parameters, so that they may be optimized. There are many such parameters, such as the slot impedance, but it turns out that at these wavelengths a parameter of almost overwhelming importance is the circuit "cold" loss! H. R. Johnson[9] has computed the dependence of the starting condition on this loss, for total losses up to 30 db, and R. W. Grow[10] has extended the results to several hundred db. Both sets of data are expressed as the product (CN) at starting vs the total loss, $\mathfrak{L}$, where C is the gain parameter (proportional to the $\frac{1}{3}$ power of beam current) and N is the active length of the circuit in *beam wavelengths*.[11] It has been found convenient to

[6] Watkins and Grow, *op. cit.*, (23), (24).

[9] H. R. Johnson, "Backward-wave oscillators," Proc. IRE. vol. 43, pp. 684–697; June, 1955.

[10] Private communication.

[11] Johnson's space charge parameter, Q/N, which is equal to $\omega_q^2N^2/4\omega^2(CN)^3$, is always negligible under the operating conditions involved at these short wavelengths.

replot these data ss the value of C at starting vs the active length, N, with ***unit*** loss $\alpha = \mathcal{L}/N$ as a parameter (see Fig. 11).

The significance of Fig. 11 is the following: When the unit loss is low, its influence is negligible, and the starting current varies inversely as the cube of the length.

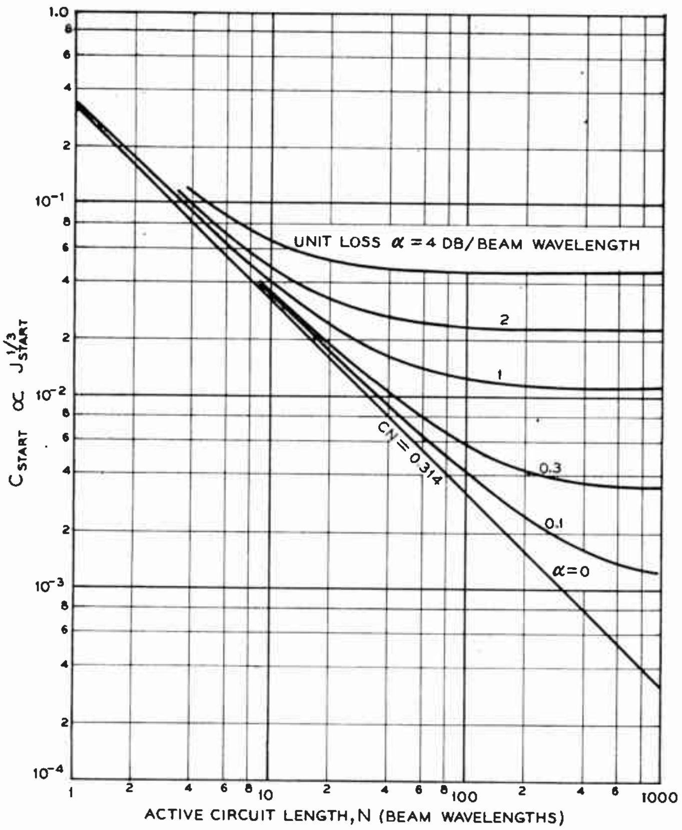

Fig. 11—Starting conditions in bwo with ohmic loss.

This is the familiar, centimeter-wave viewpoint. But when the unit loss is high, it is necessary to reverse the viewpoint, for the effect of length (beyond a minimum value) is then negligible, while the starting current varies as the cube of the loss! This situation has been borne out by the experimental work: When the tapes were gold-copper brazed, starting currents were extremely high or beyond the capacity of the gun. When the electron flow was inclined at an angle to the axis of the circuit, operation improved (at the shortest wavelengths) since the decrease in interaction length had little detrimental effect while the effect of bringing electrons closer to the tapes was favorable.

Recognition of the dominating influence of the loss should govern future work. For example, increasing the thickness of the tapes (dimension t in Fig. 1) would normally be undesirable, for it slightly decreases the coupling impedance and the bandwidth. However, it should appreciably reduce the loss (and incidentally improve thermal dissipation) and this is the more urgent need in improving the present design and in trying to reach still shorter wavelengths.

Conclusion

Extending the operating frequencies of backward-wave oscillators to a few hundred kilomegacycles requires a change in viewpoint from that assumed at lower frequencies. Constructability in the small dimensions becomes practically the sole criterion in choosing circuit structures. Decreasing the loss and increasing the proximity of the beam to the circuit become of major concern, while the length of interaction loses importance. The beam current has physical significance only when expressed as density and when considered as having a "skin effect." It may perhaps also be necessary to consider how few electrons may constitute a "bunch."

The slotted-wall waveguide structure has been found useful up to at least 200 kilomegacycles, while still being cheaply and easily made.

Acknowledgment

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Transistor Junction Temperature as a Function of Time*

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Summary—An analysis of the heat flow in transistors is presented which enables one to determine the variation of junction temperature with time for a given transistor excitation.

A one-dimensional heat flow model is considered which is representative of grown-junction and some alloy transistors. The step and impulse temperature responses are obtained from the solution of the heat equation. A comparison made between the step response so obtained and the often assumed simple exponential response indicates that the assumed response can be low by a factor of two or more.

Utilizing the impulse temperature response for the transistor, the junction temperature as a function of time is determined for periodic rectangular pulse excitation. Numerical calculations are made and curves presented for the maximum, average, and minimum junction temperatures in terms of the duty cycle, repetition rate, and fundamental thermal time constant. These curves indicate that the maximum junction temperature can be several times the average value at low duty cycles and low repetition rates.

Measurements of junction temperatures are presented which essentially substantiate both the step response and the repetitive pulse response which were theoretically obtained.

Finally, using simplified approximate expressions, the procedure for calculating the maximum, average, and minimum junction temperatures for repetitive pulse excitation is described. The use of these predicted junction temperatures with relation to the maximum junction temperature ratings, thermal stability, and electrical parameter changes is also discussed.

I. Introduction

When transistors are operated at low frequencies (below 2000 cycles), particularly in switching and pulse circuits, a considerable junction temperature variation can exist per cycle of excitation. Under these circumstances, equivalent circuit parameters can vary markedly from one portion of a waveform to another and safe dissipation limits must be determined by the peak junction temperature rather than the average. It thus becomes necessary to investigate the variation of transistor junction temperature with time (or the transient and dynamic steady-state temperatures) rather than the often assumed and often treated dc or average temperatures and dissipations. The results of such an investigation are described in this paper.

The approach to this investigation as presented here is to first determine theoretically the temperature step response for the transistor junction and then compare this response to that obtained experimentally (Part II). The second step in the investigation (Part III) consists of utilizing the verified step response to obtain the junction temperature as a function of time for rectangular pulse excitation. Again a comparison is made with experimentally obtained results. In addition to these two steps, the measurement of junction temperature is described (Part IV), and finally a discussion on the use of the knowledge gained in circuit design is presented (Part V).

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II. Junction Temperature Step Response

For the purpose of this paper all analyses will be confined to the one-dimensional, rectangular heat flow problem. Such linear heat flow exists in most bar or grown junction type transistors, as well as some alloy power transistors, depending upon the particular geometry employed.

The solution of the heat equation will first be presented followed by a comparison with the often assumed simple exponential transient and with experimental results.

A. Solution of the Heat Equation for a Step Input of Power

Considering the typical geometry of a grown junction transistor (see Fig. 1), one sees a bar of germanium or

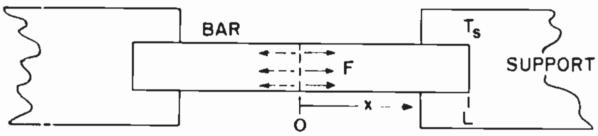

Fig. 1—Transistor model.

silicon supported on mounting wires (emitter and collector leads) which are in turn held rigidly in a header which forms part of the transistor case. Such a geometry immediately suggests a one-dimensional, rectangular heat flow from the collector junction to the extremities of the bar down the supports and finally to the case. Thus a rather simple heat flow geometry can be treated for the purposes of analysis if the following assumptions are made.

1) The heat flow is symmetrical about the collector junction.
2) The total power dissipation is considered existing uniformly across the collector junction plane, *i.e.*, the emitter and collector junctions as well as the base region are considered as one plane with the total power being equal to the sum of the collector, base, and emitter powers with no other distributed power loss in the bar.
3) The heat flow via the base lead is negligible.
4) There is no radiation, convection, or conduction from the lateral surfaces of the bar, *i.e.*, the flow is one-dimensional. (Effects discussed later.)

5) For the periods of time considered, neither the case nor supports (leads) will change their temperatures. (Effects discussed later.)

Symbols:

T = temperature in degrees C.
t = time in seconds.
k = heat conductivity in cal/sec−cm−°C.
α = thermal diffusivity in cm²/sec $= k/c\rho$ where c = specific heat in cal/gm−°C; ρ = density in gm/cc.
x = distance along the bar in cm.
L = half the length of the bar in cm.
A = cross sectional area of the bar in cm².
F = heat flux/unit area in cal/sec−cm².

Treating only half the model as it is symmetrical, the heat equation for one-dimensional flow is

$$\frac{\partial T(x, t)}{\partial t} = \alpha \frac{\partial^2 T}{\partial x^2}, \quad 0 \leqq x \leqq L, t > 0 \tag{1}$$

with the following boundary and initial conditions:

$$\frac{\partial T}{\partial t}(0, t) = -\frac{F}{k} \tag{2a}$$
$$T(L, t) = T_s \tag{2b}$$
$$T(x, 0) = T_s \tag{2c}$$

where T_s = the fixed support temperature. The solution[1] of this boundary value problem is

$$T(x, t) = T_s + \frac{F}{k}(L - x) - \frac{8FL}{k\pi^2} \sum_{n=1,3,\cdots}^{\infty} \frac{e^{-(n^2\pi^2\alpha t)/4L^2}}{n^2} \cdot \cos \frac{n\pi x}{2L} \tag{3}$$

or, if F is replaced by $P/8.36A$ where P is equal to the total power $(V_c I_c + V_e I_e)$ in watts[2], then the solution becomes

$$T(x, t) = T_s + \frac{PL}{8.36\,kA}\left\{(1 - x/L) - \frac{8}{\pi^2} \sum_{n=1,3,\cdots}^{\infty} \frac{e^{-(n^2\pi^2\alpha t)/4L^2}}{n^2} \cdot \cos \frac{n\pi x}{2L}\right\}. \tag{4}$$

If we now confine our interest to the temperature of the junction [and by virtue of assumption (2), the temperature of the base region also] by setting $x = 0$, the final expression for the temperature step response is

$$T_j(t) = T_s + P\theta_b\left\{1 - \frac{8}{\pi^2} \sum_{n=1,3,\cdots}^{\infty} \frac{e^{-t/\tau_n}}{n^2}\right\} \tag{5}$$

[1] See Appendix.
[2] The number 8.36 consists of the factor two associated with $P/2$ (half the junction power used as only half the symmetrical model is considered) and the conversion factor 4.18 to change watts into calories/second

where θ_b = the total thermal resistance of the bar (two halves in parallel) $= L/8.36\,kA$ in °C/watt and τ_n = the thermal time constants $= (2L/n\pi)^2 \cdot 1/\alpha$ in seconds. Thus the transient junction temperature is governed by an infinite summation of exponential terms.

Should sufficiently long times be considered, the summation term in (5) would vanish and the steady-state, dc condition would exist. However, since the supports are not infinite heat sinks, they too would undergo a transient change to a new steady state temperature as would the transistor case if it were not connected to an infinite sink. Thus, the actual junction temperature response with reference to ambient temperature is a series of transient changes of different fundamental ($n = 1$) time constants superimposed on each other. Nevertheless, in general, it will be adequate for normal repetitive excitation to consider only the transient change of temperature in the bar by treating the supports and case as fixed temperature bodies, as the latter two fundamental thermal time constants are so much longer than that of the bar (*e.g.*, for a typical grown junction unit (2N78), $\tau_1 = 10$ ms for the length of the bar; whereas, $\tau \cong 1$ sec for the entire length between junction and case).

B. Comparison of the Temperature Step Response to Simple Exponential Behavior and Experimental Results

In order to compare the behavior of the solution represented by (5) to the simple exponential behavior and to experimental results, the solution has been evaluated and indicated in Fig. 2.

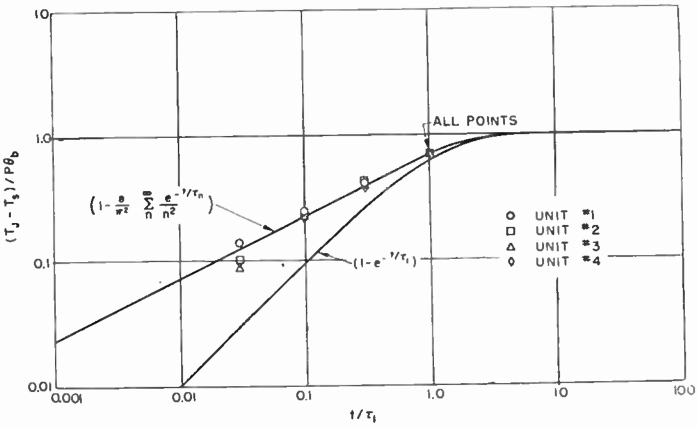

Fig. 2—Plot of normalized junction temperature vs t/τ_1 for step excitation.

Using Fig. 2 to compare the solution of (5) to the often assumed simple exponential form, it can be seen that for times greater than τ_1 the two curves coincide within ten per cent. However, for smaller values of time, the two curves split off into two straight lines (log-log plot), one with a slope of one half and the other a slope of one so that when $t/\tau_1 = 0.01$ the simple exponential form predicts a temperature of only 1/7 that given by (5). Thus, the simple exponential behavior cannot be assumed when $t<\tau_1$ without making serious errors.

It is useful to note that since the true solution is, in part, a straight line on a log-log plot, the junction temperature can be represented (within 12 per cent) by the following approximation:[3]

$$T_j(t) = T_s + P\theta_b\sqrt{t/2\tau_1}, \quad 0 \leq t/\tau_1 \leq 2.0. \tag{6}$$

Also plotted in Fig. 2 are the experimental results obtained for four grown junction transistors (experimental 2N78's). In obtaining the step response (procedure described in Part IV), the device thermal parameters were also determined and are given in Table I.

TABLE I

MEASURED AND CALCULATED VALUES OF THE THERMAL PARAMETERS

Parameter	Measured Values				Calc. Values (Evacuated)
	No. 1	No. 2	No. 3	No. 4	
τ_1 (bar)—*ms*	11	13	14	12	15
τ_1 (junction-case)—*s*	—	1.2	—	—	1.7
*θ_b—°C/w	126	130	130	103	154
θ_s—°C/w	544	506	302	195	659
θ_t—°C/w	670	636	432	298	813

* Can vary by as much as ±50 per cent between units because of bar area differences.

Note: Unit No. 1—evacuated.
No. 2—backfilled, O_2 at 1 atm.
No. 3 and 4—backfilled, 20 per cent O_2, 80 per cent He, at 1 atm.

θ_s = thermal resistance of the support.
θ_t = total thermal resistance, junction to case $=\theta_b+\theta_s$.

From Fig. 2, it can be seen that the experimental response for unit No. 1 agrees well with the theoretical response. However, the response obtained for the gas filled units is slower than the theoretical response at $t/\tau_1 \leq 0.03$. The effect of the gas is to extract heat from the surface of the solid heat path (a three dimensional heat flow) thus tending to slow down the heating up of the junction. Comparing the responses of the various units with their respective total thermal resistances, it can be seen that the gas filled units Nos. 2–4 which exhibit the slower response do indeed have appreciably smaller total thermal resistances because of the conductance of the gas. Furthermore, the helium filled units with their higher gas conductivity exhibit the most pronounced effects. No trend in the effects of the gas on thermal time constant is apparent from the data obtained, indicating that the germanium bar still controls this factor independent of the presence of the gas.

In summary, the theoretical step response obtained is substantiated by experiment when gas is not present or unimportant thermally (most cases where the thermal resistance of the support is small). Further, when the conductance of heat through the gas is comparable to that through the solid heat path, a slowing down of the theoretical response is experienced for times less than 0.03 τ_1.

[3] This approximate relationship for small values of t/τ_1 was also derived analytically by P. Weiss, Gen. Elec. Electronics Lab., Syracuse, N. Y., in a private communication.

III. JUNCTION TEMPERATURES FOR PULSE EXCITATION

The results obtained in the previous section for the junction temperature step response will now be used to determine junction temperatures for periodic excitations, in particular, for rectangular pulse excitation.

The general expression for junction temperature as a function of time for pulse excitation will first be obtained, followed by particular numerical and graphical interpretations and, finally, by some supporting experimental data.

A. Theoretical Results

The junction temperature response can now be determined for any excitation from either the step or impulse response by making use of the convolution integral,

$$T_j(t) \text{ for pulse excitation } = \int_0^t P(t')T_j'(t-t')dt' + T_s, \tag{7}$$

where $P(t)$ is the periodic power excitation, $T_j'(t)$ = the junction temperature impulse response, and T_s = support temperature for pulse excitation.

Because of its importance, the periodic rectangular pulse excitation will now be considered in detail. If the total junction power applied to the transistor is a repetitive pulse (see Fig. 3) then

$$P(t) = P_0 d\left\{1 + 2\sum_{m=1,2,3,\cdots}^{\infty} \frac{\sin m\pi d}{m\pi d}\cdot\cos 2m\pi t/\tau_r\right\} \tag{8}$$

where P_0 = the amplitude of the pulse, d = the duty cycle $=\delta/\tau_r$, and τ_r = the repetition period. Furthermore, by differentiating the unit step function response [(5), $P=1$], the impulse response can be expressed as follows:

$$T_j'(t) = \frac{2\alpha\theta_b}{L^2}\sum_{n=1,3,\cdots}^{\infty} e^{-t/\tau_n}. \tag{9}$$

Convolving these two expressions using (7), (8), and (9), the complete expression for the junction temperature response is obtained.

$$T_j(t) = \frac{4P_0 d\alpha\theta_b}{L^2}\left\{\sum_n^{\infty}\frac{\tau_n}{2}\left[1-e^{-t/\tau_n}\right] + \sum_{n,m}^{\infty}\frac{\sin m\pi d/m\pi d}{\omega_m^2+(1/\tau_n)^2}\left[\frac{1}{\tau_n}\cos\omega t + \omega_m\sin\omega_m t - \frac{1}{\tau_n}e^{-t/\tau_n}\right]\right\} + T_s$$

$$n = 1, 3, 5, \cdots \text{ and } m = 1, 2, 3, \cdots. \tag{10}$$

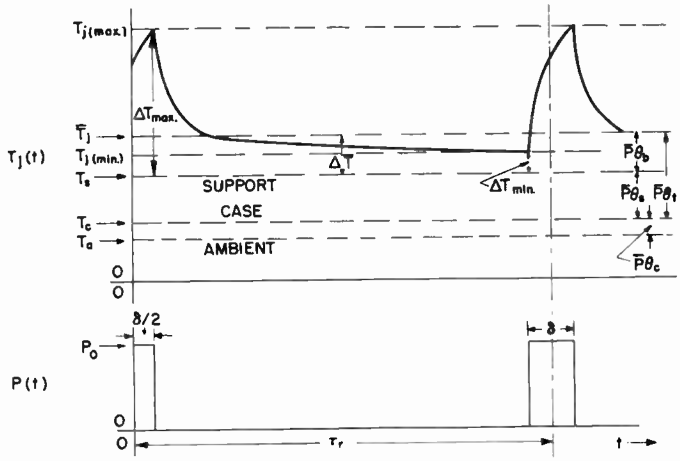

Fig 3—Junction temperature and excitation as a function of time.

Considering only the dynamic quasi-steady-state junction temperature (*i.e.*, neglecting initial transients) which can be accomplished by setting the factor e^{-t/τ_n} equal to zero, (10) reduces to

$$T_j(t) = \frac{4P_0 d\alpha\theta_b}{L^2}\left\{\sum_{n(\text{odd})}^{\infty}\frac{\tau_n}{2} + \sum_{n,(\text{odd}),m}^{\infty}\frac{\sin m\pi d/m\pi d}{\omega_m^2 + (1/\tau_n)^2}\cdot\left[\frac{1}{\tau_n}\cos\omega_m t + \omega_m \sin\omega_m t\right]\right\} + T_s. \qquad (10a)$$

To numerically and graphically interpret the solution expressed in (10a), expressions for the average, maximum, and minimum junction temperatures will now be obtained. The average junction temperature is simply the first summation plus T_s in (10a) as the sine and cosine terms of the second summation have zero average values. Thus,

$$\overline{T}_j = \frac{4P_0 d\alpha\theta_b}{L^2}\sum_n^{\infty}\frac{\tau_n}{2} + T_s = \frac{8P_0 d\theta_b}{\pi^2}\sum_n^{\infty}\frac{1}{n^2} + T_s. \qquad (11)$$

But

$$\sum_{n=1,3,\cdots}^{\infty}\frac{1}{n^2} = \frac{\pi^2}{8}.$$

Therefore,

$$\overline{T}_j = P_0\theta_b d + T_s. \qquad (11a)$$

Thus, the average junction temperature is directly proportional to the power pulse amplitude and duty cycle as one would expect.

The maximum and minimum values of junction temperature can be found most readily by realizing when they occur relative to the excitation function. The maximum junction temperature must occur at the termination of the applied pulse, whereas the minimum must occur at the termination of the interval between pulses as illustrated in Fig. 3. From Fig. 3, the times for maximum and minimum temperature can be expressed as follows:

$t = r\tau_r + \delta/2$ for maximum temperatures, and
$t = r\tau_r - \delta/2$ for minimum temperatures
where r is any integer and δ = the pulse width = $d\tau_r$. Thus,

$$\omega_m t = 2\pi m/\tau_r(r\tau_r \pm \delta/2)$$
$$= 2\pi \pm m\pi d\begin{pmatrix}+\text{ for max temp}\\-\text{ for min temp}\end{pmatrix}. \qquad (12)$$

Substituting (12) into the second term of (10a) and rearranging, the expressions for maximum and minimum junction temperatures are obtained ($p = \tau_1/\tau_r$).

$$T_j\begin{pmatrix}+,\text{max}\\-,\text{min}\end{pmatrix} = P_0\theta_b d\cdot\frac{16}{\pi^2}\sum_{n,(\text{odd}),m}^{\infty}\frac{\sin m\pi d/m\pi d}{(2m\pi p)^2 + n^4}\cdot[n^2\cos(\pm m\pi d) + 2m\pi p\sin(\pm m\pi d)] + T_s. \qquad (13)$$

Eq. (13) has been evaluated for

$$\Delta T\begin{pmatrix}\text{max}\\\text{min}\end{pmatrix} = T_j\begin{pmatrix}\text{max}\\\text{min}\end{pmatrix} - T_s$$

for a number of values of both p and d.

The resulting values for

$$\Delta T_{\max}/P_0\theta_b \quad\text{and}\quad (1 - \Delta T_{\min}/P_0\theta_b)$$

are plotted in Fig. 4 as solid points. Through these calculated points can be drawn one or more straight lines (log-log plot) to approximate the theoretical results.

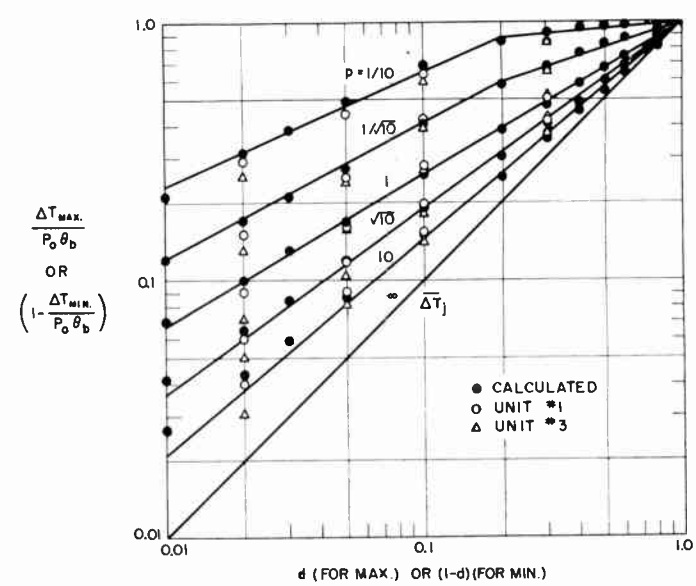

Fig. 4—Plot of normalized incremental junction temperature vs duty cycle with τ_1/τ_r the parameter for pulsed operation.

From Fig. 4, it can be seen as $p\to\infty$ (*i.e.*, $\tau_r \ll \tau_1$) the maximum and minimum values of temperature become the same and equal to the average value, *i.e.*, the frequency of excitation is so high that the junction temperature assumes a constant (average) value independent of time. As $p\to 0$ (*i.e.*, $\tau_r \gg \tau_1$) the other extreme is reached, *i.e.*, the frequency of excitation is so low that the junction temperature follows the wave shape of the transistor excitation. Between these two extremes lie values of p of practical importance, particularly when considering small duty cycle excitation. For example, consider a case where $p=1$ such as might exist if the repetition frequency were in the range 50 to 500 cps for most transistors ($\tau_1=2$ to 20 ms). Then, for $d=10$ per cent, $\Delta T_{max}/\overline{\Delta T}=2.5$, and, for $d=2$ per cent, $\Delta T_{max}/\overline{\Delta T}=4.7$. Therefore, if $T_s=n\overline{T}_j$ ($n=50$ to 80 per cent, typical values), then

$$T_j(\max)/\overline{T}_j$$

$$= n + (1 - n)\Delta T_{max}/\overline{\Delta T}$$

$$= \mathit{1.3}\ (n = 0.8, d = 0.1) \text{ or } \mathit{1.8}\ (n = 0.5, d = 0.1)$$

or

$$= \mathit{1.7}\ (n = 0.8, d = 0.02) \text{ or } \mathit{2.9}\ (n = 0.5, d = 0.02).$$

Thus, for this example, the maximum junction temperature could be from 30 to 190 per cent greater than the average junction temperature. If a smaller value of p was chosen, these ratios would be still larger; whereas, had a larger value of p been chosen they would, of course, be smaller. It may be concluded, then, that on the basis of these theoretical results and typical transistor thermal parameters, it is important to consider the maximum junction temperature rather than the average value when the duty cycle, d, is less than 20 per cent and $p=\tau_1/\tau_r$ is less than 10.

It is convenient in employing the results of this junction temperature analysis to consider the following approximate relationships. From the straight line approximations indicated in Fig. 4 (log-log plot), it can be shown that

$$\Delta T_{max} \cong C_1 P_0 \theta_b d^{(C_2 \ln (p) + 0.59)}.$$

For $1 \leqq p \leqq 10$, $d=0.01$ to 1.0,

$$C_1 = 1, \qquad C_2 = 0.11$$

and for $0.1 \leqq p < 1$

$$C_1 = (1 - 0.32 \ln (p)), \quad C_2 = 0.065 \ (d = 0.01 \text{ to } 0.2)$$

$$C_1 = 1, C_2 = 0.22\ (d = 0.2 \text{ to } 1.0). \tag{14a}$$

Furthermore, the minimum temperature increment can be expressed in terms of the maximum increment as follows:

$$\Delta T_{min}(d) = 1 - \Delta T_{max}(1 - d). \tag{14b}$$

B. Experimental Results

Experimental values of $\Delta T_{max}/P_0\theta_b$ and $\Delta T_{min}/P_0\theta_b$ were obtained for two of the grown junction units previously discussed. For simplicity in presentation and because of their greater accuracy, only the maximum values are plotted in Fig. 4.

As can be seen from Fig. 4, the agreement between theory and experiment for unit No. 1 (evacuated) is quite good except for very small duty cycles where the accuracy of measurement is poorer (± 20 per cent). Unit No. 3 (gas filled) is likewise in agreement for duty cycles in excess of 10 per cent, but below 10 per cent there is a definite fall-off in the experimental values as compared to theory. This fall-off in maximum junction temperatures for small duty cycles should be expected in view of the effect of the gas on slowing down the step response (discussed in Part II, Section B), particularly for small values of p. In fact, for $p=1$ ($\tau_r=\tau_1$) and $d=0.03$ ($\delta=0.03\ \tau_1$), the value for the maximum temperature is about 0.10 which agrees well with the step response value of 0.09 obtained for the same unit at $t/\tau_1=0.03$.

Since the effect of the gas in the transistor case, in addition to lowering the thermal resistance, is to smooth out the temperature fluctuations, it would appear that using gas under pressure or in the limit a liquid to increase the heat capacity of the transistor surroundings would be quite useful in reducing the maximum junction temperatures reached with small duty cycle pulses.

IV. Experimental Procedure

A brief description of the basic measuring approach and associated circuitry will now be presented, followed by a discussion of the actual measurements made.

A. Basic Measuring System

The basic approach used in making the junction temperature measurements consisted of monitoring the base current when the emitter was open-circuited (*i.e.*, I_{co}). By knowing the relationship between I_{co} and temperature, the junction temperature could be obtained.

The relationship between I_{co} and temperature was obtained by making collector current measurements (emitter open) for various collector voltages while the transistor was in an oven. Plotting I_{co} vs the equilibrium temperature provides the necessary calibration for a given transistor. It might be noted here that, in general, it was not found accurate enough to simply measure I_{co} at room temperature and then assume the theoretical relationship to obtain the complete calibration. Furthermore, if I_{co} is to be used for junction temperature measurements, it is of the greatest importance to insure that the calibration be completely reproducible—a situation which does not always exist.

The transistor test circuit used for measuring I_{co} was basically the same, except for some refinements, as discussed in a number of articles.[4,5] However, previous investigators were concerned only with the measurement of the average junction temperature rather than tem-

[4] J. Tellerman, "Measuring transistor temperature rise," *Electronics*, vol. 27, pp. 185–187; April, 1954.
[5] J. Ollendorf and R. E. Loofbourrow, "Equipment for measuring junction temperature of an operating transistor," *Transistors I*, RCA Labs.; March, 1956.

perature as a function of time. To observe the dynamic steady-state junction temperature response, it is necessary to insure that the measuring system response time be very much smaller than any thermal time constants involved. System response times of about 100 μs or less would be satisfactory. The basic circuit employed is shown in Fig. 5. The base circuit was selected for

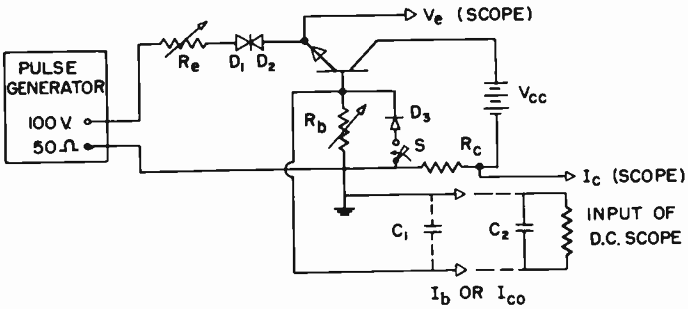

Fig. 5—Basic measuring circuit.

monitoring I_{co} (between pulses, with emitter "open") to avoid the much larger collector currents $I_c \gg I_b$ during the pulse interval. Even in comparison with I_b, I_{co} is quite small ($I_b/I_{co} \simeq 10^3$ or 10^4 for typical operation) thus requiring a sizable base resistor, R_b (=1 to 50 kilohms), for scope monitoring purposes. The diode, D_3, is a selected silicon unit used to prevent overdriving the scope by the I_bR_b voltage drop during the pulse. An over-driven scope can cause severe recovery transients which mask the I_{co} response; thus it becomes necessary to make the voltage $I_{co}R_b$ comparable to the clipped voltage pulse caused by I_b. To accomplish this, R_b can be increased to the point where the scope input just begins to appreciably shunt it and $R_b(C_1+C_2)$ is still well under 100 μs In addition to this RC time constant, the recovery times of the diodes and transistor used are important in determining the measuring system response time and, therefore must be kept small also. In the emitter circuit of the transistor under test is placed the driving source (a pulse generator), a current determining resistor, R_e, and two selected silicon diodes back-to-back. The purpose of the diodes (D_1 and D_2) is twofold: first, D_2 provides a very high impedance in the emitter circuit when the pulse is off and reverse or zero emitter voltages exist; and second, D_1 (in opposition to D_2) prevents the voltage drop $I_{co}R_b$ from biasing the emitter on while observing I_{co}. The pulse current is permitted to flow by having the applied voltage exceed the diode breakdown voltage of D_1 (about 4 volts) while the base voltage is insufficient to accomplish this. A small resistor, R_c, is included in the collector circuit for monitoring I_c, and, since I_b can be monitored by using R_b (switch, S, open), I_e can be obtained. The collector voltage, V_c, is determined by noting the base drop ($\sim$0.5 volt) and subtracting it from the collector supply voltage (I_cR_c is made negligible). Finally, the emitter voltage, V_e, is monitored by scope as indicated with $R_b = 0$.

B. Measurements Made

To verify the junction temperature step response discussed in Part II and to obtain the thermal parameters of the transistor under test, the following measurements and calculations were made. To observe the step response (in this case decay), the transistor was pulsed with a 50 per cent duty cycle (square wave) signal which produced a known junction dissipation. The repetition rate was initially started at a high rate ($p = \tau_1/\tau_r = 10$) and decreased until the I_{co} transient between pulses appeared to level off ($p \cong 1/10$). Under these operating conditions, it was presumed that the support temperature remained essentially constant (determined only by the average dissipation) and that the transistor bar went from one thermal equilibrium condition (pulse on) to another (no dissipation). From this observed transient, the maximum and minimum temperatures could be determined, while simultaneously monitoring the case and ambient temperatures. With these data the following calculations could be made with reference to Fig. 3.

$$\theta_t \text{ (total thermal resistance, junction to case)} = \frac{\overline{T}_j - T_c}{\overline{P}}$$

$$= \frac{[T_j(\max) + T_j(\min)]/2 - T_c}{\overline{P}}. \qquad (15)$$

(Note: Temperature drop equals the product of thermal resistance and power, $T = P\theta$.)

This value for θ_t must check with the value obtained by simply observing the average temperature at a high repetition rate ($p = 100$, no transient). Next, the thermal resistance of the bar can be obtained by two different calculations and checked against each other.

$$\theta_b \text{ (thermal resistance of the bar)} = \frac{T_j(\max) - T_s}{P_0}$$

$$\cong \frac{T_j(\max) - T_j(\min)}{P_0} \qquad (16a)$$

or

$$\theta_b = \theta_t - \theta_s = \theta_t - \frac{T_s}{\overline{P}} = \theta_t - \frac{T_j(\min)}{\overline{P}}, \qquad (16b)$$

where $T_s \cong T_j(\min)$ when $p \leq 1/10$, $d = 50$ per cent ($\Delta T_{\min} \cong 0$). The fundamental time constant, τ_1, can also be obtained from the I_{co} transient by noting how long it takes the junction temperature minus the support temperature ($T_j(\max) - T_s$ or $T_j(\max) - T_j(\min)$) to decay 70 per cent. The value of time constant obtained should check reasonably well with the calculated value and further should be such that $p \cong 1/10$, indicating that indeed the transient observed is essentially a step response. In general, the repetition rate should not be further decreased (p decreased beyond 1/10) if the supports are to be considered at a constant temperature (unless the entire junction to case transient is to be observed). Knowing τ_1, the response data presented in Fig. 2 were obtained by noting the existing temperature at the prescribed time intervals.

The experimental data for the maximum and minimum incremental junction temperatures, $\Delta T_{\max}$ and $\Delta T_{\min}$, plotted in Fig. 4 were obtained in a similar manner. A pulse signal of known period, τ_r (and thus known p) and duty cycle, d, was applied to the transistor and the maximum and minimum junction temperatures as well as the case temperature were determined. In addition, by noting the emitter and collector voltages and currents, the peak and average power dissipated could be calculated. Finally, using the following expressions the incremental temperatures were obtained.

$$\left.\begin{aligned}\Delta T_{\max} &= T_j(\max) - \overline{P}\theta_s - T_c \quad (17a)\\ \Delta T_{\min} &= T_j(\min) - \overline{P}\theta_s - T_c \quad (17b)\end{aligned}\right\} \text{(See Fig. 3)}$$

or for small values of duty cycle, $T_j(\min)$ for $(p=1/10) \cong (\overline{P}\theta_s - T_c)$ was used for greater relative accuracy. The values of $\Delta T_{\max}/P_0\theta_b$ in Fig. 4 are only accurate to within about ± 20 per cent at $d=0.02$ and 0.05 because of the difficulty in accurately determining P_0. At such low duty cycles with the transistors used it was difficult to avoid or accurately correct for the distributed dissipation due to the IR drop in the transistor bar. The accuracy in determining $\Delta T_{\min}/P_0\theta_b$ (not plotted) was even poorer (± 30 per cent) primarily because the difference between two large numbers is used to obtain $\Delta T_{\min}$.

V. Application of Results to Pulsed Transistor Operation

In applying the results obtained to pulsed transistor operation, three factors must be considered: maximum junction temperature rating, thermal stability, and electrical transistor parameter changes. The first two factors are directly concerned with the permissible thermal operation of the transistor, while the third factor is involved in the electrical response of the stage.

A. Consideration of the Maximum Junction Temperature Rating

Following current practice, a maximum junction temperature rating is being specified for transistors, which represents the maximum operating temperature for which satisfactory life can be expected. To comply with this rating the maximum junction temperature expected for a given transistor operation must be determined. Consider the following case of transistor pulsed operation.

Given:

$f_r = 100$ cps
$d = 3$ per cent
$I_c = 200$ ma
$V_c = 18$ v (pulse on or off)
$V_e = 0.5$ v
$I_{co} = 20\ \mu$a at 25°C
$\alpha_{ec} = 0.95$ (current gain)
$\theta_b = 145$°C/w
$\theta_s = 120$°C/w
$\tau_1 = 10$ ms
$\theta_{case} = 30$°C/w (assumed constant)
$T_a = 20$°C

Under these conditions, the maximum junction temperature will now be obtained.

$$\begin{aligned}P_0 &= V_cI_c + V_eI_e \text{ (peak power dissipated)}\\ &= 3600 + 105 = 3705 \text{ mw}\end{aligned}$$

and

$$\overline{P} = dP_0 = 111 \text{ mw.}$$

Since

$$T_j(\max) = \Delta T_{\max} + T_s = \Delta T_{\max} + \overline{P}\theta_s + \overline{P}\theta_c + T_a$$

where

$$\overline{P}\theta_s = 13.3\text{°C},\quad \overline{P}\theta_c = 3.3\text{°C, and } T_a = 20.0\text{°C},$$

it remains to determine $\Delta T_{\max}$. Recognizing that $p = \tau_1/\tau_r = 1$ and $d = 0.03$, the appropriate expression (14a) can be used to obtain $\Delta T_{\max}$. Thus,

$$\begin{aligned}\Delta T_{\max} &= P_0\theta_b d^{[0.11 \ln (p)+0.60]}\\ &= 536(0.03)^{(0.60)} = 536\cdot 0.123 = 65.9\text{°C.}\end{aligned}$$

Therefore,

$$T_j(\max) = 65.9 + 13.3 + 3.3 + 20.0 = 102.5\text{°C.}$$

Whereas

$$\overline{T}_j = \overline{\Delta T} + T_s = \overline{P}\theta_b + T_s = 16.1 + 36.6 = 52.7\text{°C.}$$

Two corrections to the above results should be considered before accepting them as final values. First, the average support temperature, T_s, should be corrected if the added average dissipation caused by the increase in $I_{co}(\overline{T}_j)$ is sufficient to warrant it. Second, $\Delta T_{\max}$ must also be corrected if the increase in peak dissipation caused by $I_{co}(T_j(\max))$ warrants it.

In the example under consideration, $I_{co}(\overline{T}_j) = 126\ \mu$a and $I_{co}(T_j(\max)) = 3.51$ ma as calculated from $I_{co}(25\text{°C}) = 20\ \mu$a and (18). If the stability factor[6] $S = \partial I_c/\partial I_{co} = 1$ as it would for a simple grounded base stage, then the incremental powers become

$$\begin{aligned}\overline{\Delta P} &= SV_cI_{co}(\overline{T}_j) = 2.3 \text{ mw}\\ \Delta P(\max) &= SV_cI_{co}(T_j(\max)) = 63.2 \text{ mw.}\end{aligned}$$

Since these incremental powers are small compared to the average and peak values of dissipation in this case, no correction to the calculated temperatures is necessary. However, for this case, if $S = 20$, as it would for the simple grounded emitter stage ($S \cong \alpha_{bc}$), corrections would definitely be required.

[6] See R. F. Shea, *et al.*, "Principles of Transistor Circuits," John Wiley and Sons, Inc., New York, N. Y.; 1953.

Thus, for the operation considered, the maximum junction temperature would be approximately 103°C. If the permitted junction temperature (maximum rating) were 110°C., this operation would be permissible with regard to this criterion (maximum junction temperature rating).

It is interesting to note here that had the simple exponential response been used to obtain $T_j(\max)$ for the repetitive pulse, $\Delta T_{\max}$, in this example, would have been only 25.1°C. as compared to 65.9°C.; $T_j(\max)$ would have been only 61.7°C. as compared to 102.5°C.

B. Consideration of Thermal Stability

In addition to comparing the maximum operating junction temperature expected to the maximum rating, it is also necessary to determine if thermal stability will prevail before passing on the proposed operation. The system is said to be thermally stable if the increase in dissipation due to a temperature increase is less than the corresponding increase in heat flow from the junction. This condition for dc steady-state can be stated as follows:

$$I_{co}(\overline{T}_j) < \frac{15}{S\overline{V}_c(\theta_t + \theta_c)} \tag{17}$$

where $(\theta_t+\theta_c)$ is the total thermal resistance from junction to ambient.[7] This expression assumes that I_{co} is related to temperature by

$$I_{co} = I_0 e^{T/15}, \tag{18}$$

which is sufficiently accurate for the temperature range of interest, and that the gain is constant. However, where this criterion for thermal stability can be applied to the average transistor excitation (dc component) and temperature, an additional criterion must be developed for the pulsed operation.

Consider the change in junction temperature, $\Delta T_j'$, caused by an incremental change in dissipation, ΔP, which is brought about by a change in I_{co} with an assumed incremental change in junction temperature, ΔT_j (gain assumed constant). If thermal stability is to be maintained, then

$$\Delta T_j' \text{ caused by } \Delta P < \Delta T_j \text{ causing } \Delta P. \tag{19}$$

At t_1, if a ΔT_j is assumed which corresponds to the rise in temperature during Δt caused by $P(t)$, then ΔP can be calculated as follows: (See Fig. 6)

$$\Delta P = V_c \Delta I_c = V_c S \Delta I_{co} \text{ (assuming } V_c = \text{const).} \tag{20a}$$

But,

$$\Delta I_{co} = I_{co} \Delta T_j/15 \text{ from (18).}$$

Thus,

$$\Delta P = V_c S I_{co}(T_j) \Delta T_j/15. \tag{20b}$$

[7] J. S. Saby, "Transistors for high power application," 1954 IRE Convention Record, part 3, pp. 80–83.

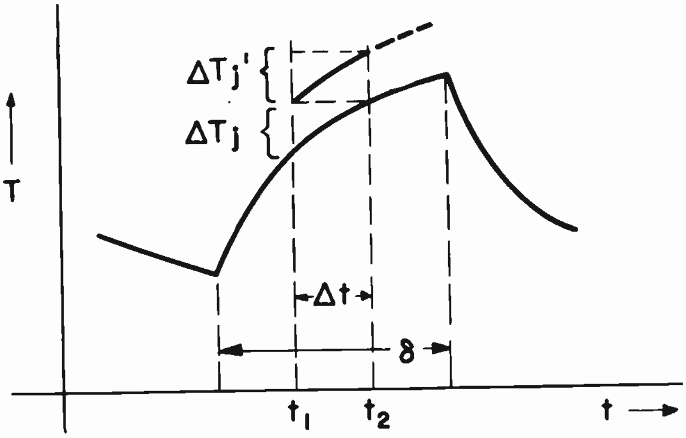

Fig. 6—Temperature vs time considering thermal stability.

The $\Delta T_j'$ produced by ΔP can be calculated by using the approximate step response, (6). This expression, which can be considered applicable for $d<0.3$ and $p>1/10$ $(t/\tau_1<2)$, actually predicts temperatures in excess of the true values and will thus give a more stringent stability criterion by as much as a factor of 1.4 for large values of d. Thus,

$$\Delta T_j' = \Delta P \theta_b \sqrt{\Delta t/\tau_1} \cdot y\sqrt{2}$$

or

$$\Delta T_j' = V_c S I_{co}(T_j) \theta_b \sqrt{\Delta t/\tau_1} \cdot \Delta T_j/21. \tag{21a}$$

If now Δt is assumed to approach δ, and I_{co} is evaluated at $T_j(\max)$, the worst situation, then (21a) becomes

$$\Delta T_j' = V_c S I_{co}[T_j(\max)] \theta_b \sqrt{d/p} \Delta T_j/21, \tag{21b}$$

and the second criterion for thermal stability becomes

$$I_{co}[T_j(\max)] < \frac{21}{S V_c \theta_b \sqrt{d/p}}. \tag{22}$$

Note that the additive effect of ΔP for many cycles can modify the average temperature which is taken into account in (22) through the evaluation of I_{co} at the corrected maximum junction temperature as discussed in Section A.

Thus, two conditions for thermal stability must be met when considering pulsed operation; first, (17) for the average condition which involves the total thermal path; and second, (22) for the maximum condition which involves the thermal resistance of the bar only and d/p for the chosen operation. Applying (17) to the example cited in Section A, the condition,

$$I_{co}(\overline{T}_j) = 1.26 \cdot 10^{-4} < \frac{15}{S\overline{V}_c(\theta_t + \theta_c)}$$

$$= \frac{15}{20 \cdot 18 \cdot 295} = 1.41 \cdot 10^{-4},$$

is satisfied in this case even with $S=20$. However, applying

$$I_{co}[T_j(\max)] = 3.51\cdot 10^{-3} < \frac{21}{SV_c\theta_b\sqrt{d/p}}$$

$$= \frac{21}{20\cdot 18\cdot 145\cdot 0.173} = 2.33\cdot 10^{-3}$$

is not satisfied with $S=20$. Thus the pulsed operation proposed in the example is completely permissible thermally only if S is made less than 13.

It should be remarked here that the above calculations are made assuming S constant. However, since S depends on α_{ec} or α_{bc} which can change with temperature, it may be necessary to correct the calculations accordingly. Of course, if α_{ec} changes too rapidly with temperature so that either $d(1-\alpha_{ec})/dT$ or $d\alpha_{bc}/dT$ are comparable with dI_{co}/dT, then the change in I_c with the total base current, I_b, (not just I_{co}) must be considered and (17) and (22) are invalid.

C. Consideration of Parameter Changes

In designing pulse circuits, it can often be helpful to at least know the limiting parameter values so that appropriate calculations of gain and terminal impedances can be made. If the various electrical parameters are known as a function of temperature (sometimes supplied by the manufacturer), a determination of the limiting values (at least with respect to temperature) can be obtained by noting the values of $T_j(\max)$ and $T_j(\min)$. How the parameters vary with excitation between these limiting temperatures would require further knowledge of the variation of junction temperature with time [evaluation of (13) for a number of points] which would, in general, be too laborious to be of value.

VI. Conclusion

Both the temperature step response and the pulse excitation response which have been derived from an assumed thermal model have been essentially substantiated by experiment. Thus, all the analytical results, including the approximate expressions for maximum and minimum temperatures, are directly applicable to the problem of predicting transistor junction temperatures where the one-dimensional heat flow model exists. Furthermore, utilizing the verified step or impulse response, the junction temperature response can be obtained for other periodic excitations (in addition to the repetitive pulse) following the procedure used here.

With the information gained from this study, the circuit designer should be better able to determine the satisfactory operation of a transistor stage for pulse excitation from both thermal and electrical considerations.

Finally, this study indicates that, with the current procedure of reducing the thermal resistance of the internal transistor supports (θ_s) to increase the device dissipation, the problem of junction temperature variations, particularly in switching circuits, may (assuming θ_b is kept constant) be even more important in the future.

VII. Acknowledgment

The author wishes to express his thanks to C. V. Jakowatz and J. E. Taylor for many helpful discussions on the heat flow problem. Further, the author is grateful to R. L. Pritchard and W. N. Coffey for their critical review of this paper and for their many constructive suggestions. All experimental transistor units were supplied by R. E. Shepp.

VIII. Appendix

The solution to the boundary value problem, (1) and (2), can be obtained using the classical procedure or by use of the Laplace transform. The classical procedure will be outlined here.

Given:

$$\frac{\partial T}{\partial t}(x, t) = \alpha \frac{\partial^2 T}{\partial X^2}, \qquad 0 \leqq x \leqq L,\ t > 0, \tag{1}$$

and

$$\left.\begin{aligned} \frac{\partial T}{\partial t}(0, t) &= -\frac{F}{k} \quad &(2a)\\ T(L, t) &= T_s &(2b)\\ T(x, 0) &= T_s &(2c)\end{aligned}\right\}$$

Solve for $T(x, t)$, $0 \leqq x \leqq L$, $t>0$.

Since (2a) and (2b) are nonhomogeneous boundary conditions a change in variable is made so that they become homogeneous.

Let $T(x, t) = G(x, t) + F/k(L-x) + T_s$ so that

$$\frac{\partial G(x, t)}{\partial t} = \alpha \frac{\partial^2 G}{\partial x^2}, \quad 0 \leqq x \leqq L, \quad t > 0 \tag{1a}$$

and

$$\left.\begin{aligned} \frac{\partial G}{\partial t}(0, t) &= 0 \quad &(2a')\\ G(L, t) &= 0 &(2b')\\ G(x, 0) &= \frac{F}{k}(x - L) &(2c')\end{aligned}\right\}.$$

Treating (1a) by the method of separation of variables, two equations are obtained.

$$X''(x) + aX = 0 \tag{23}$$

$$T'(t) + a\alpha T = 0. \tag{24}$$

The solution of (23) which also satisfies the boundary conditions is

$$X(x) = \cos \frac{n\pi}{2L} x, \qquad n = 1, 3, \cdots, \tag{25}$$

and the solution of (24) is

$$T(t) = e^{-(n^2\pi^2/4L^2)\alpha t}. \tag{26}$$

Thus the total solution becomes the product of (25) and (26) or

$$G(x, t) = \sum_{n=1,3,\cdots}^{\infty} B_n e^{-(n^2\pi^2/4L^2)\alpha t} \cos \frac{n\pi}{2L} x \tag{27}$$

where

$$B_n = 2/L \int_0^L G(x, 0) \cos \frac{n\pi}{2L} x dx = -\frac{8FL}{kn^2\pi^2}. \tag{28}$$

Substituting (28) into (27) and changing back to the original variable, $T(x, t)$, (3), representing the complete solution is obtained.

$$T(x, t) = T_s + \frac{F}{k}(L - x) - \frac{8FL}{k\pi^2} \sum_{n=1,3,\cdots}^{\infty} \frac{e^{-(n^2\pi^2/4L^2)\alpha t}}{n^2} \cos \frac{n\pi}{2L} x. \tag{3}$$

Shutter Image Converter Tubes*

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Summary—**This paper discusses recently developed shutter image converter tubes. Part I is concerned with electrostatically focused tubes while part II considers magnetically focused tubes. Both types employ a mesh spaced close to the cathode which can be used to control the passage of photoelectrons from cathode to phosphor layer (anode). The electron-optical theory is presented for both types of tubes and their static characteristics are discussed.**

Introduction

IN RECENT TIMES image converters have found many new applications. Besides uses as snooperscopes[1] and light amplifiers, they have been used widely in the field of ultraspeed photography.[2] In this latter field the tubes have been mainly magnetic focus image converters and the image was pulsed on and off the screen by pulsing the total accelerating voltage (5 to 12 kv) on and off the tube.

The purpose of the present paper is to describe two types of image converters which were investigated in these laboratories. Both types incorporate a mesh or grid spaced closely to the cathode so that the image on the phosphor screen may be pulsed on and off with a low voltage pulse rather than having to pulse the overall voltage. Part I of this paper describes an electrostatically focused tube while part II describes the work done on magnetically focused image converters. In both parts only the static characteristics of the tubes are considered.

I. An Electrostatically Focused Shutter Image Converter

The electrostatic image converter investigated in these laboratories was based on a concept of Schagen, *et al.*[3] These workers took advantage of the fact that the motion of an electron in an electrostatic field whose intensity varies as an inverse square law from some central point is relatively easy to calculate. This is one of the few cases in electron optics where the equations of motion may be solved in a closed form. The system used by Schagen, *et al.* is that of two concentric spheres at different potentials, *i.e.*, the photocathode and anode spheres of Fig. 1. Electrons released from a point on the inside surface of the outer sphere are accelerated toward the more positive inner sphere and the paths are such as to form a first order focus at some point beyond the center of the system. One of the difficulties in constructing a two-electrode tube is that the position of the focal plane (or in this case one may say more accurately, spherical focal surface) depends only on the geometry of the tube. Thus, if the tube is not build to the exact proportions, the image is not at its best focus and nothing can be done to improve it.

By inserting a highly transparent mesh of spherical form between cathode and anode spheres in such a way as not to disturb the spherical geometry of the system, one has a means of cutting the electron flow on and off just as in an ordinary triode radio tube. Not only does the mesh act as a shutter electrode, but it is also a focus-

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[1] G. A. Morton and L. E. Flory, "An infra-red image tube and its military applications," *RCA Rev.*, vol. 7, pp. 385–413; September, 1946

[2] A. W. Hogan, "Use of image converter tubes for high-speed shutter action," PROC. IRE, vol. 39, pp. 268–270; March, 1951.

[3] P. Schagen, H. Bruining, and J. C. Francken, "A simple electrostatic electron-optical system with only one voltage," *Phillips Res. Rep.*, vol. 7, pp. 119–130; April, 1952.

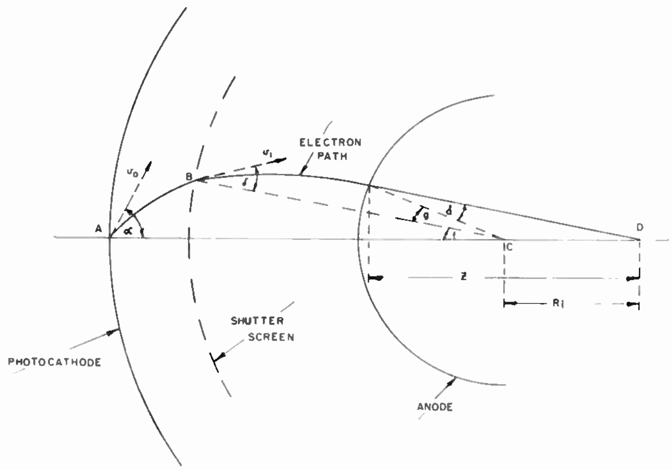

Fig. 1—The basic focusing system for the electrostatic image converter.

ing electrode. Thus the dimensions of the tube can vary over quite a large range and there will still be a mesh voltage for which the image is in optimum focus.

A. Theory

The theory presented by Schagen, *et al.*, for the two-electrode case will now be extended to include three electrodes (cathode, anode, and shutter screen). To do this one merely applies the two-electrode theory in two steps.

1) The first step considers the cathode-screen system only and makes use of the formulas for the two electrode case. In this way the point B and the angles δ and f of Fig. 1 are calculated.
2) The second step considers the screen as the cathode which emits an electron with velocity v_1 at an angle δ with respect to the normal to the screen. The theory for the two-electrode case is then applied again and the angles g and d of Fig. 1 are then obtained.

These two steps are sufficient to determine the radius, R_i, of the virtual image focal surface; *i.e.*,

$$R_i = R_a \frac{d}{g + f - d}. \tag{1}$$

Having found R_i it will then be necessary to take into account the diverging lens effect of the anode aperture, (see Fig. 8) in order to get the radius of the real image focal surface.

In developing the theory of the three-electrode tube all angles except the initial angle of emission, α, involved in the calculation will be considered so small that the sines and tangents of the angles may be replaced by the angles themselves. This is standard procedure in calculating out the first order theory of an electron optical system. This first order theory then gives the Gaussian focal point which will be independent of the angle of emission, α, and initial velocity, v_0. Higher order theories will show a dependence of the focal point on α (spherical aberration) and on v_0 (chromatic aberration). In experimental tubes so far constructed, the size of the phosphor grains has been the main limitation on resolution and these particular geometrical aberrations have not been observed in the tube.

The notation to be used in the following sections is explained in Fig. 1 and Table I.

TABLE I

NOTATION FOR THE THEORY OF THE ELECTROSTATIC IMAGE CONVERTER

R_a = anode radius.
R_c = cathode radius.
R_s = mesh radius.
R_i = radius of the virtual image focal surface.
$n = R_c/R_a$.
$y = R_c/R_s$.
$x = R_s/R_a$.
ρ = the magnitude of the radius vector from the center of curvature of the focusing system to the electron.
ϕ = the angle between the radius vector to the electron and the axis of symmetry of the system.
$\mu = R_c/\rho$.
z = the virtual object distance from the anode aperture lens.
z' = the virtual image distance from the anode aperture lens.
s = the real image distance from the anode aperture lens.
F = the focal length of the anode aperture lens.
v_0 = the initial velocity of emission of the photoelectron.
v_1 = the velocity of the photoelectron on reaching the mesh.
m = the mass of the electron.
e = the charge of the electron.
E_0 = the initial energy of emission of the photoelectron.
E_1 = the energy of the photoelectron on reaching the mesh.
$\vec{\mathcal{E}}(\vec{\rho})$ = the electric field vector at the point defined by the radius vector $\vec{\rho}$.
Φ_ρ = the potential at a sphere of radius $\rho (R_c \geq \rho \geq R_a)$.
Φ_a' = the potential of the anode relative to the cathode.
Φ_a = the potential of the anode relative to the mesh.
Φ_s = the potential of the mesh relative to the cathode.
Φ_c = the potential at the center of the anode aperture relative to the cathode.
$\Phi_0 = \Phi_s/(y-1)$.
$\Phi_1 = \Phi_a/(x-1)$.
$r^2 = \Phi_a'/\Phi_s$.
$q = e\Phi_0/E_0$.
$t = e\Phi_1/E_1$.
$p = e\Phi_a'/[E_0(n-1)]$.

1) *The Virtual Image Position:* The mathematical derivation of the position of the virtual image surface has been relegated to Appendix I. There it is shown that

$$R_i = R_a \frac{n(r+1)}{r(n-2) + 2(y-1)r^2 - n}. \tag{2}$$

It is also shown in Appendix I that for the case where the mesh potential is such that the three-electrode case degenerates to the two-electrode case, (2) above reduces to

$$R_i = R_a \frac{n}{(n-2)}. \tag{3}$$

This is just the expression for the two electrode case derived by Schagen, *et al.*

It will now be noticed that for the two-electrode case, the virtual image focal surface is determined only by the geometrical parameters. Thus, if the tube is not built to exactly the correct dimensions, the image will be out of focus with no means for adjustment. On the other hand, (2) for the three-electrode case involves the ratio of mesh to anode voltage. Thus, one can obtain a focus over a wide range of geometrical parameters by varying this voltage ratio. Once the geometrical parameters are fixed in the tube, however, there is only one voltage ratio which gives a focused image.

2) *Magnification:* The magnification of the virtual image with respect to the object is given by

$$M' = \frac{R_i}{R_c}. \tag{4}$$

The real image distance, however, is somewhat larger than the virtual image distance due to the fact that the aperture in the anode sphere acts as a diverging lens. If z' and s are the distances of the virtual and real images, respectively, from the anode aperture then s/z' is the magnification of the real image with respect to the virtual image. Thus, the magnification of the system as a whole is given by

$$M = \frac{sR_i}{z'R_c}. \tag{5}$$

Assuming that the thin lens formula holds for the anode aperture lens, one may write

$$\frac{1}{s} + \frac{1}{z} = \frac{1}{F}. \tag{6}$$

Then (5) becomes

$$M = \frac{R_iF}{R_c(z - F)}. \tag{7}$$

It should be pointed out that F is negative because the lens is divergent and z is negative $(= -z')$ because the virtual object is in the image space.

3) *Focal Length of the Anode Aperture Lens:* In the article by Schagen *et al.* the Davisson-Calbick formula is used for the focal length of the anode aperture lens. This formula is for an aperture in an infinite plane and as such is not quite applicable here. The formula for the focal length of an aperture in the inner sphere of a system consisting of two concentric spheres is derived in Appendix II. This derivation gives the following

$$F = -4R_a \frac{(x - 1)\Phi_c}{x\Phi_a}. \tag{8}$$

Taking $\Phi_c \simeq \Phi_a'$, we have

$$F = -4R_a \frac{r^2(x - 1)}{(r^2 - 1)x}. \tag{9}$$

In experimental tubes built in these laboratories, the values of F calculated from measurements on the tubes correspond within about 10 per cent to the value given by (9). The deviation between the measured and calculated values of F are probably due to the various assumptions involved in the derivation of (9).

B. Procedure for Tube Design

In designing an image converter, one is usually interested in designing for a particular magnification, M, and voltage ratio, r^2. The size of the cathode required determines R_c, and the resolution required will determine how close one can place the shutter mesh to the cathode, *i.e.*, R_s is also determined. If R_s extends too close to the cathode, the shutter mesh structure will be apparent in the image and cause a deterioration of resolution. One then chooses convenient values of anode radius, R_a, and anode aperture diameter, D. From these choices the value of R_i may be calculated from (2) and the distance, z, of the virtual image from the anode aperture can then be obtained. The value of F may be estimated from (9) and the distance, s, of the real image from the aperture plane may be calculated with the use of (6). The value of M may then be found from (7). If this value of M corresponds closely to the desired value, then the initial choice of geometrical parameters is correct and the tube is built according to those values. If the calculated value of M is not correct, then one of the parameters must be changed (usually R_a is most convenient) and the calculation repeated. Fig. 2 shows that

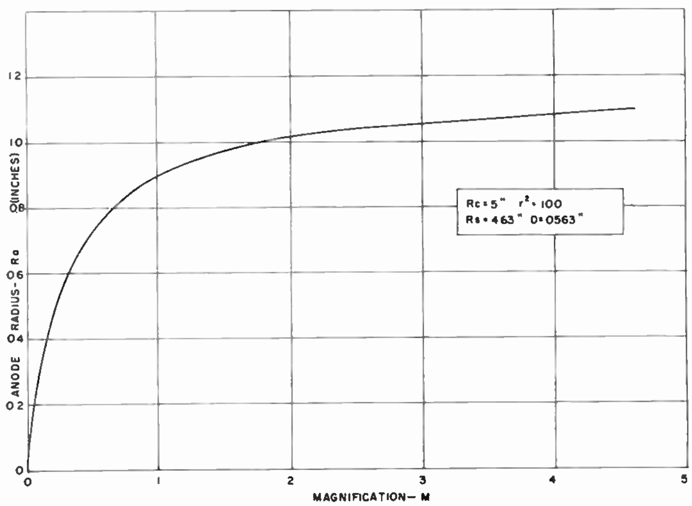

Fig. 2—Calculated variation of anode radius, R_a, with magnification, M.

the magnification is a monotonically increasing function of the anode radius. Thus, if the value of M calculated for a given R_a is too small, one may increase R_a in the subsequent calculations to increase the value of M. Figs. 3 through 7 illustrate how the various parameters vary among themselves for certain fixed parameters.

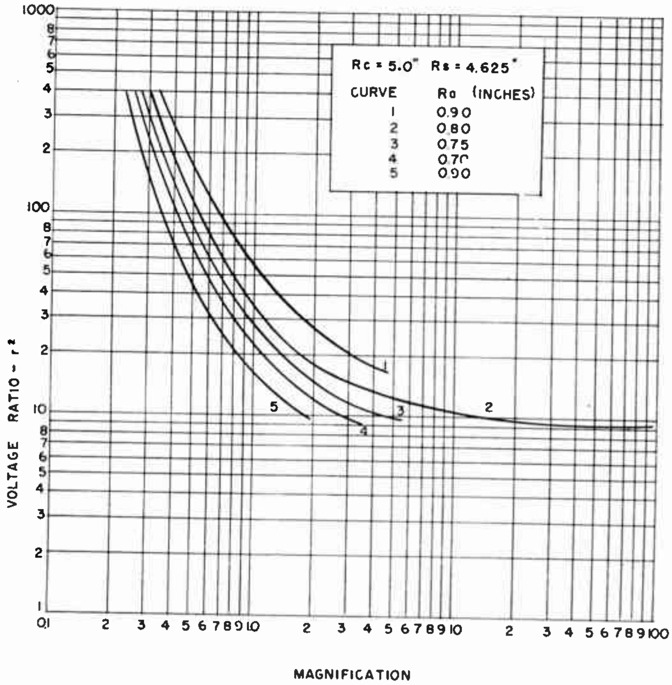

Fig. 3—Calculated variation of voltage ratio, r^2, with magnification, M.

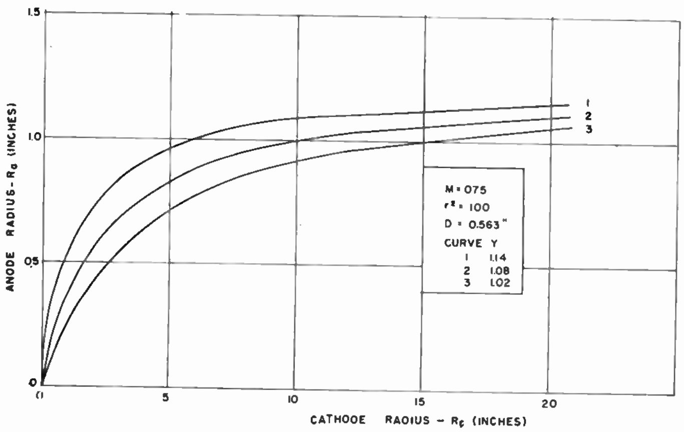

Fig. 4—Calculated variation of anode radius, R_a, with cathode radius, R_c.

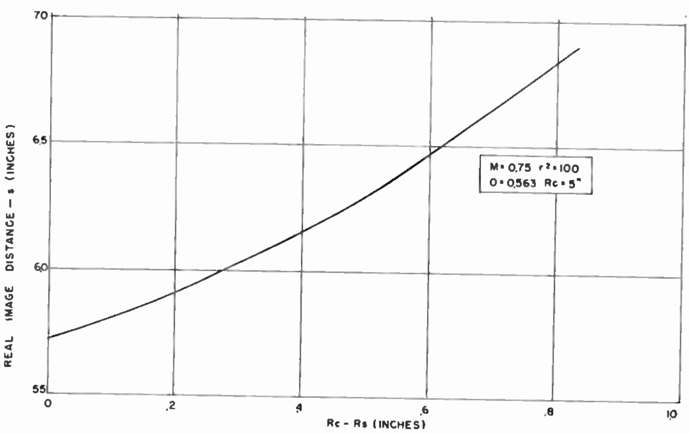

Fig. 5—Calculated variation of real image distance, s, with distance between cathode and screen, $R_c - R_s$.

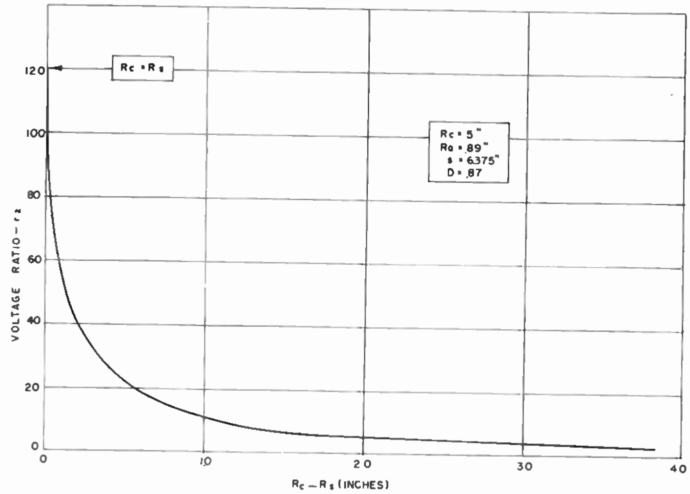

Fig. 6—Calculated variation of voltage ratio, r^2, with distance between cathode and screen, $R_c - R_s$.

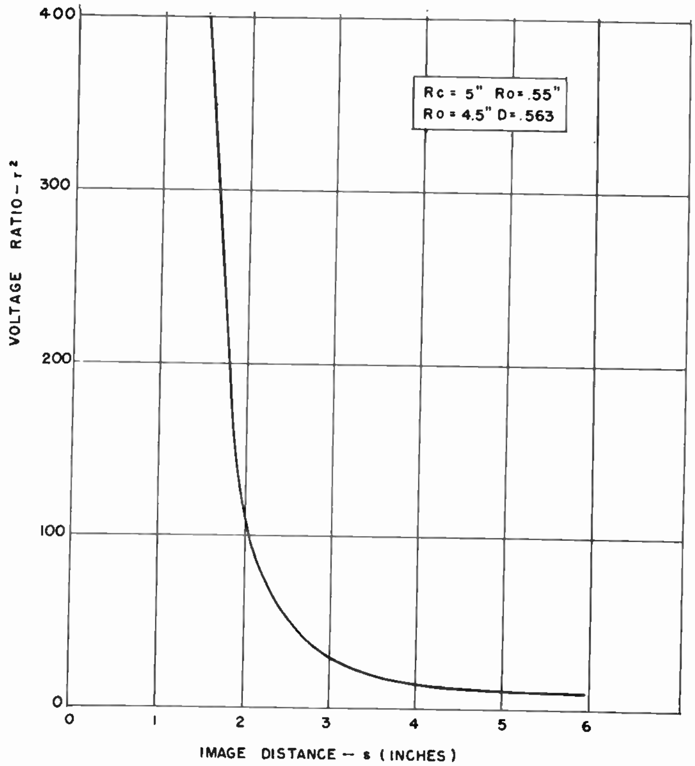

Fig. 7—Calculated variation of voltage ratio, r^2, with real image distance, s.

C. Experimental Results

A number of these electrostatic image converters have been built and tested at these laboratories. Fig. 8 is a sketch of the tube showing the essential features. The photocathode is deposited on a transparent conducting layer which has been formed on the inside glass surface. The reason for this is that otherwise high intensity light flashes may cause a resistance drop along the cathode so that the cathode surface would not be an equipotential surface. The transparent conductive coating pre-

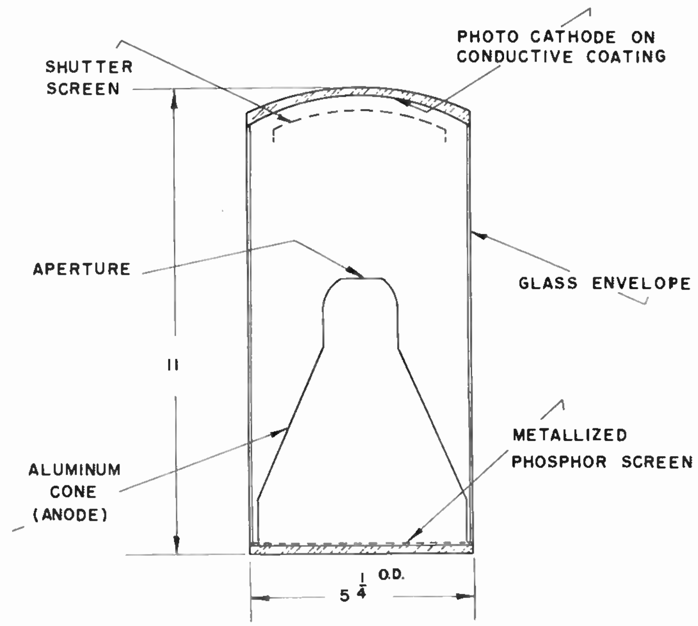

Fig. 8—Sketch of the electrostatic image converter.

vents this effect. The phosphor screen is aluminized since the tube is usually run at between 9 to 15 kv and the aluminum backing enhances the contrast and brightness. The aluminum backing serves two other important purposes, namely, it eliminates light feedback from the phosphor to the photocathode and it prevents light which may pass through the photocathode from being observed by the viewer. Although the anode cone will block most of the light that passes between cathode and phosphor, the anode aperture (see Fig. 8) still provides a direct optical path.

One very desirable feature of this tube type is the fact that the anode cone shields the phosphor screen from any stray emission effects. These effects usually show up as bright patches on the viewing screen and often limit the over-all accelerating voltage at which the image converter may be run. Fig. 9 is a photograph of an electrostatic image converter built at these laboratories.

Fig. 10 is a photograph of a television resolution chart observed on one of the tubes. The best resolution obtained to date is 6 optical lines per millimeter on the phosphor screen. In all tubes built so far, the resolution has been limited by the grain size of the phosphor. No aberration effects have been large enough to noticeably affect the resolution, except at the very edge of the screen. In tubes built so far, the viewing screen has had a diameter of 4 inches.

One type of observable aberration is pincushioning. This aberration will theoretically not affect the resolution, however. The amount of pincushioning depends on the design of the tube. Those tubes with smaller anode radii have less pincushioning than tubes with larger anode radii.

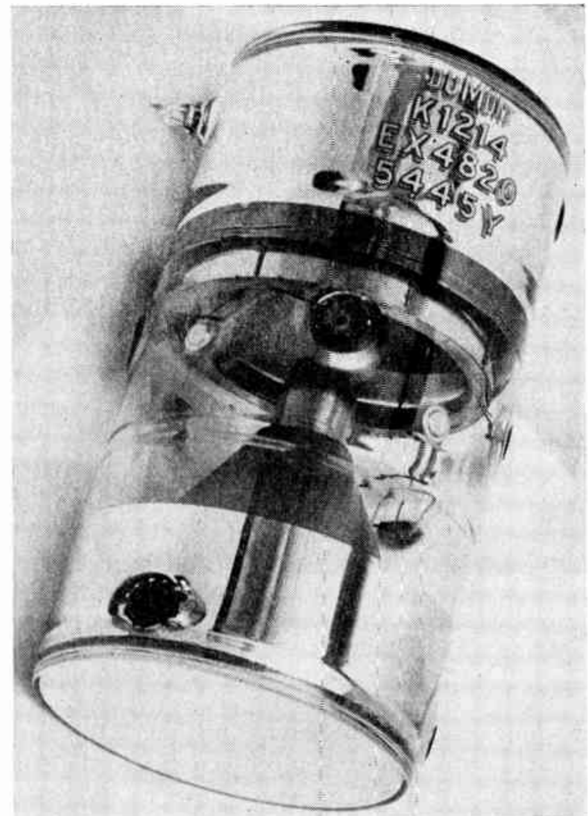

Fig. 9—Photograph of an electrostatic image converter.

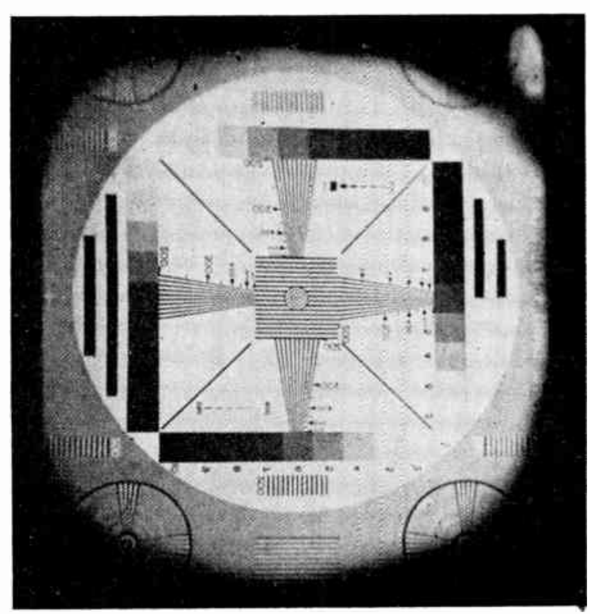

Fig. 10—Photograph of a test pattern from the electrostatic image converter. Note that the focusing screen is definitely in the image at the edge of the field although it has little or no effect at the center.

This is probably due to the fact that the crossover of the electron beam is closer to the anode aperture lens for smaller anodes. In a sense this is unfortunate since one way of designing for lower mesh-cathode voltage is to increase the anode radius.

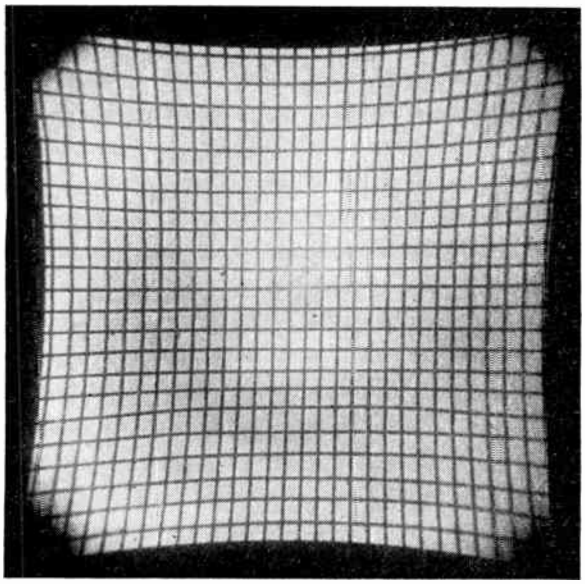

Fig. 11—Photograph from the electrostatic image converter illustrating the degree of pincushioning. With correct design this pincushioning can be eliminated as shown in Fig. 10.

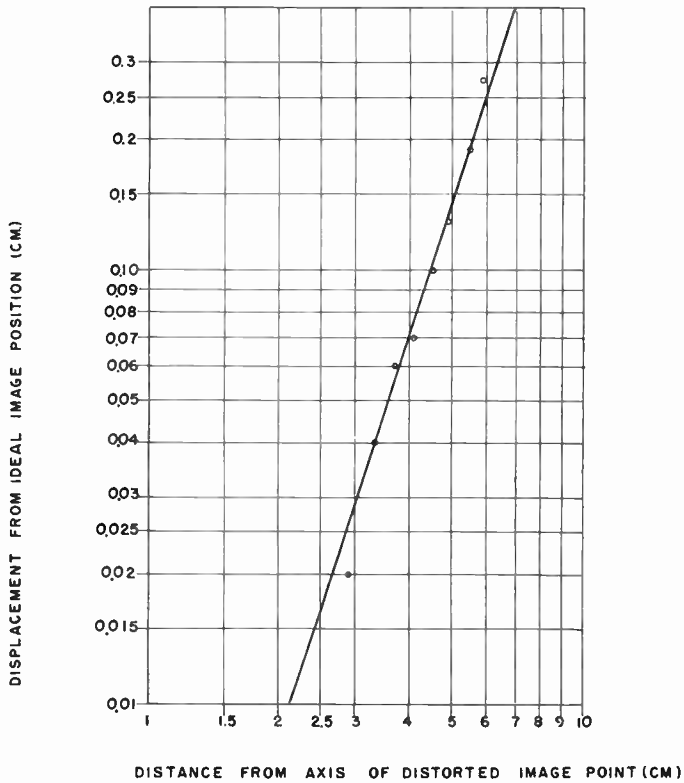

Fig. 12—Empirical distortion characteristic of the tube used for the photograph in Fig. 11.

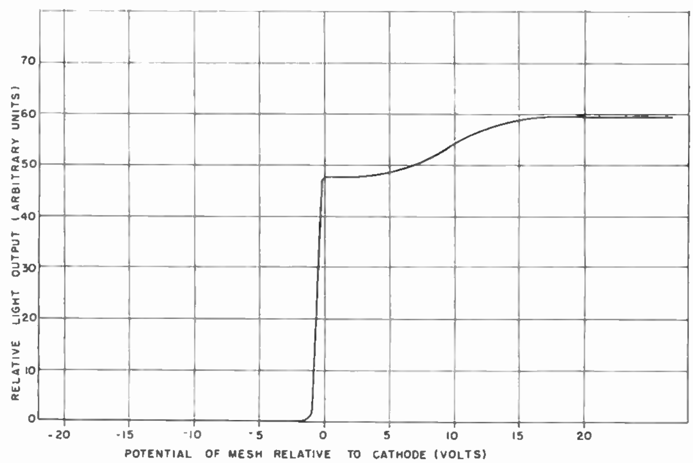

Fig. 13—Variation of light out from the anode as a function of mesh voltage. The over-all voltage applied was 10,000 volts.

This pincushioning is shown in Fig. 11 where a crossed-grid pattern has been projected onto the photocathode. Two other effects which tend to increase the pincushioning are the fact that a flat viewing screen is used rather than a spherical one. This latter effect is small, however, since tubes relatively free of this aberration have been built as seen in Fig. 10.

The theory of lens aberrations[4] shows that, for pincushioning in a cylindrically symmetric system, the displacement of an actual image point from its position for an undistorted image is proportional to the cube of the distance from the axis of symmetry of the system. As shown in Fig. 12, the actual dependence is the 3.42 power rather than the 3rd power. This may be due to the above mentioned effect of the flat viewing screen.

It is usually of interest to design a tube to have low cathode-to-mesh voltage for easy on-off pulsing of the image. By varying the geometrical parameters, it is possible to design for almost any mesh voltage. One may even design to have the tube work with an over-all voltage of 10,000 volts while a pulse of 10 volts will cut off the photoelectron beam. Tubes built so far have had mesh voltages between 100 to 300 volts for 10,000 volts over-all on the tube.

The cutoff characteristic of the focusing mesh is shown in Fig. 13.

II. Shutter Image Converter Tubes with Magnetic Focusing

A. Introduction

Part I of this paper discussed a shutter image converter tube with electrostatic focusing. In the general investigation of shutter tubes work was also done on magnetic focus image converters with a fine mesh spaced close to the cathode. This mesh then allowed pulsing the tube on and off with a low voltage pulse rather than pulsing the total voltage on the tube. The only tubes considered were those that use a long solenoid for focusing. Table II is a list of symbols used in part II of this paper.

B. Theory

A sketch of the system is shown in Fig. 14. To have a focus it is necessary that the time, T, required for the

[4] V. E. Cosslett, "Introduction to Electron Optics," Oxford University Press, New York, N. Y., pp. 114–115; 1946.

TABLE II

NOTATION FOR THE THEORY OF THE MAGNETIC IMAGE CONVERTER

H_z = the axial component of the magnetic field.
V_m = the voltage of mesh relative to cathode.
V_s = the voltage of screen (anode) relative to mesh.
d = the distance between cathode and mesh.
L = the distance between mesh and screen.
n = the order of focus.
β = the angle of electron velocity relative to axis at the mesh.
T_1 = the transit time of an electron from cathode to mesh.
T_2 = the transit time of an electron from mesh to anode.
T = the transit time of an electron from cathode to anode.

electron to pass down the length of the tube be given by[5]

$$T = \frac{2\pi n m}{e H_z} \tag{10}$$

i.e., T must be equal to an integral number of cyclotron periods of the electron.

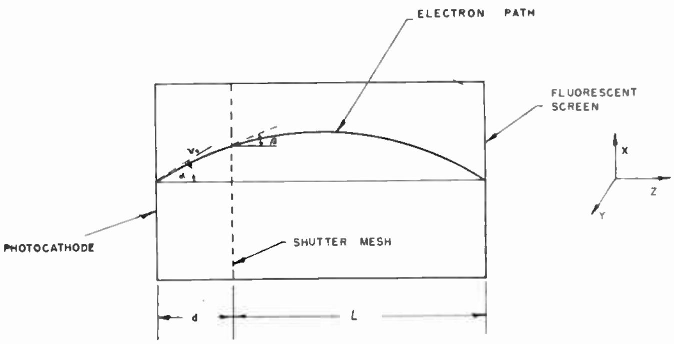

Fig. 14—The basic focusing system for the magnetically focused image converter.

Here e is the charge on an electron, m the mass of the electron, H_z the axial component of the magnetic field, and n is an integer. MKS units are used.

The time T_1 from cathode to mesh is

$$T_1 = d\left(\frac{2m}{eV_m}\right)^{1/2}. \tag{11}$$

Here we have neglected the initial velocity since this is only 2 or 3 volts compared to about 40 volts on the mesh.

The time, T_2, to travel between $z=d$ and $z=L+d$ may be shown to be

$$T_2 = \left(\frac{2m}{e}\right)^{1/2} \frac{L\left\{\left(1+\frac{V_m}{V_s}\right)^{1/2} - \left(\frac{V_m}{V_s}\right)^{1/2}\right\}}{V_s^{1/2}}. \tag{12}$$

[5] *Ibid.*, p. 79.

It has been assumed here that $\cos\beta \simeq 1$ (see Fig. 14).

Then from (30) and (31), we have

$$T = T_1 + T_2 = \left(\frac{2m}{e}\right)^{1/2} \cdot \left[\frac{d}{V_m^{1/2}} + \frac{L\left\{\left(\frac{V_m}{V_s}+1\right)^{1/2} - \left(\frac{V_m}{V_s}\right)^{1/2}\right\}}{V_s^{1/2}}\right]. \tag{13}$$

Substituting (32) into (29), we have

$$H_z = \frac{2\pi n}{\left(\frac{2e}{m}\right)^{1/2}\left[\frac{d}{V_m^{1/2}} + \frac{L\{(1+V_m/V_s)^{1/2} - (V_m/V_s)^{1/2}\}}{V_s^{1/2}}\right]}. \tag{14}$$

Using $e/m = 1.77 \times 10^{11}$ coulombs/kg and multiplying by 10^4 to change webers/(meter)^2 into Gauss, we obtain

$$H_z \text{ (Gauss)} = \frac{0.106n}{\left[\frac{d}{V_m^{1/2}} + \frac{L}{V_s^{1/2}}\left\{\left(1+\frac{V_m}{V_s}\right)^{1/2} - \left(\frac{V_m}{V_s}\right)^{1/2}\right\}\right]} \tag{15}$$

where d and L are in meters, while V_m and V_s are in volts.

C. Experimental Verification of the Theory

Experimental data on the static characteristics of the tube are presented in Figs. 15 through 17.

Theoretically, the lines in Fig. 15 should pass through the origin according to (15). The fact that they do not pass through the origin may be attributed to inaccuracies in the experimental data associated with the difficulties in judging best focus. This difficulty was increased by the presence of the mesh. What was actually chosen as best focus was really a compromise between actual image focus and mesh interference.

Fig. 16 shows the variation of the magnetic field strength required for best focus as a function of the applied anode voltage. The dotted line for the sixth order focus is the predicted variation according to (15).

Fig. 17 shows the variation of the magnetic field strength required for best focus as a function of the applied mesh voltage. The circles represent the experimental data while the solid line is plotted from (15). The agreement between the simple theory and the experimental data is relatively good.

It will be noted that no data for the first order focus is presented. Actually, the first order focus was not discernible due to the interaction between magnetic focusing and the electrostatic focusing by the cylinder supporting the shutter mesh. As a matter of fact, with the magnetic field at zero, it is possible to obtain an electrostatically focused image which is inverted. As the magnetic field is increased, the image is then rotated to

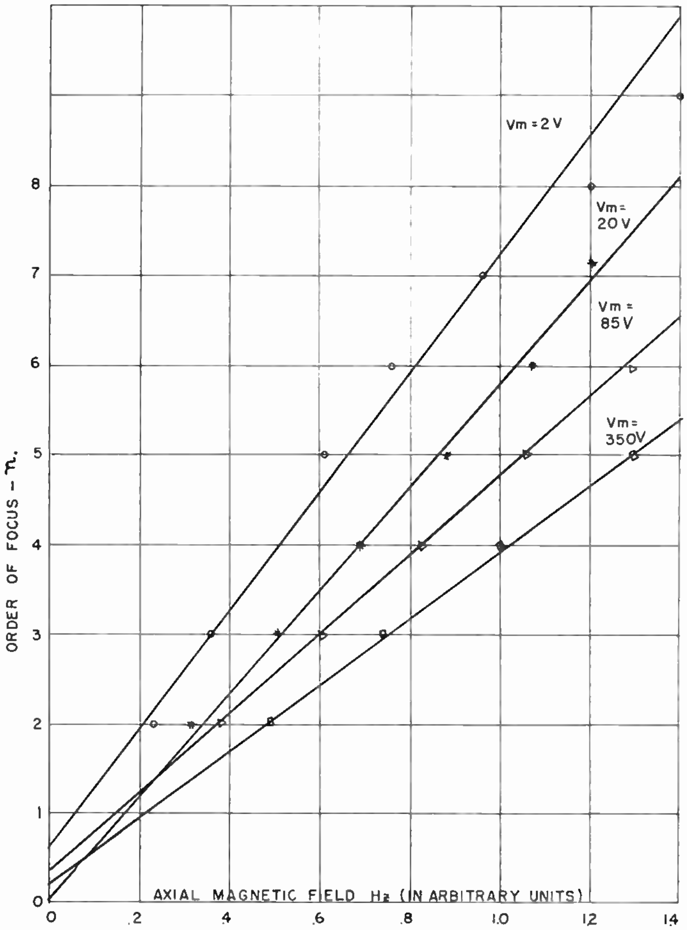

Fig. 15—Variation of order of focus, n, with the **axial** magnetic field (experimental).

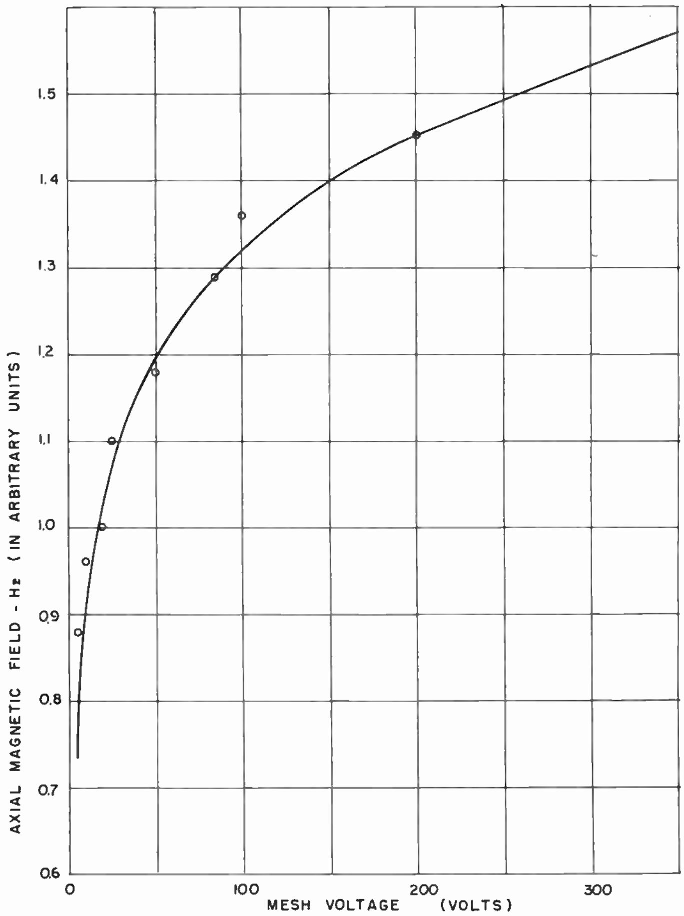

Fig. 17—Variation of the axial magnetic field, H_z, with mesh voltage, V_m, for the sixth order focus. The circles are experimental points. The solid line is the theoretical curve.

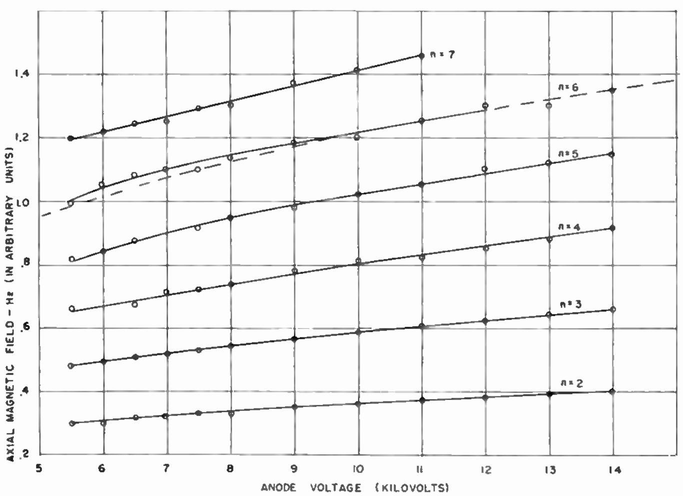

Fig 16—Variation of the axial magnetic field, H_z, with applied anode potential for various orders of focus. The circles are experimental points and the solid lines are curves drawn to fit the experimental data. The dashed line is the theoretical curve for $n=6$.

an upright position. Somewhere during the rotation of the image the magnetic field reaches a high enough value to give the first order focus. However, the electrostatic focusing interaction is still strong enough to prevent the formation of a focused image.

Fig. 17 illustrates an important advantage of the magnetic focus shutter tubes over the electrostatic focus tubes discussed in part I; namely, that for a given over-all voltage one has a wide choice of mesh voltages for which a focus is still attainable by varying the magnetic field. This means that it is possible to use a low mesh voltage for ease in pulsing, but still keep a high anode voltage for brightness. With the electrostatic tube the ratio of mesh to anode voltage was a function of the geometry of the tube only and therefore could not be varied once the tube had been constructed.

Unfortunately, the magnetic focus image converter has other difficulties. The solenoid current and applied voltages must be carefully filtered and regulated, the components are bulky and heavy and application of deflection fields is very difficult if high speed is required.

Fig. 18 is a photograph of one of the magnetic focus shutter image converter tubes.

Appendix I

Determination of the Virtual Image Position

The notation to be used in this appendix is explained in Fig. 1 and Table I.

Consider first the cathode-screen system. The electron is moving in a field whose intensity varies according to

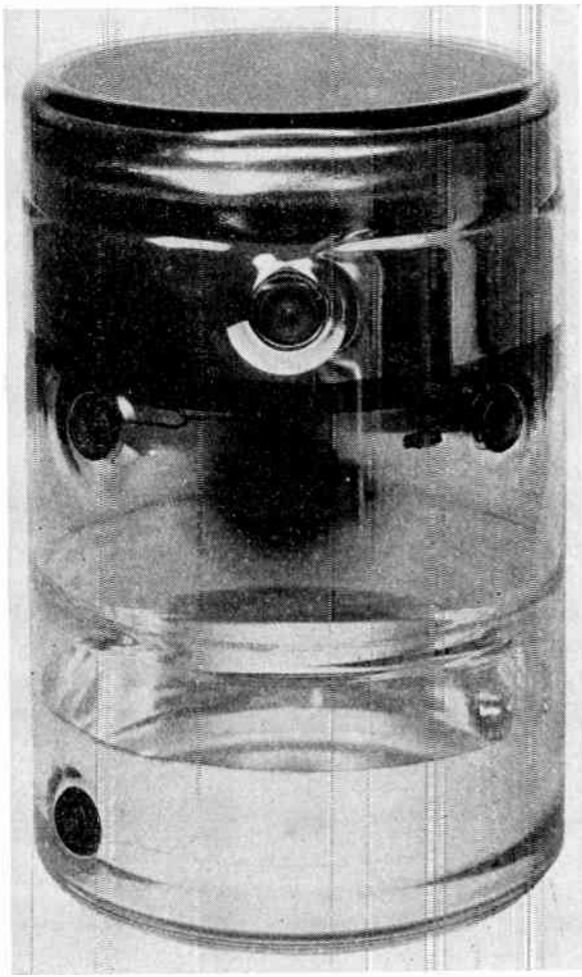

Fig. 18—Photograph of a magnetically focused image converter.

the inverse square law. Thus the well known theory of the motion of a particle in a central field of force may be applied.[6] Accordingly, the trajectory of the electron may be described by two coordinates, the radial distance ρ, and the polar angle ϕ. The two basic equations used express the conservation of angular momentum and energy, *i.e.*,

$$\rho^2\dot{\phi} = \kappa = \text{constant} \tag{16}$$

$$\tfrac{1}{2}mv^2 = \tfrac{1}{2}mv_0^2 + e\Phi_\rho. \tag{17}$$

Since $R_c\dot{\phi}_0 = v_0 \sin\alpha$, we have that $K = R_c v_0 \sin\alpha$. The velocity v may be written

$$v^2 = (\dot{\rho})^2 + (\rho\dot{\phi})^2 \tag{18}$$

Eliminating parameter time from (16) and (17),

$$d\phi = \frac{R_c \sin\alpha \dfrac{d\rho}{\rho^2}}{\left\{\left(1 - \dfrac{e\Phi_0}{\frac{1}{2}mv_0^2}\right) + \dfrac{e\Phi_0}{\frac{1}{2}mv_0^2}\dfrac{R_c}{\rho} - \left(\dfrac{R_c}{\rho}\sin\alpha\right)^2\right\}^{1/2}}$$

[6] L. Goldstein, "Classical Mechanics," Addison-Wesley Press, Inc., Cambridge, Mass., ch. 3; 1950.

$$= \frac{(-\sin\alpha)d\mu}{\{(1-q) + q\mu - \mu^2\sin^2\alpha\}^{1/2}}. \tag{19}$$

Integrating (19) we have

$$\tan\frac{\phi}{2} = \frac{-\cos\alpha + \{(1-q) + q\mu - \mu^2\sin^2\alpha\}^{1/2}}{\left(\dfrac{q}{\sin\alpha}\right) - (\mu+1)\sin\alpha}. \tag{20}$$

In (20) the value of ϕ at $\rho = R_c$ has been taken as zero. This will not affect the generality of the theory. At $\rho = R_s$, one has $\phi = f$ so that

$$\tan\frac{f}{2} = \frac{-\cos\alpha + \{(1-q) + qy - y^2\sin^2\alpha\}^{1/2}}{\left(\dfrac{q}{\sin\alpha}\right) - (y+1)\sin\alpha}. \tag{21}$$

Since $\tan\delta$ is equal to the ratio of polar to radial velocity at $\rho = R_c$ we may write

$$\tan\delta = \left(\frac{\rho\dot{\phi}}{\dot{\rho}}\right)_{\rho=R_c} = -\left(\mu\frac{d\phi}{d\mu}\right)_{\mu=1}. \tag{22}$$

Substituting (19) into (22) we get

$$\tan\delta = \frac{y\sin\alpha}{\{(1-q) + qy - y^2\sin^2\alpha\}^{1/2}}. \tag{23}$$

A similar calculation shows that in the region between the mesh and the anode

$$\tan\frac{g}{2} = \frac{-\cos\delta + \{(1-t) + tx - x^2\sin^2\delta\}^{1/2}}{\left(\dfrac{t}{\sin\delta}\right) - (x+1)\sin\delta} \tag{24}$$

$$\tan d = \frac{x\sin\delta}{\{(1-t) + tx - x^2\sin^2\delta\}^{1/2}}. \tag{25}$$

It will be noted that (24) and (25) are obtained immediately from (21) and (23) by the following change of variables:

$$\alpha \to \delta,\quad q \to t,\quad y \to x,\quad f \to g,\quad \delta \to d.$$

To estimate the value of q we take $e\Phi_s = 100$ electron-volts, $E_0 = 3$ electron-volts and $y - 1 = 0.1$. Then q is approximately 300. Thus with respect to q we may neglect terms of the order of magnitude of 1 in (21) and (23). This gives

$$f = 2\left(\frac{y-1}{q}\right)^{1/2}\sin\alpha \tag{26}$$

$$\delta = \frac{y\sin\alpha}{\{q(y-1)\}^{1/2}} \tag{27}$$

where in (26) and (27) we have set $\tan f/2 = f/2$ and $\tan\delta = \delta$. To estimate the value of t take $e\Phi_a = 10^4$ electron-volts and $x - 1 = 10$. This gives $t \simeq 3$. Here we cannot neglect terms of the order of magnitude of 1 with respect to t so that we must look for a different approximation in simplifying (24) and (25). We have

$$t = \frac{e\Phi_1}{E_1} = \frac{e(\Phi_a' - \Phi_s)}{(x-1)(E_0 + e\Phi_a)}. \tag{28}$$

But from Table I

$$\frac{e\Phi_a'}{E_0 + e\Phi_s} = \frac{p(n-1)}{1 + q(y-1)} \tag{29}$$

and

$$\frac{e\Phi_s}{E_0 + e\Phi_s} = \frac{q(y-1)}{1 + q(y-1)}. \tag{30}$$

Taking $1+q(y-1) \simeq q(y-1)$ and substituting (29) and (30) into (28) we have

$$t \simeq \frac{1}{(x-1)}\left\{\frac{p(n-1)}{q(y-1)} - 1\right\} = \frac{r^2 - 1}{x - 1}. \tag{31}$$

Since sin δ is small compared to 1 we have from (31)

$$1 - t + tx - x^2 \sin^2 \delta \simeq r^2 \tag{32}$$

$$\frac{t}{\sin \delta} - (x+1)\sin\delta = \frac{r^2 - 1}{(x-1)\sin\delta}. \tag{33}$$

Substituting (27), and (31)–(33) into (24) and (25) gives

$$g = \frac{2y(x-1)\sin\alpha}{(r+1)\{q(y-1)\}^{1/2}} \tag{34}$$

$$d = \frac{n \sin \alpha}{r\{q(y-1)\}^{1/2}}. \tag{35}$$

From Fig. 1 it can be seen that

$$R_i = R_a \frac{d}{g + f - d}. \tag{36}$$

Substituting (26), (34), and (35) into (36) we obtain

$$R_i = R_a \frac{n(r+1)}{r(n-2) + 2(y-1)r^2 - n}. \tag{37}$$

This is the expression for the radius of the virtual image sphere given in terms of the geometrical and electrical parameters of the tube. For the case where the mesh potential is such that it does not disturb the cathode-anode potential distribution (*i.e.*, a degeneration from the three-electrode to the two-electrode case) it can be shown that

$$r^2 = \frac{n-1}{y-1} \quad \text{or} \quad p = q$$

and (37) reduces to

$$R_i = R_a \frac{n}{n-2}. \tag{38}$$

Appendix II

Derivation of the Focal Length Formula for an Aperture in the Inner Sphere of a System of Two Concentric Spheres

In deriving the formula for the focal length of the aperture the field inside the inner sphere will also be taken as a radial field as though there were a point charge at the center of the system. For the particular case in hand we will want to set this internal field equal to zero later because the image converters will have a field free region inside the inner sphere. Performing the calculation with a radial field inside the inner sphere adds a little more generality to the results. It will be assumed that the aperture in the inner sphere is small compared to the radius of the inner sphere. This allows us to use the thin lens formula for the determination of the focal length expression. A second assumption is that all angles are small enough so that their sines and tangents may be replaced by the angles themselves.

Consider Fig. 19. $AB(=z)$ is the virtual image distance for no focusing by the aperture. $AD(=s)$ is the real image distance determined by the lens action of the aperture.

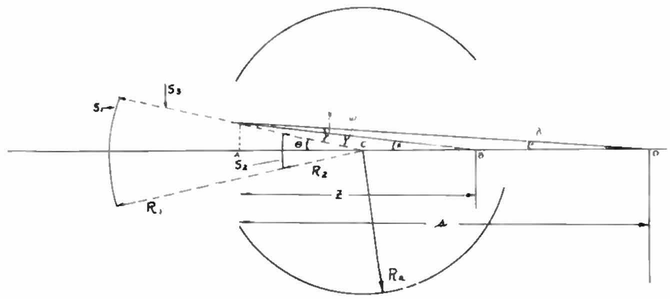

Fig. 19—Sketch of anode aperture system in the electrostatic image converter.

The polar angular momentum, P_ϕ, of the electron whose radius vector has length ρ is given by

$$P_\phi = m\rho^2\dot{\phi}. \tag{39}$$

The change in P_ϕ when the electron passes through the aperture is

$$\Delta P_\phi = e\int_{t_1}^{t_2} [\vec{\rho} \times \vec{\mathcal{E}}]_\phi dt = e\int_{t_1}^{t_2} \rho \mathcal{E}_\phi dt. \tag{40}$$

Here $\mathcal{E}_\phi$ is the polar component of $\vec{\mathcal{E}}$. t_1 and t_2 are times corresponding to two surfaces of radii R_1 and R_2. R_1 and R_2 are so chosen that the field may be considered to be a central field at those two surfaces. From Gauss' theorem we have

$$\int_S \vec{\mathcal{E}} \cdot \vec{dS} = 0 \tag{41}$$

where the integral is taken over a closed surface, S. The surface we will take is that of the frustrum of a cone with half angle θ bounded by the spherical surfaces S_1 and S_2 and the surface S_3. The integrals over S_1, S_2, and S_3 then give

$$\int_{S_1} \vec{\mathcal{E}} \cdot \vec{dS} = 2\pi R_1^2 \mathcal{E}_1 (1 - \cos\theta) \tag{42}$$

$$\int_{S_2} \vec{\mathcal{E}} \cdot \vec{dS} = 2\pi R_2^2 \mathcal{E}_2 (1 - \cos\theta) \tag{43}$$

$$\int_{S_3} \vec{\mathcal{E}} \cdot \vec{dS} = 2\pi \sin\theta \int_{R_1}^{R_2} \mathcal{E}_\phi \rho d\rho. \tag{44}$$

Here $\mathcal{E}_1$ and $\mathcal{E}_2$ are the magnitudes of $\vec{\mathcal{E}}$ at R_1 and R_2, respectively, and $\mathcal{E}_\phi$ is the polar component of $\vec{\mathcal{E}}$. Hence from (41)–(44) we have

$$\int_{R_1}^{R_2} \mathcal{E}_\phi \rho d\rho = \frac{(1 - \cos\theta)}{\sin\theta} (R_2^2 \mathcal{E}_2 - R_1^2 \mathcal{E}_1) \tag{45}$$

but

$$\Delta P_\phi = e \int_{t_1}^{t_2} \mathcal{E}_\phi \rho dt = e \int_{R_1}^{R_2} \frac{\rho \mathcal{E}_\phi}{v} d\rho \simeq \frac{e}{v_c} \int_{R_1}^{R_2} \mathcal{E}_\phi \rho d\rho \tag{46}$$

where v_c is the velocity of the electron on passing through the center of the aperture. It is here assumed that the lens is so thin that the velocity is almost constant over that part of the trajectory where $\mathcal{E}_\phi$ is different from zero.

Combining (45) and (46)

$$\Delta P_\phi = \frac{e(1 - \cos\theta)}{v_c \sin\theta} (R_2^2 \mathcal{E}_2 - R_1^2 \mathcal{E}_1). \tag{47}$$

If it is assumed that the change in angular momentum takes place abruptly on passing through the aperture then the angular momentum before entering the aperture $= m v_c R_a \sin w$; and the angular momentum after passing through the aperture $= m v_c R_a \sin \psi$.

Thus

$$\Delta P_\phi = m v_c R_a (\sin\psi - \sin\omega) \simeq m v_c R_a (\psi - \omega). \tag{48}$$

From Fig. 19 we have

$$-\frac{R_a \sin\phi}{z} = \tan\kappa \simeq \kappa = \phi - \omega \tag{49a}$$

$$-\frac{R_a \sin\phi}{s} = \tan\lambda \simeq \lambda = \phi - \psi. \tag{49b}$$

Using the thin lens formula we have

$$\frac{1}{F} = \frac{1}{z} + \frac{1}{s} = \frac{\omega - \psi}{R_a \sin\phi} = -\frac{\Delta P_\phi}{m v_c R_a^2 \sin\phi}. \tag{50}$$

Substituting (47) into (50)

$$\begin{aligned} \frac{1}{F} &= -\frac{e(1 - \cos\theta)}{m v_c^2 R_a^2 \sin^2\phi} (R_2^2 \mathcal{E}_2 - R_1^2 \mathcal{E}_1) \\ &\cong \frac{-e(R_2^2 \mathcal{E}_2 - R_1^2 \mathcal{E}_1)}{2 m v_c^2 R_a^2}. \end{aligned} \tag{51}$$

For the case under consideration

$$\mathcal{E}_2 = 0, \qquad \mathcal{E}_1 = -\frac{\Phi_a R_s}{R_1^2 (x - 1)}.$$

Hence

$$\frac{1}{F} = -\frac{x}{4R_a(x - 1)} \frac{\Phi_a}{\Phi_c}. \tag{52}$$

It is interesting to note that when R_1, R_2, and R_a approach infinity, (51) reduces to the Davisson-Calbick formula for the focal length of an aperture in an infinite plane.

Acknowledgment

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Minimizing Incidental Frequency Modulation in Amplitude-Modulated UHF Oscillators*

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Summary—It is possible to minimize incidental fm in amplitude-modulated uhf oscillators. The frequency changes at uhf that occur over an amplitude-modulated cycle stem, primarily, from the variation of the transit time and space charge within the oscillator tube. By appropriate selection of parameters of the feedback and cathode lines a compensation is obtained for changes in cathode-to-grid transit time. The exact parameters can be approximately calculated from an analysis of the oscillator circuit. However, a final experimental adjustment is usually necessary.

Cathode or grid modulation is better than plate modulation for low incidental fm. The frequency changes caused by variations of grid-to-plate transit time and space charge tend to cancel for cathode of grid modulation; whereas for plate modulation they add. By use of cathode modulation and the previously mentioned compensation, the frequency changes in a 400-mc oscillator were less than 5 kc for a cathode-current variation of 9 to 13 ma.

Introduction

In this paper principles are given for minimizing incidental fm in amplitude-modulated uhf oscillators. The term "incidental fm" refers to the frequency changes which occur when the electrode potentials of an oscillator are varied to produce an amplitude-modulated wave. From the analytical and experimental work which has been done, two important conclusions can be drawn. These are as follows:

1) Incidental fm can be minimized by proper selection of parameters of the feedback and cathode circuits.
2) Cathode or grid modulation is better than plate modulation for low incidental fm.

There are many instances where incidental fm is a serious problem; this is particularly true in low-power uhf transmitters where economy is an important factor. Very often in order to meet FCC specifications on incidental fm, the use of a very low per cent modulation is necessary; sometimes an expensive master-oscillator-power amplifier combination is required. It is expected that the method of reducing incidental fm described in this paper can help eliminate some of these uneconomical practices.

Researchers have met with considerable success in reducing the incidental frequency change at low frequencies;[1,2] whereas, at uhf this reduction has not as yet been satisfactorily accomplished. The reason for the lack of success at the higher frequencies is that the effect of variation of transit time is added to and overshadows the effect of variation of resistive parameters of the oscillator tube. Variations of transit time at uhf are enough to change oscillation frequencies from $\frac{1}{4}$ to 1 per cent for amplitude modulations of 80 to 90 per cent.[3] Compensation for the transit time variations is the substance of the method of reducing incidental fm which is described in this paper.

The Oscillator to be Compensated

Compensation will be developed for a triode-transmission-line oscillator which has external feedback. Fig. 1 shows a schematic diagram of this type of oscillator. Admittances Y_L and Y_k represent tuned-transmission-line plate and cathode circuits, respectively. The M denotes the mutual inductance between the plate circuit and the feedback-line-coupling loop, which has a self-reactance of $+X_f$. Other elements shown in Fig. 1

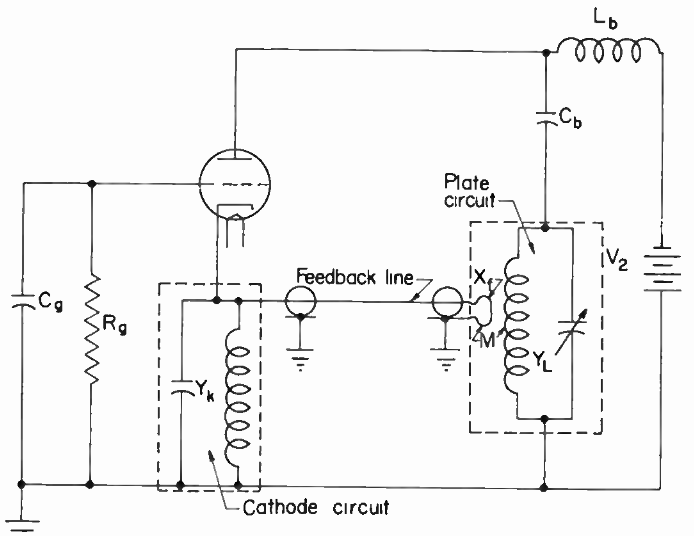

Fig. 1—Schematic diagram of a transmission-line oscillator with external feedback.

are the $R_g - C_g$ grid-leak-bias network, plate blocking capacitor C_b, and rf choke, L_b. It will be assumed that these latter elements provide either perfect short or open circuits to the rf energy.

Although it is possible to derive the compensation for the general circuit represented by Fig. 1, it has been found convenient to specify the oscillator configuration more exactly. Both plate and cathode circuits are as-

* Original manuscript received by the IRE, October 10, 1956; revised manuscript received, November 15, 1956.

† Stewart-Warner Electronics, Chicago 51, Ill.

[1] F. B. Llewellyn, "Constant frequency oscillators," *Bell Sys. Tech. J.*, vol. 2, pp. 67–100; January, 1932.

[2] J. K. Clapp, "Frequency stable LC oscillators," Proc. IRE, vol. 42, pp. 1295–1300; August, 1954.

[3] W. G. Dow, "Transit-time effects in ultra-high frequency class C operation," Proc. IRE, vol. 35, pp. 35–42; January, 1947.

sumed to be shorted-transmission lines with Y_L and Y_k the equivalent admittances at the open end of their respective lines. In addition the feedback-line-coupling loop is assumed to be at the shorted end of the plate-circuit-transmission line, although this is a matter of convenience and not necessity.

The Oscillator-Tube-Equivalent Circuit

The analysis of the oscillator shown in Fig. 1 is possible because of the development of an equivalent circuit for the triode accurate at uhf.[4] Such an equivalent circuit, derived by assuming parallel-plane electrodes, is shown in Fig. 2. The effect of transit-time on the parameters of Fig. 2 can be calculated because the development of that circuit takes into consideration the electronic motion in the tube. Cathode, grid, and plate terminals have voltages denoted by V_k, V_{gt}, and V_p, respectively. The internal junction of the three branches of the circuit is called the equivalent grid. In reality, this equivalent grid represents a region bounded by equipotential planes which are at a potential V_g. The action of this region on distant electrons is the same as that of the grid wires. The y_g is the admittance or capacitance,[4] when grid current is zero, between the equivalent grid and grid wires. Other elements are cathode to equivalent-grid admittance y_{11} (consisting of conductance g_{11} in parallel with susceptance $+b_{11}$), transadmittance y_{12}, and equivalent grid-to-plate admittance y_{22} (a pure capacitance). Normally, y_g is small enough to be considered a short-circuit at the operating frequency; in such a case Fig. 2 can be redrawn as Fig. 3.

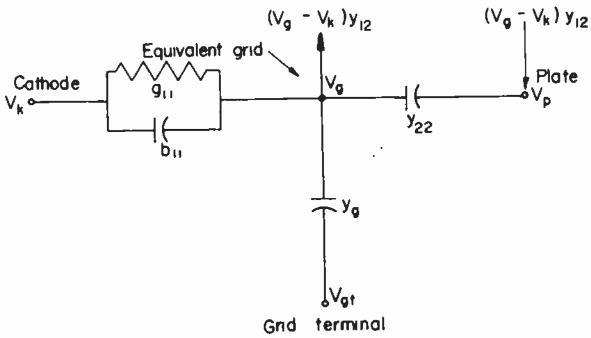

Fig. 2—Complete equivalent circuit of a triode.

For relatively low values to transit time y_{11} can be approximately given by (1) in terms of dc cathode current I_0, dc equivalent-grid voltage V_1, and average or dc cathode-to-grid transit angle θ_1.

$$y_{11} \doteq g_0 + j0.3g_0\theta_1 \tag{1}$$

where

$$g_0 \equiv \frac{3I_0}{2V_1} \tag{2}$$

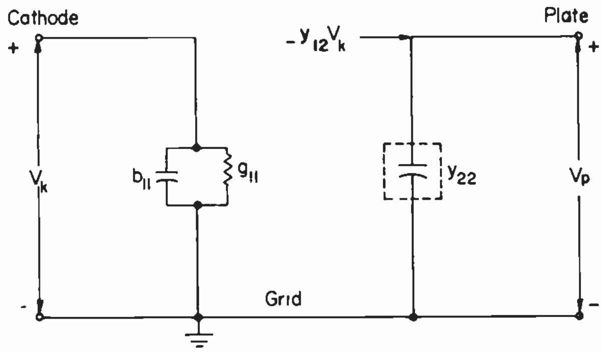

Fig. 3—Equivalent circuit of a triode with negligible y_g for grounded-grid operation.

and

$$\theta_1 \equiv 2\pi fT_1 \tag{3}$$

with f the operating frequency and T_1 the dc cathode-to-grid transit time. Eq. (1) is valid under the following conditions:

1) The rf fluctuations in transit time are small compared to the dc transit time.
2) The transit angle θ_1 is less than 2 radians.
3) The cathode-to-grid region is space-charge limited.
4) There exists one and only one potential minimum plane in the cathode-to-grid region, and that plane is near the cathode.

The assumption of these conditions does not usually produce serious errors for frequencies below 1000 megacycles.

By assuming that both grid-to-plate transit angle θ_2 and grid-to-plate region space charge are zero, y_{12} can also be given in terms of g_0 and θ_1.

$$y_{12} \doteq -g_0e^{-j0.366\theta_1}. \tag{4}$$

The effect of a nonzero θ_2 on y_{12} will be discussed later.

Under conditions of zero space charge in the grid-to-plate region y_{22} is equal to the admittance of the measured grid-to-plate capacitance c_{22}.

$$y_{22} = jwc_{22}. \tag{5}$$

The effect of a nonzero space charge will also be discussed later.

The Oscillator Equivalent Circuit

An equivalent circuit of the oscillator whose schematic is shown in Fig. 1 can now be drawn; it is shown in Fig. 4. Cathode-line admittance is assumed to be a susceptance B_k. Plate-line admittance is combined with y_{22} and is shown as Y_L, consisting of conductance G_L in parallel with susceptance B_L. Conductance G_L accounts for the plate-transmission-line loading and losses. Susceptance B_L consists of the grid-to-plate capacitance in parallel with the equivalent capacitance and inductance of the plate-transmission line at the operating frequency.

[4] F. B. Llewellyn and L. C. Peterson, "Vacuum-tube networks," Proc. IRE, vol. 32, pp. 144–166; March, 1944.

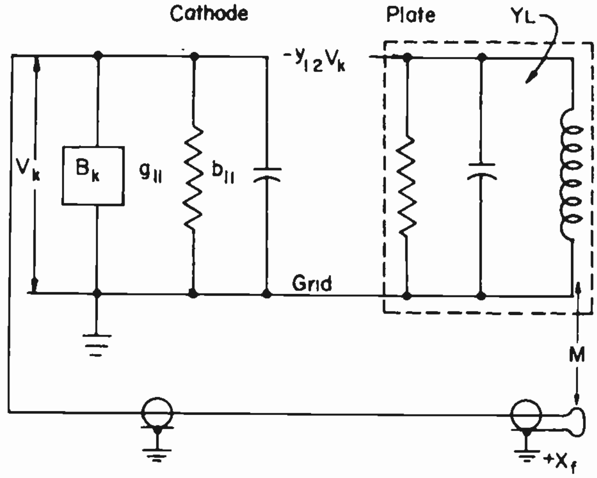

Fig. 4—New equivalent circuit of the oscillator.

Analysis of the Oscillator

The method of reducing incidental fm is developed from an analysis of the circuit in Fig. 4. Basically, if a voltage V_k is impressed from cathode to ground, conditions of oscillation must be such that the same V_k is fed back to the cathode. An identity between circuit parameters, if assumed linear, can be written, from which two equations, one for magnitude and one for phase, are obtained. Generally, the phase equation controls oscillation frequency.

As a first step the effect of impressed voltage V_k is traced through the tube and plate circuit up to the feedback-line-coupling loop. The tube's output voltage V_p, which is across the open end of the plate-circuit-transmission line, is given by

$$V_p = -\frac{y_{12}V_k}{Y_L}. \tag{6}$$

The current I_s at the shorted end of the plate line is obtained in terms of V_p by use of the transmission-line equations.

$$I_s = (-jG_0 \csc\theta)V_p = \frac{(jG_0y_{12}\csc\theta)V_k}{Y_L} \tag{7}$$

where G_0 and θ are the plate-line-characteristic conductance and electrical length, respectively. It is assumed that the feedback-line-coupling loop is at the shorted end of the plate line. The induced voltage V_i on the coupling loop, therefore, can be given by $jwMI_s$, which leads to

$$V_i = -\frac{(wMG_0y_{12}\csc\theta)V_k}{Y_L}. \tag{8}$$

Voltage V_i, modified by the self-reactance $+X_f$ of the feedback-line-coupling loop, is applied to the input end of the transmission line which feeds energy back to the cathode to sustain oscillations. In order to obtain the cathode voltage V_k at the output end of the feedback line, further recourse to transmission-line theory has to be made. Assuming a characteristic conductance G_{0f} and electrical length θ_f for the feedback line, (9) gives the cathode voltage out of the feedback line in terms of V_i.

$$V_k = \frac{V_i}{G_{0f}\cos\theta_f + jY_{in}\sin\theta_f + jX_fG_{0f}(Y_{in}\cos\theta + jG_{0f}\sin\theta_f,} \tag{9}$$

where Y_{in} is the admittance looking into the cathode circuit from the cathode end of the feedback line. It is evident from Fig. 4 that Y_{in} can be given by

$$Y_{in} = g_{11} + j(b_{11} + B_k) = g_0 + j(0.3g_0\theta_1 + B_k). \tag{10}$$

An important identity results from the substitution of (8) and (10) into (9) and the cancellation of the common V_k's. In writing this identity it is convenient to use the following abbreviated notation:

$$R \equiv G_{0f}\cos\theta_f - X_fG_{0f}^2\sin\theta_f \tag{11}$$

$$S \equiv \sin\theta_f + X_fG_{0f}\cos\theta_f. \tag{12}$$

It is also useful to express Y_L and y_{12} in polar form. The identity can, therefore, be written as follows:

$$1 = \frac{(wMG_0G_{0f}\csc\theta)g_0e^{-j0.366\theta_1}}{|Y_L|e^{j\theta_L}(R - B_kS - 0.3\theta_1g_0S + jg_0S)} \tag{13}$$

where θ_L and $|Y_L|$ are the phase angle and magnitude, respectively, of the plate-circuit admittance.

The phase angle θ_L of the plate-circuit admittance is obtained by equating the total phase of the right-hand expression of (13) to 2π radians.

$$\theta_L = 2\pi - 0.366\theta_1 - \text{ctn}^{-1}\left(\frac{R}{g_0S} - \frac{B_k}{g_0} - 0.3\theta_1\right). \tag{14}$$

Eq. (14) is the final necessary step in the analysis of the oscillator circuit; the compensation can now be developed.

Reduction of the Effect of θ_1 Changes on θ_L

Since oscillation frequency is a function of θ_L, the reduction of variations of θ_L as θ_1 changes would be important in reducing incidental fm. Conditions for this reduction of θ_L changes are determined by setting the derivative, with respect to θ_1, of θ_L in (14) equal to zero.

To find the derivative of (14), g_0 must be expressed in terms of θ_1. The exact relation between g_0 and θ_1 is obtained by multiplying together expressions found in the literature for g_0 per unit area and transit time T_1.[5] The product of g_0 per unit area and T_1 is then multiplied by the area of the cathode a_1 and $2\pi f$. Therefore,

$$\frac{2\pi fT_1a_1g_0}{a_1} = \theta_1g_0 = \frac{111 \times 10^{-14}fa_1}{d_1} \tag{15}$$

with a_1 in cm² and cathode-to-grid spacing d_1 in cm.

Eliminating g_0 in (14) by use of (15)

$$\theta_L = 2\pi - 0.366\theta_1 - \text{ctn}^{-1}(A\theta_1) \tag{16}$$

[5] A. H. W. Beck, "Thermionic Valves: Their Theory and Design," Cambridge University Press, Cambridge, Eng., pp. 403, 407; 1953.

where

$$A \equiv \frac{Rd_1}{111 \times 10^{-14} f a_1 S} - \frac{B_k d_1}{111 \times 10^{-14} f a_1} - 0.3. \quad (17)$$

A quadratic equation in A, with functions of θ_1 as coefficients, results from the differentiation of (16) with respect to θ_1 (A considered constant) and subsequent setting of $d\theta_L/d\theta_1$ equal to zero. Of the two solutions of A in terms of θ_1 the one using the minus term in the quadratic formula is the more useful. This particular A, denoted by A_d, is given by (18) and is plotted vs θ_1 in Fig. 5.

$$A_d = \frac{1.36 - (1.85 - \theta_1^2)^{1/2}}{\theta_1^2}. \quad (18)$$

The use of such an A_d means that changes in the corresponding initial θ_1 have a minimum effect on θ_L and, therefore, on frequency. In other words proper selection of feedback-line parameters and cathode-line susceptance, which are determined by (17) and (18), results in reduced incidental fm.

A study of the second derivative $d^2\theta_L/d\theta_1^2$ yielded no further results other than the conclusion that low values of θ_1 helps keep incidental fm low.

Effect of the Grid-to-Plate Region

The previous analysis neglected the grid-to-plate region space charge and transit angle, the effects of which are discussed in this section.

By assuming a small grid-to-plate transit angle θ_2, a modified expression for y_{12} can be obtained from the general y_{12} as given by Llewellyn and Peterson.[4]

$$y_{12} = -g_0 e^{-j0.366\theta_1}\left[1 - \frac{j\theta_2(V_1^{1/2} + 2V_2^{1/2})}{3(V_1^{1/2} + V_2^{1/2})}\right] \quad (19)$$

where V_2 is the plate voltage. Considering $V_2 \gg V_1$, y_{12} can be approximately given by

$$y_{12} \doteq -g_0 e^{-j0.366\theta_1}(1 - j0.667\theta_2). \quad (20)$$

The $0.667\theta_2$ term is usually much smaller than 1 so that further simplification of y_{12} is possible.

$$y_{12} \doteq -g_0 e^{-j(0.366\theta_1 - 0.667\theta_2)}. \quad (21)$$

Eq. (16) for θ_L can now be modified as follows to include the effect of a small but not negligible θ_2:

$$\theta_L = 2\pi - 0.336\theta_1 - 0.667\theta_2 - \mathrm{ctn}^{-1}(A\theta_1). \quad (22)$$

The formula used to calculate θ_2 is given by (23), which is the expression for transit angle in a zero-space-charge region.

$$\theta_2 = \frac{4\pi f d_2}{5.95 \times 10^7 (V_1^{1/2} + V_2^{1/2})} \text{ radians} \quad (23)$$

where d_2 (the grid-to-plate spacing) is in cm, f is in cycles per second, and V_1 and V_2 are in volts.

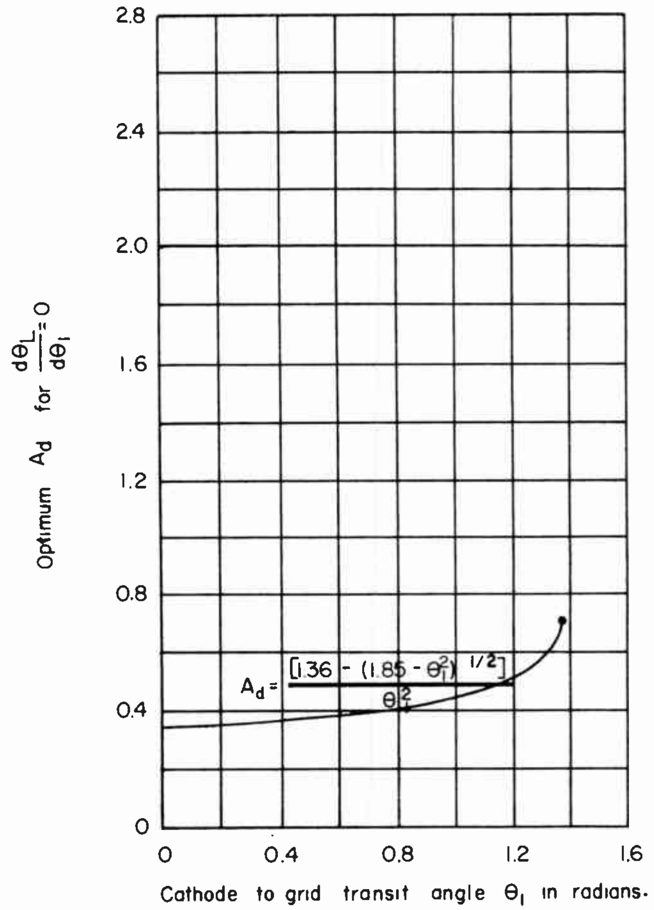

Fig. 5—Optimum A_d for $d\theta_L/d\theta_1 = 0$ vs cathode-to-grid transit angle θ_1.

The presence of a small space charge in the grid-to-plate region alters the value of the grid-to-plate admittance y_{22}. From a general expression of y_{22} given in the literature and assuming small θ_2 and grid-to-plate region space charge, the following can be written:

$$y_{22} = jwc_{22}\left(1 + \frac{2\psi_2}{3}\right) \quad (24)$$

where ψ_2 is a space-charge factor defined from the ratio of grid-to-plate transit time T_{02} if space charge were zero to the actual grid-to-plate transit time T_2.

$$\psi_2 \equiv 3\left(1 - \frac{T_{02}}{T_2}\right) \quad (25)$$

In the grid-to-plate region ψ_2 is small enough to be calculated by the following approximation:

$$\frac{9\psi_2}{4} = \frac{I_2 d_2^2}{2.33 \times 10^{-6} a_2 (V_1^{1/2} + V_2^{1/2})^3} \quad (26)$$

where the dc plate current I_2 is in amps, d_2 is in cm, the grid-to-plate region area a_2 perpendicular to electron motion is in cm², and V_1 and V_2 are in volts.

The effects of changes in the grid-to-plate region space charge and transit angle are discussed in the next section.

Relation Determining Frequency Change

Frequency change can be given in terms of changes of the three parameters of electronic motion previously discussed, θ_1, θ_2, and ψ_2. The conclusions given at the beginning of this paper are based on that relationship.

The procedure is to write an expression for the plate-circuit susceptance and to add increments to those terms dependent on θ_1, θ_2, and ψ_2. The plate-circuit susceptance B_L can be represented by an equivalent capacitance C_{eq} in parallel with an equivalent inductance L_{eq}. Therefore,

$$B_L = 2\pi f C_{eq} - \frac{1}{2\pi f L_{eq}} = G_L \tan\theta_L = \frac{G_L \sin\theta_L}{\cos\theta_L}. \quad (27)$$

If C_{eq} is increased by dC_{eq} and θ_L is increased by $d\theta_L$, then f is increased by df (the df can actually have a negative sign).

$$2\pi(f+df)(C_{eq}+dC_{eq}) - \frac{1}{2\pi(f+df)L_{eq}} = \frac{G_L(\sin\theta_L\cos d\theta_L + \cos\theta_L \sin d\theta_L)}{\cos\theta_L \cos d\theta_L - \sin\theta_L \sin d\theta_L}. \quad (28)$$

Subtracting (27) from (28), assuming $\sin d\theta_L$ is $d\theta_L$, $\cos d\theta_L$ is 1, and neglecting a $d\theta_L \sin\theta_L \cos\theta_L$ term in the denominator and all second order increments, the following results:

$$df \doteq \frac{4\pi^2 f^2 L_{eq}}{2\pi(4\pi^2 f^2 L_{eq} C_{eq} + 1)}\left(\frac{G_L d\theta_L}{\cos^2\theta_L} - 2\pi f dC_{eq}\right). \quad (29)$$

The increment dC_{eq} is equal to the increment of c_{22} resulting from a change of grid-to-plate region space charge (denoted by $d\psi_2$). Therefore,

$$df \doteq \left(\frac{2\pi f^2 L_{eq}}{4\pi^2 f^2 L_{eq} C_{eq} + 1}\right)\left(\frac{G_L d\theta_L}{\cos^2\theta_L} - \frac{4\pi f c_{22} d\psi_2}{3}\right). \quad (30)$$

A study of (30) yields the following with regard to keeping incidental fm low.

1) The plate-line conductance G_L should be small; this means light coupling to the load.
2) The change of the phase angle of the plate-line admittance $d\theta_L$ should be small; this is accomplished by use of the compensation for the cathode-to-grid transit angle θ_1 previously discussed.
3) $\cos\theta_L$ should be large; this means the plate circuit should be near resonance, although this condition is interdependent on the compensation for θ_1.
4) The component of $d\theta_L$ caused by change of the grid-to-plate transit angle θ_2 should have the same sign as the change of grid-to-plate region space charge, $d\psi_2$; these increments would then tend to cancel. This condition warrants further discussion, which is forthcoming.

Condition 4) gives information as to the proper modulation system for low incidental fm. With plate modulation, as plate voltage goes up, for example, both ψ_2 and θ_2 decrease; this means $d\theta_L$ is positive and $d\psi_2$ is negative causing additive frequency changes. Furthermore, the change of θ_2 is large since the necessary change of V_2 is large. However, with cathode or grid modulation as current goes up, for example, ψ_2 increases while θ_2 decreases. The frequency changes associated with these variations not only are small, they also tend to cancel.

Experimental Verification

Conclusions 2) and 4) of the previous section were experimentally verified for an oscillator, using a 5876 pencil triode, operating at 400 megacycles. A cross section of the test oscillator is shown in Fig. 6, and a photograph is shown in Fig. 7.

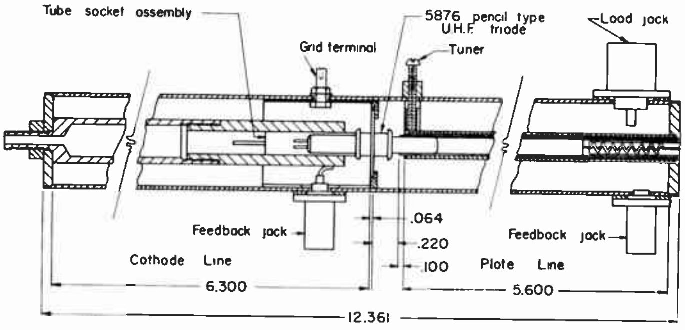

Fig. 6—Cross section of 400-megacycle test oscillator.

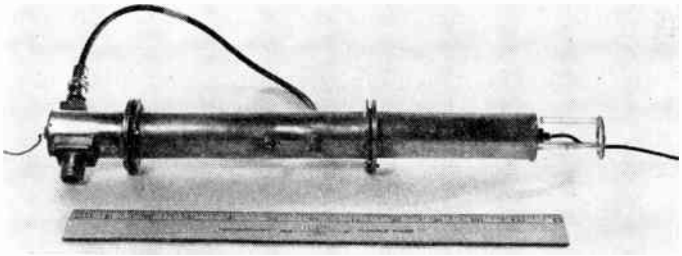

Fig. 7—Photograph of 400-megacycle test oscillator.

Cathode modulation was accomplished by placing a modulating sine-wave signal in series between the cathode and the negative terminal of the plate-power supply. Plate modulation was accomplished by feeding the modulating signal through a capacitor to the plate; a choke was used in series with the plate-power supply.

Theoretical frequency changes were calculated by use of (30). Some of the terms of (30) were obtained experimentally because discontinuities in the various coaxial lines made theoretical calculations prohibitively difficult. The transit times and space charge were obtained from previously discussed formulas applied to the 5876 pencil triode. Even though the 5876 has cylindrical electrodes, parallel-plane formulas were used; this is valid because of the low ratio of electrode diameters in the 5876.[6]

[6] W. R. Ferris, "Input resistance of vacuum tubes as ultra-high frequency amplifiers," Proc. IRE, vol. 24, pp. 82–105; January, 1936.

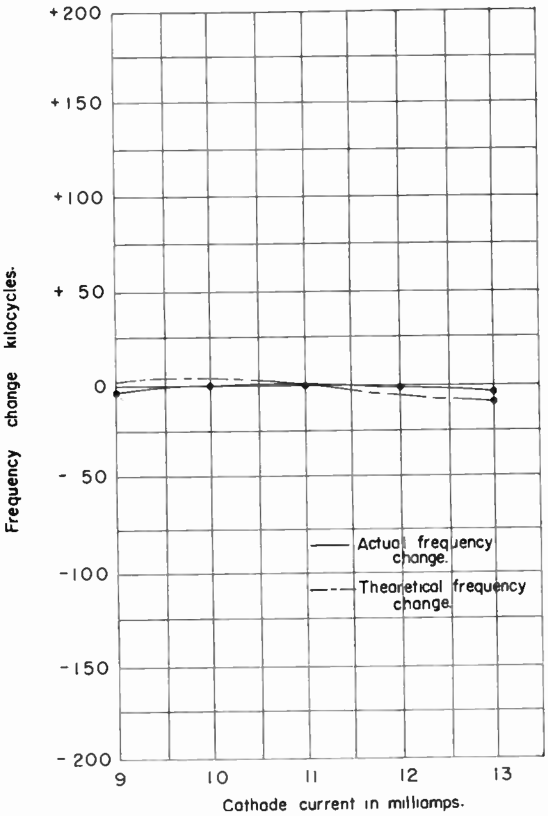

Fig. 8—Frequency change vs cathode current for cathode-modulated 400-mc oscillator adjusted for compensation.

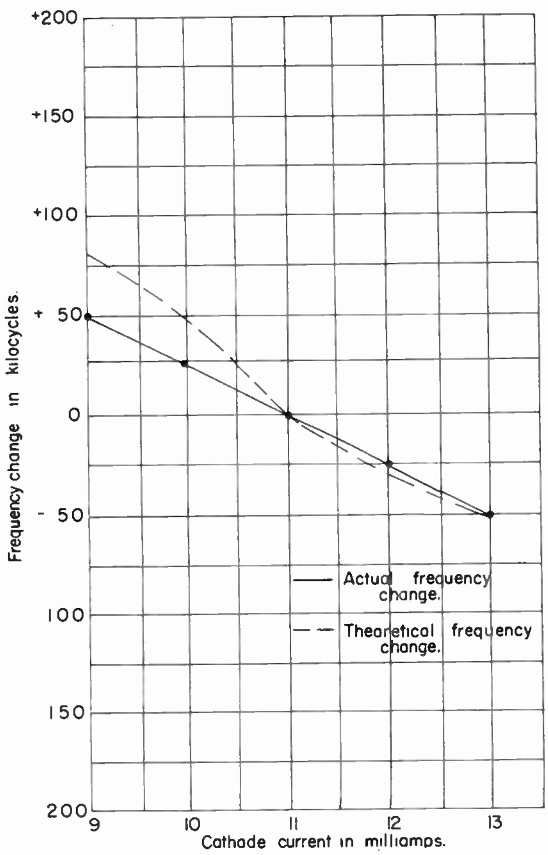

Fig. 9—Frequency change vs cathode current for cathode-modulated 400-mc oscillator not adjusted for compensation.

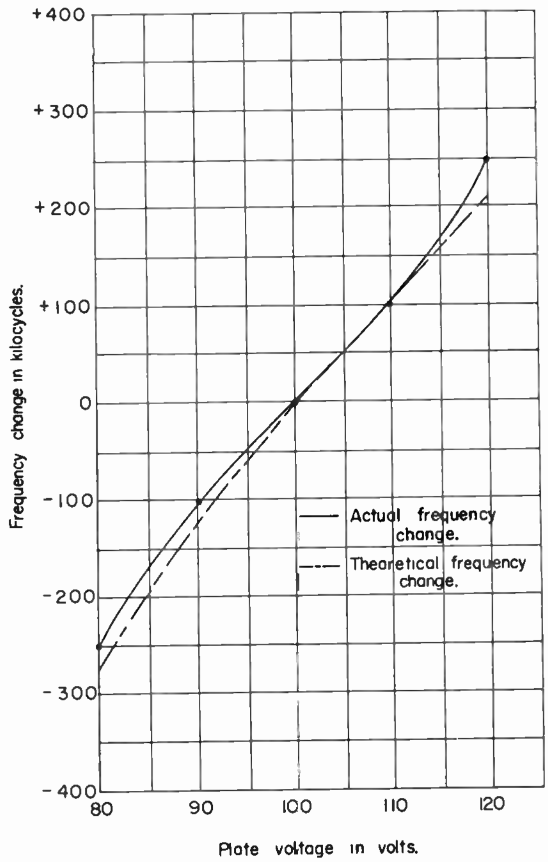

Fig. 10—Frequency change vs plate voltage for plate-modulated 400-mc oscillator adjusted for compensation.

To show the effectiveness of the methods described in this paper, sample data are presented in graphical forms. Fig. 8 shows, for the oscillator adjusted to compensation ($A = A_d$), the frequency changes resulting from a variation of cathode current from 9.0 to 13 ma under cathode modulation. The solid line shows the actual frequency changes while the broken line shows the theoretical changes. Fig. 9 shows the frequency changes for the same variation of cathode current occurring when the oscillator is not set at compensation ($A > A_d$). It is clear that compensation is quite effective in reducing the incidental fm. The incidental fm obtained for plate modulation is shown in Fig. 10. Frequency changes were large in spite of the fact that A was set equal to compensation A_d.

Conclusion

In summarizing, two facts stand out as important for the reduction of incidental fm at uhf. One fact is the advantage of cathode or grid modulation over plate modulation. The frequency changes caused by the two transit-time effects in the grid-to-plate region tend to cancel for cathode and grid modulation; whereas, for the plate modulation they add. The second fact is that a compensation for change of cathode-to-grid transit time is possible by a proper adjustment of the cathode circuit and feedback line. These adjustments are best made experimentally. A suggested procedure for this is:

1) Amplitude modulate the oscillator about 10 per cent by applying a square wave to the cathode.
2) Mix the oscillator output with an external cw signal of the desired frequency, and observe the resultant beat pattern on an oscilloscope.
3) For a given feedback-line length adjust the cathode-line length and plate tuning control until a zero beat appears simultaneously on top and bottom of the square wave. Care must be exercised as the adjustment of cathode-line length is fairly critical.

If simultaneous zero beats are not obtainable either the feedback line is not of proper length and should be changed or the frequency is too high for perfect compensation. In the latter case the incidental fm can be minimized by adjusting the cathode-line length so that the two zero beats are as close together in frequency as possible.

The utilization of such a procedure in the type of uhf oscillator described in this paper should producea signal relatively low in incidental fm.

Acknowledgment

The author wishes to express appreciation to Dr. C. W. McMullen of Northwestern University for his valuable aid and direction.

Improved Keep-Alive Design for TR Tubes*

LAWRENCE GOULD†

Summary—The conditions leading to crystal deterioration and burnout in tr tubes, which exhibit apparent satisfactory leakage characteristics, are investigated experimentally. The characteristics of coaxial keep-alive structures are evaluated by monitoring with probes and light intensity measurements the residual electron density in the interaction gap spacing. The prime factor leading to crystal deterioration is random fluctuations in the electron density which are attributed to a wandering of the keep-alive discharge along the cone wall. During the period of fluctuation the spike energy can become excessive. An improved keep-alive design, in which the cone wall is insulated and the cathode is fabricated from stainless steel, eliminates this effect. TR tubes with the improved keep-alive structure have successfully operated in excess of 1000 hours without any apparent crystal deterioration.

Introduction

ONE IMPORTANT function of a tr tube in a pulse radar system is the protection of the associated crystal receiver from electrical deterioration and burnout. The degree of protection provided by the tr tube during the time interval of rf high-power transmission is determined by the rf leakage power. The prime damage to the crystal results from the leakage energy during the period in which the rf discharge is being formed. This energy is referred to as the "spike" energy.

A dc keep-alive discharge must be placed in the tube in order to reduce the values of spike leakage energy so as to provide adequate crystal protection. In general, failure of the keep-alive discharge will result in excessive leakage energy, causing crystal deterioration and burnout. Recently, sufficient evidence has been accumulated which substantiates that tubes possessing the required low values of spike energy as measured with a thermistor bridge still do not protect the crystal against deterioration and burnout.

A recent article[1] attributed crystal deterioration to the occurrence of occasional transitions from a glow to an arc of the keep-alive discharge. If an arc discharge does occur, the voltage across the resistor in series with the discharge increases to the extent that the discharge is cut off momentarily. During this time interval, the spike energy may rise to values sufficient to cause crystal damage.

The investigation described here demonstrates that as severe a condition as a glow to arc transition is not required for crystal impairment. The work was performed at 9300 mc on coaxial keep-alive structures in broad-band tr tubes.

Electron Density Measurements

The keep-alive assembly in a conventional tr tube consists of a keep-alive electrode placed coaxially within one of the hollow truncated cones which forms part of the tube filter structure. The keep-alive electrode is made of kovar and is glass sheathed up to its tip, where a bare metal area of about 0.020 inch in diameter is left, forming the cathode for the glow discharge. The tip of the keep-alive lead is 0.040 inch away from the tip of the cone. The inside cone wall forms the anode for the glow discharge.

* Original manuscript received by the IRE, October 5, 1956; revised manuscript received, December 13, 1956. This paper is based upon work performed for the U. S. Army Signal Corps, on Contract DA-039-SC-42720.

† Microwave Associates Inc., Boston, Mass.

[1] T. J. Bridges, P. O. Hawkins, and D. Walsh, "Keep-alive instabilities in a tr switch " Proc. IRE, vol. 44, pp. 535–538; April, 1956.

The prime function of the keep-alive discharge is to provide the requisite electron density in the microwave discharge region of the filter structure: namely, the interaction gap between the cones. Measurements of the electron density in the interaction gap as a function of experimental conditions and operation on life were performed to evaluate the behavior of the keep-alive structure. Observations of the light intensity from the plasma in the interaction gap were made with a photomultiplier. The light intensity, which was correlated with dc probe measurements, yields a relative value of electron density and is convenient for rapid measurements. It was noticed that, at times, the electron density decreased momentarily by a factor of two or more, while the keep-alive current and voltage changed by only a few per cent. A closer inspection into the nature of the discharge within the cone, by placing holes along the cone wall, indicated that the decrease in electron density was associated with a movement of the discharge away from the tip of the cone towards the keep-alive lead.

The light intensity was monitored as a function of time for keep-alive operation at 100 microamperes current. A gradual decrease in electron density with time and random fluctuations in electron density after about 100 hours of operation were observed. These effects can be explained qualitatively as follows. The electron density, penetrating through the hole of the cone tip, is very sensitive to the discharge position along the cone wall. The gradual decrease of electron density on life was attributed to a change in the characteristics of the gas, since refilling the tubes with a fresh gas returns the electron density to its original value. Impurities liberated from the cathode and surrounding walls, the presence of oil vapor, and insufficient outgassing of the tubes are possible causes for the observed phenomena. The random fluctuations in electron density can be attributed to changes in the cathode characteristics. A more stable cathode than kovar such as a rhodium suppresses the degree of fluctuation. Erratic changes in the electron secondary emission from the cathode can be produced by oxidation of the cathode by sharp irregularities of the surface which may become temporarily heated by ionic bombardment and then removed by sputtering.

Verification of Crystal Deterioration

The following approach was established to verify that the random fluctuations in electron density, superimposed on the gradual decrease of electron density, are sufficient to produce crystal impairment although the average values of leakage energy are satisfactory. A conventional 5863 tr tube was placed on dc life test at a keep-alive current of 100 microamperes. At various intervals, the spike leakage energy was measured as a function of light intensity. The results for the measurements up to 140 hours of dc life are shown in Fig. 1. At this time, the light intensity became unstable and was

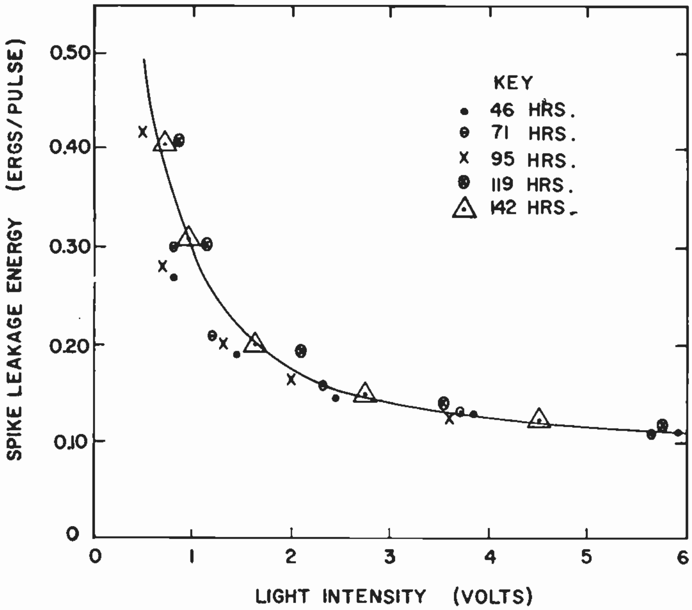

Fig. 1—Light intensity as a function of time on dc operation.

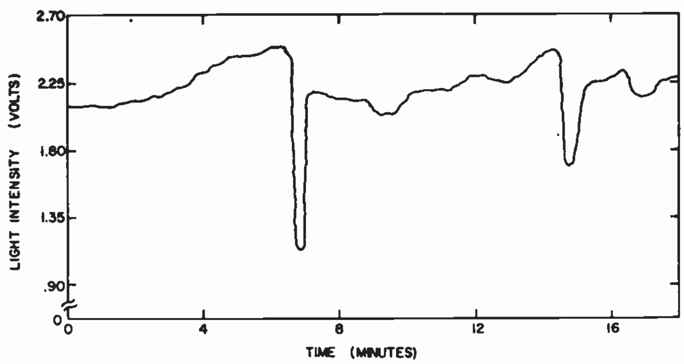

Fig. 2—Spike leakage energy as a function of light intensity.

recorded on an automatic Honeywell Brown recorder. The recording of light intensity as a function of time is shown in Fig. 2. The lowest value of light intensity during a fluctuation is the desirable quantity since this determines the largest value of spike energy through the tube. The recorder, because of its slow response time, gives an approximate indication of the decrease in light intensity, although the actual decrease may have been greater. From Fig. 1, it is seen that the spike energy increases rapidly as the light intensity drops below 2 volts. The average spike energy for this tube was 0.16 ergs per pulse, whereas, from Fig. 2, the maximum spike energy is at least 0.28 erg per pulse if not greater. According to JAN specifications, the maximum allowable spike leakage energy is 0.2 erg per pulse. Thus this tube exhibited the potential characteristics for producing crystal deterioration.

The tube was placed on rf high-power operation with a crystal mounted behind it. After 42 hours, deterioration was evident. A new crystal showed deterioration after 23 hours of additional tr operation. The average spike energy during this time interval remained approximately unchanged. Tests of this kind were performed on other tubes with the same results. Thus, the fluctuation of electron density in the interaction gap due to the wandering of the keep-alive discharge along the cone wall can produce crystal deterioration even though the spike energies are within specification.

Improved Keep-Alive Design

The simplest technique for minimizing the wandering of the discharge and confining it to the tip of the cone is to insulate the inside of the cone except for the region in the cone tip. This was accomplished by inserting an insulating bushing, fabricated from synthetic mica, into the cone. Synthetic mica,[2] technically known as hot-pressed synthetic fluor pholgopite mica, has properties similar to natural mica, but is machinable. The improved keep-alive assembly is shown in Fig. 3. The results of a materials investigation showed that a high chrome content stainless steel, #431, is very satisfactory as a stable cathode material. Since the lead need not be insulated in the vicinity of the tip, the lead tip was fabricated from stainless steel, 0.020 inch in diameter, and was welded to the kovar.

For optimum keep-alive design, the spacing between the end of the keep-alive lead and the cone tip is of the order of 0.050 to 0.080 inch. Measurements indicate that the requisite electron density in the interaction gap is obtained when the mica tip to cone tip spacing is of the order of 0.005 inch. With these dimensions and structure, the gradual decrease and rapid fluctuations in electron density on life, which are so detrimental to adequate crystal protection, are suppressed and, to a greater extent eliminated. Additional advantages of the improved structure are:

1) The probability of keep-alive shorts is minimized,
2) The centering of the lead within the structure is not critical,
3) The distance from the lead to cone tip is not critical, and
4) The fabrication of the assembly is simplified.

5863 tr tubes containing the improved keep-alive structure were life tested at 200 kilowatts peak rf power with a pulse width of one microsecond and a repetition rate of 1000 cps. The gas fill in the tubes was eight to ten mm Hg of argon and two mm Hg of water vapor. IN23C crystals were mounted behind the tubes with a crystal in a dummy mount as a control. All tubes showed no crystal deterioration for a minimum of 1000 hours of operation. An extended life test on two tubes showed no crystal deterioration for 5000 hours. After this time, crystal deterioration occurred. The exact nature of the deterioration was not established. The operating characteristics of the tubes were within specification throughout the entire life period.

[2] Manufactured by Brush Beryllium Co., Cleveland, Ohio.

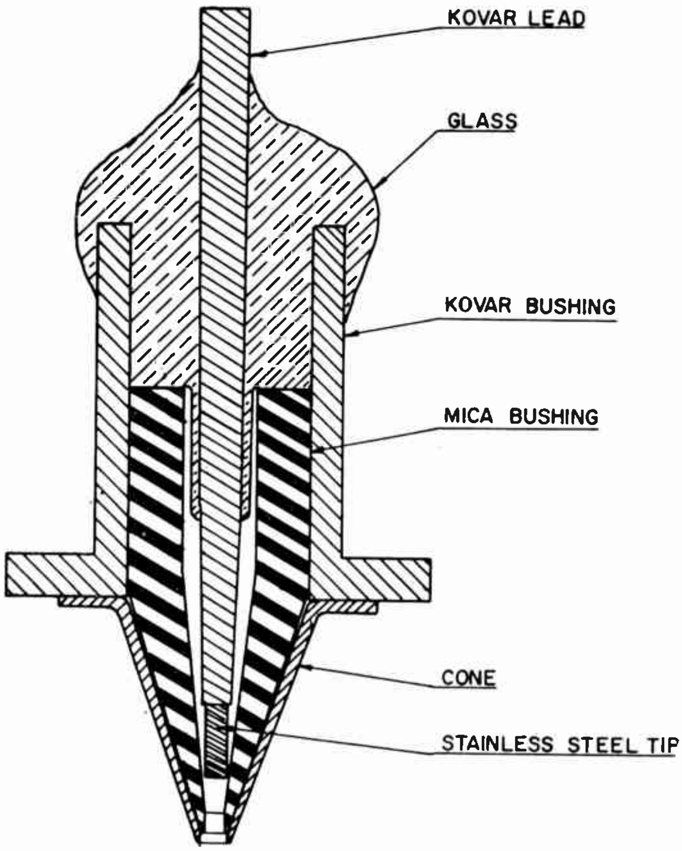

Fig. 3—Improved keep-alive structure.

Optimum Gas Pressure

Measurements of the pressure dependence of spike leakage energy indicated that erroneous values for the optimum gas fill can be obtained with the conventional keep-alive structure. Fig. 4 shows the data of spike energy as a function of gas pressure for constant values of keep-alive current with a conventional keep-alive structure. The gas consisted of two mm Hg of water vapor and varying amounts of argon. The curves indicate that the leakage power is a minimum at a gas pressure of five to six mm Hg. However, such data lead to false conclusions. The reason for the apparent increase in leakage energy with increasing pressure is because of the rapid decrease in the electron density in the interaction gap with increasing pressure. Fig. 5 shows the light intensity and hence, electron density in the interaction gap as a function of gas pressure for the conventional keep-alive at constant current. The decrease in electron density with increasing pressure is attributed to a movement of the discharge up along the inside cone wall toward the keep-alive lead.

Curves of spike leakage energy as a function of gas pressure for constant values of light intensity, using the improved keep-alive structure are shown in Fig. 6. It is evident that the actual minimum leakage energy occurs at a total gas pressure of 14 to 16 mm Hg. The apparent minimum at the lower pressure with conventional structure comes about because the decrease in leakage energy for constant electron density in the gap is over-

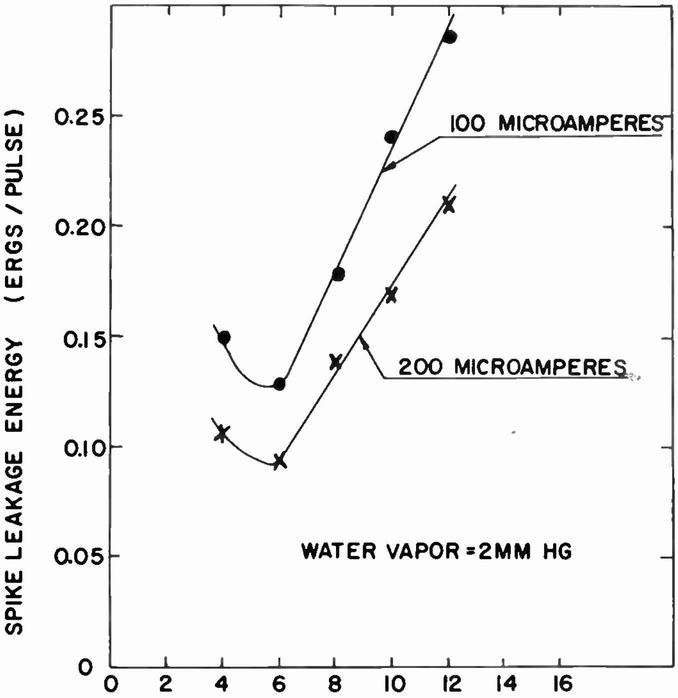

Fig. 4—Spike energy as a function of gas pressure for constant keep-alive current.

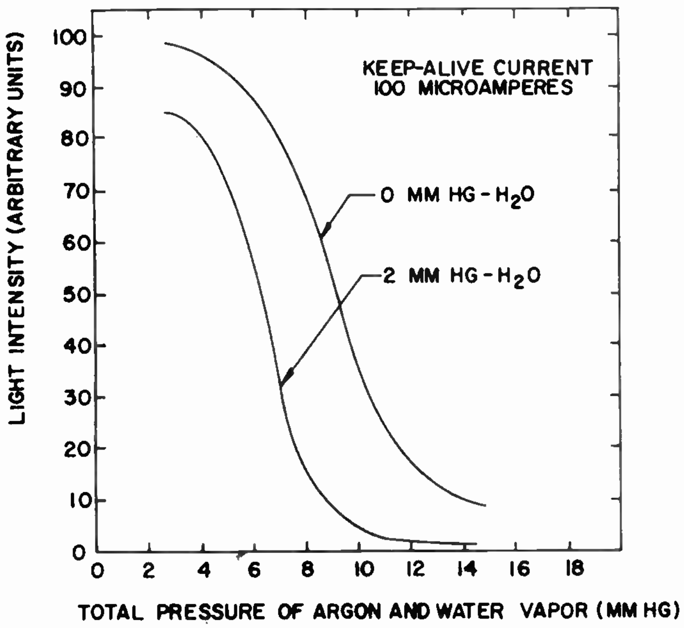

Fig. 5—Light intensity as a function of gas pressure for the conventional keep-alive structure.

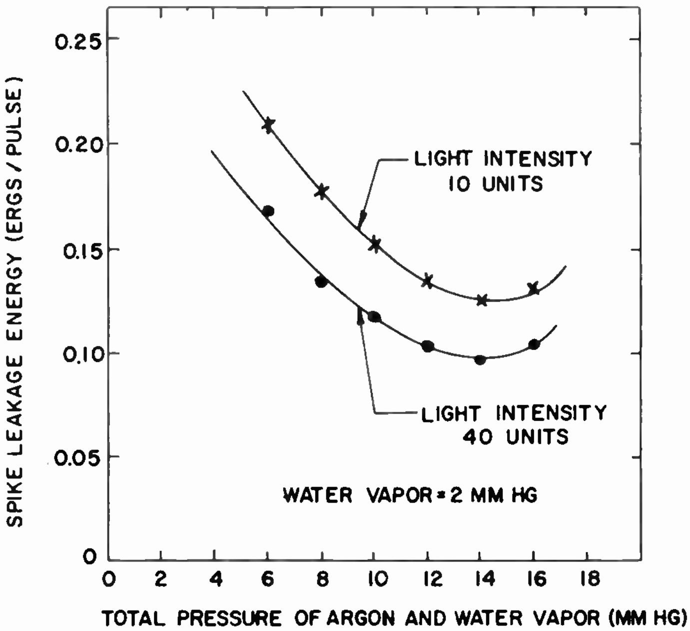

Fig. 6—Spike leakage as a function of pressure for light intensity.

shadowed by the increase in leakage due to the rapid decrease in electron density for constant keep-alive current as the pressure is increased. Life tests at the higher pressures show that the tubes possess greater stability and longer life than at the lower pressures.

Conclusion

An investigation into the ability of conventional keep-alive structures to provide the value of electron density in the interaction gap required for low spike leakage energies over the life time of the tube shows evidence of faulty keep-alive behavior. The prime factor leading to crystal deterioration is the rapid random fluctuations in electron density which are attributed to a wandering of the keep-alive discharge along the cone wall. During the period of fluctuation, the spike energy can become excessive and produce crystal deterioration. A condition as severe as a glow to arc transition is not required to explain the occurrence of crystal failures. The improved keep-alive design maintains an electron density in the interaction gap which is fairly insensitive to conditions and enables operation at higher pressure. TR tubes with the improved keep-alive structure have successfully operated at 200 kilowatts peak rf power in excess of 1000 hours without any apparent crystal deterioration.

Discussion of the Single-Sideband Issue

Double Sideband vs Single Sideband Systems*

With regard to the December issue of PROCEEDINGS devoted to Single Sideband, I have some comments which some of the readers may wish to consider. To begin with, I am in the uncomfortable position of having submitted a paper which presents a point of view which is diametrically opposed to the very spirit of this particular issue. I am sincerely grateful to the IRE and to the reviewers for permitting me the opportunity to express my views in print.

The point I have attempted to make in my paper is that AM, if brought up to date by giving it the same advantages (carrier suppression, improved frequency stability, and improved receiving techniques) proposed for SSB, would be by far the more desirable system for the large majority of communications applications. After reading the December issue I am more convinced than ever of the validity of this point of view. In introducing the material which follows, let me say that never in engineering do we expect to get something for nothing. The elimination of one sideband from a double-sideband signal may have certain advantages, but there is also a price to be paid in one form or another. The advantages (both real and imagined) have been emphasized and re-emphasized, but the disadvantages and penalties have either barely been mentioned or have been completely ignored by most writers.

Spectrum Conservation

In this discussion let us dispense with the usual cries of alarm concerning the congestion in the hf spectrum, which normally serve as an introduction to most SSB presentations. The congestion problem is real and it is serious; there can be no argument on this point. The claim that SSB will offer significant relief in this area over DSB represents, in my opinion, a false hope. *A significant increase in usable channels cannot be obtained by use of SSB except in those very special communications applications where the dynamic range of received signals can be controlled.* To explain further, consider the paper by Brown[1] which is concerned with mobile applications of SSB. In Fig. 4 of this paper is shown a frequency allocation for eight mobile units calling in to a common central station receiver. Each mobile transmitter is SSB with 5-kc spacing between SSB carrier frequencies. Brown points out that there may be as much as 100 db difference in the level of received signals at the central station due to the changing positions of the mobile transmitters. He is rightly concerned about receiver overload problems, but even if we assume an ideal central station receiver, a much more serious problem is encountered. If we can assume that in the future 50 to 60 db suppression of the unwanted sideband in SSB can be consistently maintained, we are still faced with intolerable adjacent channel interference in situations such as this. Obviously, 50 to 60 db of sideband suppression will result in adjacent-channel interference which exceeds the *desired* signal in that channel by 40 to 50 db if there exists 100 db difference in received signal levels. The same situation exists, of course, for aircraft or for any other mobile communications situation. In short, 50 db of sideband suppression may sound impressive (and indeed it represents an outstanding engineering accomplishment) but it is not nearly good enough to allow the adjacent channel to be used with any reliability in many important situations. Thus, in applications where received signal dynamic range cannot be controlled, conversion to SSB will result in considerable inconvenience in terms of cost and complexity for the user with only a marginal, if any, increase in the number of usable channels.

Of course, there are situations such as telephone carrier systems where dynamic range can be carefully controlled and indeed SSB is used here to good advantage. It is a mistake, however, to use these special situations as examples of how SSB can save spectrum space without considering the applications.

There is another approach to spectrum conservation which has not been getting too much attention of late, but which has far greater potential than SSB. We refer here to signal processing techniques which reduce required bandwidth by removing redundancy from the original intelligence signal. The bandwidth reduction in this area for speech, for example, can be easily 10/1, and 100/1 ratios are not at all unlikely. Thus, the ultimate 2/1 reduction offered by SSB begins to look rather small compared to what can be accomplished by efforts in other areas. For example, use of teletype in place of voice for routine data handling yields a bandwidth reduction of about 30/1. The argument that these signal processing techniques can be used with SSB for an additional bandwidth reduction is not necessarily sound. Many of these reduction techniques require waveform preservation in transmission and conventional SSB systems do not preserve waveform. Thus, it should be remembered that SSB is not the *only* method for reducing required channel bandwidth.

Power Gain

Much has been said by the various authos as to the power advantage which is obtained through the use of SSB techniques. The figures quoted for this power gain appear to have a lower bound of about 9 db. (Figures much larger than 9 db are also often seen in print. After some investigation it appears that the power gain of SSB over AM is approximately $(9+N!)$ db, where N represents the number of co-authors of a particular paper.) This confusion as to the exact amount of the power gain and as to the exact reason for its existence reminds me of the statement someone once made concerning the law of errors. "Everybody believes in the Law of Errors, the experimenters because they think it is a mathematical theorem, the mathematicians because they think it is an experimental fact." Much the same situation apparently exists with regard to the power gain of SSB over AM. This is an important matter because I feel that there does not exist, nor has there ever really existed, any power advantage of SSB even over the old AM system with full carrier. In computing the 9-db figure, a comparison is made which assumes *sine wave* modulation and equal peak powers for both SSB and AM, with a full carrier assumed in the AM system. Under these conditions the average SSB power is indeed 9 db higher than that of the sidebands of the AM signal, and a 9-db improvement in signal-to-noise ratio will result for SSB over AM if optimum receiving techniques are employed in both cases. Note that bandwidth did not enter into this consideration, which is as it should be. In the first place, some of the assumptions made in arriving at the 9-db figure are entirely unwarranted since, for example, pure sine wave modulation is seldom employed when intelligence is to be conveyed. If we had used a square wave instead of a sine wave as the modulating voltage, we would have found that AM would have a sizable power advantage over SSB for a given peak power limitation. It can be shown that a perfect square wave cannot be transmitted through an SSB system without requiring infinite peak power as well as infinite bandwidth. This can be shown mathematically quite easily but the point to be made here is that the waveform of the modulating intelligence is of paramount importance in determining the peak-to-average power output ratios in SSB and AM. The sine wave happens to be a member of that very limited class of modulating waveforms which gives SSB all of the advantages. It should always be remembered that the rf envelope of an SSB signal bears little or no direct relationship to the modulating waveform, whereas the rf envelope and modulating waveform are directly related in DSB or AM. This means that certain types of modulating waveforms can produce very large peak SSB envelope amplitudes even though the amplitude of the modulating waveform is well behaved. The square wave, or for that matter any flat-topped wave with a rise time which is small in proportion to the pulse duration, will produce high peak voltages in the resultant SSB signal. This means that the simple expedient of speech clipping and filtering which can easily increase the average transmitted power in a DSB or AM signal by at least 10 db cannot be used by SSB. If an attempt is made to use such straightforward techniques with SSB it will be found that very little if any actual improvement in peak-to-average power ratio will result. Tests have been made and there is every indication that for voice transmission DSB

* Received by the IRE, January 7, 1957.

[1] A. Brown, "Single-sideband techniques applied to coordinated mobile communications systems," PROC. IRE, vol. 44, pp. 1824–1828; December, 1956.

may well show a 10-db average power advantage over SSB. This is not to say that the SSB cause is lost with regard to voice communications but rather that different, and from what I have seen to date, far more complicated voice processing techniques must be used in order to make up at least part of the deficit. If we realize that a SSB system is far more complicated than an equivalent DSB system, it becomes a bit unnerving to think of complicating the SSB system even more in order to gain back at least some of the power advantage lost when speech transmission is a requirement.

I, therefore, object to the indiscriminate claim that there exists any power advantage whatsoever for voice transmission in favor of SSB over DSB or even the old AM system with full carrier. It should be clear that any power advantage claim is meaningless without specifying the modulation waveform to be used, since, as we have seen, certain types of waveforms such as pulses or clipped speech put SSB in a position of definite disadvantage.

Waveform Preservation

The problem of *waveform preservation* which was mentioned in the paper by Honey and Weaver[2] deserves, I think, a considerable amount of attention. I have gotten the impression that in considering SSB most engineers tend to think in terms of voice transmission. While it is true that voice is certainly a consideration, it is also true that voice will be used less and less in future communication systems. Thus, we must turn our attention from voice to other types of transmission such as binary data and consider the problems that may arise from the universal use of SSB. Waveform distortion, although it is not particularly serious when voice is considered, becomes all-important when other types of data are to be transmitted. The transmission of a pulse by means of AM or DSB systems is so simple and so easily done that we may tend to forget that this will become a major accomplishment if SSB transmitting and receiving techniques are employed. It is a revealing experiment to transmit pulses through an SSB system in which stable clocks are being depended upon to maintain the proper frequency at each end of the system. We are not thinking here of the peak power considerations which were mentioned earlier but merely of waveform considerations. The first thing one discovers when such an experiment is performed is that absolutely no frequency error can be tolerated in the SSB system. As a matter of fact, not only must the frequency of the receiver oscillator be correct but its phase must be exactly controlled, for a 90° change in demodulator oscillator phase will result wondrous new and different waveforms from those seen in the 0° phase position. So, we quickly discover that we can no longer depend upon accurate clocks at each end of the system for pulse transmission but we must establish an exact phase lock at the receiver. Once this phase lock is established, it will in general be found that a considerable amount of phase compensation must be employed to correct for the phase distortion of the sharp SSB filters at the transmitter and the receiver. I do not say that this cannot be done, but I merely wish to point out that what was so easily accomplished with AM or DSB now requires even more complications to be added to the already complicated SSB system. This in my mind is a very serious situation for it means for example that time division multiplex signals could never be put through an SSB link which was designed for voice operation. The same applies for slow-scan tv and a host of other signals which require waveform preservation in transmission. SSB has been used to transmit keyed subchannel tones but it must be realized that in this case the keyed tones themselves are DSB transmissions and the SSB system merely serves as a frequency translator. It is further true that under severe multipath conditions frequency division multiplex schemes having a slow data rate per subchannel may be desirable and SSB can certainly handle this type of transmission. The point still remains, however, that this is about the only way that SSB can handle data. Thus, we see that far from being a general purpose communication system, SSB will impose severe restrictions as to the type of signals which can be handled. This last fact, I believe, represents a serious limitation in the widespread use of SSB for general purpose communications.

Propagation

I have often heard the argument that the DSB system requires a coherent addition of the two sidebands in order for this system to make full use of transmitted signal power. It has been further argued that in a turbulent medium coherent sideband addition cannot be obtained and hence SSB will be preferable under these conditions since there is no dependence on the sideband-to-carrier phase with regard to the amplitude of the detected sideband. Although this line of reasoning appears on the surface to have some merit, actual tests as reported in my paper did not justify this criticism. AM signals received over long paths when conditions were very poor have shown very much the same results when DSB synchronous detection was employed as compared to SSB reception of the same signals and at the same time. As a matter of fact, on some occasions it was noted that DSB detection gave decidedly better results than SSB detection. The following analysis, I believe, may tend to explain these results and I hope give pause to those who would automatically assume that SSB is better than DSB for the reception of signals propagated through very turbulent media.

The assumptions to be made are for the most violent turbulence imaginable short of a complete dropout. We shall assume that every sideband component transmitted through the medium suffers amplitude variations according to a Rayleigh distribution. We shall further assume that one sideband component bears no recognizable phase relationship to any other sideband component and that the amplitude variations are also completely random and independent. SSB and DSB transmissions are to be compared with the average received signal power in both cases equal to $2\sigma^2$. In the SSB case the demodulated audio vector will, of course, have an amplitude distribution which is Rayleigh and a resulting distribution function which may be written as:

$$F_{\mathrm{SSB}}(x) = 1 - \epsilon^{-x^2/2\sigma^2}. \quad (1)$$

Now in the DSB case with absolutely no phase control assumed, the receiver output will be composed of two audio vectors (representing the demodulation of each of the two sideband components) having the same frequency but a completely random relative phase relationship. The resulting audio vector can easily be shown to be Rayleigh distributed in amplitude with a distribution function given by:

$$\underset{\mathrm{no}\phi}{F_{\mathrm{DSB}}}(x) = 1 - \epsilon^{-x^2/2\sigma^2}. \quad (2)$$

A very interesting thing about (1) and (2) is that they are identical. At first hand this might seem strange but a little further thought shows that in order to get a null in audio output in the DSB case the two audio vectors must be in phase opposition to each other and at this time must have equal magnitudes. Thus, we see that the conditions for an SSB fade are simply a fade of the one received sideband whereas in the DSB case two conditions must be satisfied for a fade; namely, proper phase orientation and equality of amplitudes for the two received sidebands.

If we now make the assumption that although there exists no phase coherence between the two sidebands in the DSB case, the phase variations are slow enough to be followed by the phase control system, we will find that the two audio vectors previously considered now always add in phase to one another giving a distribution function for the amplitude of DSB receiver output as given by

$$\underset{\mathrm{opt}\phi}{F_{\mathrm{DSB}}}(x) = (1 - \epsilon^{-x^2/\sigma^2}) - \frac{x}{\sigma}\epsilon^{-x^2/2\sigma^2}\sqrt{\frac{\pi}{2}}\, I\left(\frac{x}{\sqrt{2}\sigma}\right) \quad (3)$$

where

$$I(y) = \frac{2}{\sqrt{\pi}}\int_0^y \epsilon^{-t^2}dt. \quad (4)$$

Eqs. (1)–(3) are plotted in Fig. 1 (next page) for comparison. If it is assumed that the received signal in either case is not usable when it drops below a certain threshold value, the plot of the distribution functions will yield immediately the relationship between probability of drop-out and threshold setting. The upper curve represents both SSB performance and DSB performance when no phase control attempt is made or possible. The lower curve represents the performance of a DSB system with optimum phase control. It is clear from Fig. 1 that the use of DSB modulation through a highly turbulent medium guarantees at least SSB performance since SSB performance represents the worst that can ever happen to the DSB system. The lower curve, of course, also represents the performance of two SSB systems in a frequency diversity arrangement if optimum phase control is made in the recombination process. These results confirm very well some

[2] J. F. Honey and D. K. Weaver, Jr., "An introduction to single-sideband communications," Proc. IRE, vol. 44, pp. 1667–1675; December, 1956.

of usable channels in our already overcrowded spectrum. Indeed we even see the statement that SSB will *more* than double the number of usable channels. The "more than double" statement is apparently founded on the premise that the improved stability necessary for SSB operation (and DSB operation for that matter) will result in a reduction of the presently required guard bands. Now this is quite true, but in effect what these people are saying is that one of the advantages of SSB is increased frequency stability when in actual fact increased frequency stability is a *requirement* rather than an advantage. Improved frequency stability can result in narrower guard bands for any modulation system, not just SSB alone. Secondly, the 2-to-1 increase argument, I believe, does not hold in many of the important applications which are being considered due to large variations in dynamic range of received signals. I am quite aware that there may be some advantage for some services from a frequency conservation point of view through the use of SSB but I cannot conceive of any gain on the order to 2-to-1 in this regard. It would be worthwhile indeed for some disinterested group, familiar with allocation problems, to evaluate the spectrum conservation advantage which can be expected from SSB for various applications. Until this is done, I feel that the proponents of this system should refrain from even a 2-to-1 claim.

I should like to mention a few items concerning military applications of DSB and SSB although this is obviously a difficult subject to discuss due to security restrictions. It would appear to me that many of the arguments concerning spectrum utilization which might be used in civil applications cannot be carried over without modification into the military communications field. It is rather obvious that there will be no FCC protection in a combat area. In this regard it would appear that the effectiveness of our military communications would be directly affected by the "talking power" of equipments having, in many cases, very restricted space and weight limitations. In this respect, I should like to point out again that DSB transmitting equipment is even simpler than standard AM transmitting equipment. Further, the use of clipping and filtering in DSB can result in a sizable power advantage over an equivalent SSB transmitter even if the original complexity argument is set aside. Thus, it would seem that for tactical use DSB would be far preferable especially when one considers that the performance under actual combat conditions is the final payoff.

A more philosophical question to be asked with regard to military communications would be the availability of adequately trained operator and maintenance personnel. SSB equipment is admittedly complicated and expensive and if proof of this last statement is required, I should only like to cite the very limited commercial use that has been made of this technique in spite of its well-known advantages over conventional AM. If the technical personnel shortage in the military is as serious as it has been described to me, it would appear that the present problems will be compounded with the adoption of a complex system such as SSB for general communications use. With the rapid strides that other nations are apparently making in technology it would appear that we have passed the time when increased complexity in our military equipment is a point of minor consequence. We can no longer afford to squander the technical talent we now have on a complicated communications system when a much simpler one will not only do the same job but in many respects will do this job better. In short, our technically skilled people are a precious national asset and the maximum possible efficiency should be realized in their use.

I foresee one interesting development in amateur radio which might prove of considerable significance to the military. For a good many years now various groups have been attempting to convert the radio amateur to SSB and to date these groups have enjoyed some measure of success. Recently a small number of amateurs have been told about DSB and have been using this modulation method. Some of these people have employed clipping and filtering in their equipment and their experiences to date seem to bear out the claim that there is a sizable power gain over SSB to be had. Thus, we may be facing an actual battle between DSB and SSB for survival under conditions that in many respects are not unlike the conditions to be found in a military combat area. Although any amateur operating experience must be interpreted very carefully when applied to areas outside this field, the results of the DSB-SSB battle on the amateur bands bears some watching. This situation will be altered, of course, by any amateur regulations which discriminate against DSB either on an input power basis or by giving SSB exclusive use of certain frequency assignments. Barring this unexpected event we should obtain some relative idea of the merits of DSB and SSB under conditions where simplicity, economy, and effectiveness are of paramount importance.

Finally, let me state that in this controversy there are important factors involved which are not entirely technical. There is a definite psychological problem confronting any individual today who attempts to promote a system which is competitive to SSB. I have had some experience along these lines and have come to the conclusion that much of the objection to any non-SSB system is quite often based on psychological rather than technical factors. To understand this situation we must realize that when SSB was first proposed several years ago as the logical replacement for AM, a group of forward looking technical people grasped the advantages of SSB over standard AM and they naturally became quite excited about the prospects. Many of them threw their support behind SSB and as a consequence these people tend to associate themselves personally with this particular modulation process. Consequently, I have come to the conclusion that the vast majority of those who promote and defend SSB are forward looking people who have seen the advantages of a new system and who are anxious to put this new system into general use for the common good. As commendable as this attitude might be, there has been the tendency on the part of many of these people to make somewhat of a "sacred cow" out of SSB. This is unfortunate because true progress will be hindered rather than helped by such an attitude. Thus, both the PROCEEDINGS paper and this letter have been written not for the sake of controversy itself, but in the hope that a technical debate might be carried on in which the *technical* pros and cons of the DSB and SSB systems can be brought forward and examined.

My thanks again to the IRE for the courtesy they have extended to me.

JOHN P. COSTAS
Defense Electronics Div.
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Syracuse, N. Y.

Synchronous Communications*

From theoretical considerations only, I want to reinforce the conclusions and claims presented by Dr. Costas for his system of synchronous communications.[1] Further, it seems to me that his system does not necessarily represent a bandwidth disadvantage with respect to single-sideband systems, even considering the theoretical doubling of channel capacity by single-sideband over conventional AM. The reason for this is that the possibility exists for making each single transmitter-receiver link a double-channel system; then either two independent modulations could be used to occupy the two channels, or a single modulation could be separated into two components for application to a half-bandwidth double-channel rf system. The latter system is attractive enough to qualify as a (partial) eclipsing type considered by Costas.[2]

The double-channel synchronous communications (DCSC) system is provided in the following theoretical manner: in Fig. 2 of Costas' paper the audio phase discriminator is replaced by a four-quadrant multiplier;[3] the local oscillator and 90° phase shifter are replaced by a two-phase frequency-modulated oscillator whose outputs are $\sin(\phi(t))$ and $\cos(\phi(t))$, where the phase $\phi(t)$ is dynamically determined by the closed loop system at rates up to the highest Q channel modulation frequency. Then the output at the summing point of Fig. 2 is the amplitude modulation component, and the frequency of the two-phase oscillator is the frequency modulation component. With linear modulation the frequency control output is the frequency modulation component, so that both AM and fm demodulations are obtained.[4] The possibility of this DCSC system depends on the fact that the synchronous AM demodulation is independent of the frequency of the (suppressed) carrier, so long as the local oscillator is of the *same* fre-

* Received by the IRE, January 7, 1957.

[1] J. P. Costas, PROC. IRE, vol. 44, pp. 1713–1718; December, 1956.

[2] *Ibid.*, p. 1717.

[3] For discussions leading to this interpretation, the writer is indebted to Dr. Lyman W. Orr.

[4] W. C. Moore, "Simultaneous am and fm in rocket telemetering," *Electronics*, vol. 25, pp. 102–105; March, 1952.

quency and phase as the carrier in the I channel of Costas' Fig. 2. This requirement in turn means that the bandwidth of the phase-control loop must be great enough to keep the Q channel output at servo zero. The four-quadrant multiplier behaves as an instantaneous audio phase discriminator. The term *polarity* discriminator is preferable, since not phase but relative Q audio polarity is changed as a result of local oscillator phase change through 90°.

The introduction of a modulation-bandwidth phase-control loop provides a possible double-channel system. This system could be used with two independent modulations, or it could be used as a single-modulation, half-bandwidth, double-channel system in direct competition with the theoretical double-channel capacity of single-sideband systems. The method for deriving the two half-bandwidth signals has been given by Barber[5] and recently by Weaver.[6] Effectively, a two-phase midaudioband local oscillator translates the middle of the input spectrum to zero frequency, so that two-phase, low-pass signals of half-bandwidth are obtained, midway in the systems of Barber and Weaver. These two signals are then applied as amplitude and frequency modulations of a single rf carrier and are demodulated by the DCSC system under discussion. The second half of the Barber filter[5] is then employed. Ideally, the same midaudioband local oscillator frequency and phase as was used at the transmitter is required, in order to restore the low-pass, half-bandwidth, two-phase signals to their proper spectrum positions for addition to regain the single input modulation. However, it is true that small frequency translations are permissible at this point; further, frequency errors less than ten cps are easily maintained at midaudio frequencies.

From the foregoing considerations, it appears to be at least theoretically possible to provide a DCSC system embodying the claims presented by Costas,[1] but in addition requiring only the bandwidth of a single-sideband system. Widening the phase-control loop bandwidth will of course deteriorate lock-in performance under some interference conditions, and interesting problems are introduced by the low-pass requirements on the two half-bandwidth channels. It is not known whether they must be good down to dc, for example. And it is true that even well-stocked shelves do not carry two-phase frequency-modulated oscillators at the present time. Thus no claims are made except to advance a theoretical possibility. Whether it would be easier to instrument as AM and fm as described, or as in-phase and quadrature[7] amplitude modulations with one of the carrier phases tagged is difficult to say. These questions may require the type of investigation now under way on the single-sideband system, for their resolution.

[5] N. F. Barber, "Narrow band-pass filter using modulation," *Wireless Engr.*, vol. 24, pp. 132–134; May, 1947.
[6] D. K. Weaver, Jr., "A third method of generation and detection of single-sideband signals," PROC. IRE, vol. 44, pp. 1703–1705; December, 1956.
[7] D. B. Harris, "Selective demodulation," PROC. IRE, vol. 35, pp. 565–572; June, 1947.

With respect to the advantages of synchronous AM for long-range communications, Costas mentions[2] that there is a lack of a complete explanation for occasional superior performance relative to single sideband. Toward filling this gap, it may be worthwhile considering the multipath conditions the same as a comb of transmission dips and sharp nulls moving back and forth in frequency. The flutter with single sideband could be due to rapid motion of one of these nulls across the transmission band. Constructing a case for synchronous AM, this flutter would be absent due to automatic change-over to synchronous single-sideband demodulation, which occurs when one sideband is seriously reduced in amplitude. The accompanying phase modulation due to the phase characteristic of the comb of dips would result in observable phase changes of the local oscillator toward the phase of the short-time average carrier, as determined by the bandwidth of the phase-control loop. It is also possible that the latter bandwidth in the receiver as designed and used was sufficient to compensate for flutter rates, whereas in the single-sideband receiver the local carrier frequency may not have been capable of such rapid changes, holding steady through the flutter frequency modulation produced by the varying phase characteristic of the moving dips of the comb. This phenomenon would depend on flutter fm being noticeable even though small fixed frequency shifts are not particularly noticeable.

It is interesting as a final remark that the phase-lock system presented by Costas is a most valuable addition to the types of locking system considered by Kallmann.[8]

ROBERT R. MCPHERSON
Ann Arbor, Mich.

[8] H. E. Kallmann, "Single-sideband transmission without transient distortion," PROC. IRE, vol. 43, pp. 485–486; April, 1955.

The DB is the Argument of SSB*

The December issue of PROCEEDINGS is an excellent presentation of many articles on a timely subject that will affect the future of radio transmission techniques. It is also noteworthy for the minimum use of mathematics in the description of single-sideband characteristics.

Relative advantages of single-sideband to amplitude modulation systems as pointed out by Honey and Weaver in their introductory paper[1] varies from 7 to as many as 16 db depending upon the comparison level: 7 db for equal equipment size and weight; 12 db for equal peak antenna voltage; 12 to 16 db for equal radiated power.

* Received by the IRE, December 27, 1956.
[1] J. F. Honey and D. K. Weaver, "An introduction to single-sideband communications," PROC. IRE, vol. 44, pp. 1667–1675; December, 1956.

This last value of 12 to 16 db radiated power gain, assumed with the use of the same final amplifier stage, that has been claimed by SSB techniques is theoretically difficult to realize for the following reason:

Let A be the peak voltage of unmodulated carrier wave, W_m be the angular velocity of modulating wave, R_a be the radiation resistance of antenna.

The average useful power output at 100 per cent amplitude modulation of the carrier wave as represented by the integral of

$$\frac{\int_{-\pi/W_m}^{\pi/W_m} \frac{A^2(1+\cos W_m t)^2 dt}{4\pi R_a}}{W_m} \text{ is } \frac{3\,A^2}{4\,R_a}.$$

The average single-sideband power component in the above term is $A^2/8R_a$.

Therefore, the relative maximum gain for single-sideband, to amplitude modulated power output level, is $10 \log_{10} 6/1$, or 7.78 decibels.

The battle of the db has just begun.

JOHN P. NICOLOSI
Defense Projects Div.
Western Electric Co.
New York, N. Y.

Author's Comment[2]

The SSB system has been demonstrated to provide performance equivalent to the AM system under typical long-range propagation conditions if the power of the SSB signal is equal to the power in one of the two sidebands of the AM signal. Therefore, when an AM communications system is replaced with an equivalent SSB system, the total power which must be radiated is greatly reduced, with consequent benefit to the radio environment as a whole. The amount of reduction of total radiated power may be found by determining the ratio of the total power of an AM signal, including the carrier and both sidebands, to the power in one sideband alone.

Nicolosi's calculations concern the case in which the AM transmitter is 100 per cent modulated by a sinusoidal signal. Our calculations concern the case in which the AM transmitter is modulated by a voice signal, and we feel that this may be a more meaningful basis of comparison than 100 per cent modulation by a sinusoidal signal. For 30 per cent average modulation index, a value typically achieved when speech processing techniques are not used, the ratio of the total power of the AM signal to the power in one sideband is 46:1 or 16.6 db. For 50 per cent average modulation index, the ratio is 18:1 or 12.5 db. Hence, the values in our text of 12 to 16 db.

J. F. HONEY
Hoffman Labs., Inc.,
Los Angeles, Calif.
D. K. WEAVER, Jr.
Elec. Eng. Dept.
Montana State College
Bozeman, Mont.

[2] Received by the IRE, January 25, 1957.

Letter from Mr. Nicolosi[3]

The comparison level as revealed in the last paragraph of Honey and Weaver's comment is considered very acceptable to this writer.

JOHN P. NICOLOSI

Received by the IRE, February 15, 1957.

A Third Method of Generation and Detection of Single-Sideband Signals*

Letter from Mr. Frank

I should like to comment on the above paper by Weaver.[1]

Mr. Weaver is to be commended for calling attention to a neglected modulation method which was clearly explained by Madella a number of years ago,[2] but apparently not recognized as applicable to single-sideband communication. This lack of recognition is really quite remarkable in view of the extensive treatments which have been given by Tucker and others on polyphase modulation methods.[3] The paper by Tucker and McDiarmid is to be recommended, incidentally, for its excellent treatment and bibliography of the whole field of polyphase modulation, which forms the background for so much of the single-sideband art as well as that of synchronous detection.[4]

I believe a clearer appreciation of the "third method" can be obtained by noting that Weaver describes two rather distinct ideas. The first is the use of modulators with quadrature carriers to generate quadrature-phased audio signals, in place of the wide-band phase shift networks used by the "second method;" basically the location of the first carrier frequency is not critical—it could just as well be at the high end of the audio band.

The location of the first carrier at the center of the audio band is a second distinct idea, which has the great merit of causing the residual sideband to fall on top of the desired sideband; however, for some applications it might be desirable to have the residual sideband and carrier fall on distinct frequencies, so they could be further attenuated by filters.

ROBERT L. FRANK
Sperry Gyroscope Co.
Great Neck, N. Y.

* Received by the IRE, December 26, 1956.

[1] D. K. Weaver, Jr., PROC. IRE, vol. 44, pp. 1703–1705; December, 1956.

[2] G. B. Madella, "Single-phase and polyphase devices using modulation," *Wireless Engr.*, vol. 24, pp. 310–331; October, 1947.

[3] D.G. Tucker and I. E. MacDiarmid, "Polyphase modulation as a solution to certain filtration problems in telecommunications," *Proc. IEE*, vol. 97, part III, pp. 349–358; September, 1950.

[4] J. P. Costas, "Synchronous communications," PROC. IRE, vol. 44, pp. 1713–1718; December, 1956.

Letter from Mr. McPherson[5]

The single-sideband method presented by Weaver[1] is not new, but has been considered in several British publications since 1947.[2,3,6,7] Applications to filtering and to frequency translation are also given.[2,3,6,7] Tucker's Fig. 4,[8] is essentially the same as Weaver's Fig. 7.[9] Tucker's application is to transmission measurement and shows the diversity of utilization of the method, stressing its adaptability to quantitative work.

ROBERT R. MCPHERSON
Ann Arbor, Mich.

[5] Received by the IRE, January 7, 1957.

[6] N. F. Barber, "Narrow band-pass filter using modulation," *Wireless Engr.*, vol. 24, pp. 132–134; May, 1947.

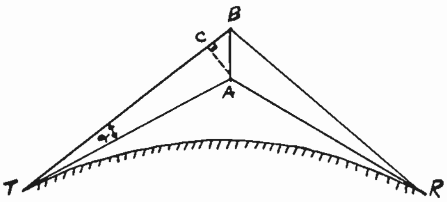
[7] D. G. Tucker, "Highly selective transmission measuring equipment for communication circuits," *J. IEE*, vol. 94, part III, pp. 211–216; May, 1947.

[8] Tucker, *op. cit.*, p. 215.

[9] Weaver, *op. cit.*, p. 1705.

Single-Sideband Techniques in UHF Long-Range Communications*

In the above paper[1] an error appears in (1) on page 1855. The correct analysis is as follows:



$$TA = \frac{}{2}$$

$$AC = \frac{}{2}\sin\alpha$$

$$CB = \frac{l}{2}\sin\alpha\tan\left(\alpha + \frac{l}{2r}\right)$$

$$TC = \frac{l}{2}\cos\alpha$$

$$TB - TA$$

$$= \frac{l}{2}\left(\sin\alpha\tan\left(\alpha + \frac{l}{2r}\right) + \cos\alpha - 1\right)$$

$$= \frac{l}{2}\left(\alpha^2 + \frac{l}{2r}\alpha - \frac{\alpha^2}{2}\right)$$

$$\Delta T = 2\,\frac{(TB - TA)}{C} = \frac{l}{C}\left(\frac{l}{2r}\alpha + \frac{\alpha^2}{2}\right).$$

The authors are to be commended for an excellent presentation of the problems associated with beyond-the-horizon uhf communication.

JAMES E. BARTOW
Signal Corps Eng. Labs.
Fort Monmouth, N.J.

* Received by the IRE, December 26, 1956.

[1] W. E. Morrow, Jr., C. L. Mack, B. E. Nichols, and J. Leonhard, PROC. IRE, vol. 44, pp. 1854–1873; December, 1956.

Author's Comment[2]

The error noted by Mr. Bartow in (1) is correct. The equation should read

$$\Delta T \cong \frac{l}{C}\left(\frac{l}{2r}\alpha + \frac{\alpha^2}{2}\right). \qquad (1)$$

The factor of 2 in the second term results in minor corrections to Fig. 2, which results in slightly smaller multipath delays for the longer circuits.

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Mass. Inst. Tech.
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Received by the IRE, January 7, 1957.

Application of Single-Sideband Technique to Frequency Shift Telegraph*

The above paper by Buff[1] is very interesting and contains much food for thought. There are a few aspects which appear to require clarification, and regarding which I wish to submit certain comment as follows.

I. REDUCTION OF SHIFT AND BANDWIDTH

In discussing the question of 100-cps shift vs 400-cps shift, the statement is made that: "This signal, properly detected at the receivers through narrow-band filters, should render a gain of approximately 6 db over present 400 cycle shift operation."

The gain referred to apparently is in signal/noise ratio.

Analysis, supported by certain comparative tests over commercial facilities, indicates that this gain is not obtained when using the usual sloping-frequency-characteristic type of discriminator such as indicated by the graphs of Fig. 7.

On the contrary, this type of discriminator can be expected to give a 3-db loss in signal/noise ratio for each 2/1 reduction in shift and corresponding bandwidth. The explanation is that while the 2/1 reduction in bandwidth gives a 3-db reduction in noise output—for a given discriminator—the 2/1 reduction in shift gives a 6-db reduction in signal-voltage output; the net result being not a gain but a 3-db loss in signal/noise ratio out of the discriminator.

If, however, we assume a discriminator system consisting of separate MARK and

* Received by the IRE, January 7, 1957.

[1] Christopher Buff, PROC. IRE, vol. 44, pp. 1692–1697; December, 1956.

SPACE band-pass filters, we can expect the gain stated by Buff.

At some value of shift, and corresponding bandwidth, these two methods conceivably should provide equal performance. The location or value of this point, if it does exist as such, should prove to be a fundamental consideration in the design of radio receiving equipment for such services.

II. Wide-Band FM Theory

It seems surprising, and unfortunate, that the quoted statement—"deviation ratio of at least 4.7 was necessary for satisfactory fsk operation"—ever was made. Wide-band or narrow-band, basic fm theory applied to the case of telegraphic keying (fsk) indicates that a deviation ratio of 2.0 should be quite satisfactory, and that the practical minimum probably is 1.0, for hf radio-circuit applications.

To illustrate such analysis, let us take the case of 4-channel, time-division multiplex with values rounded to convenient figures:

Channel speed: 25 dot-cycles/sec or 50 bauds.
Aggregate speed: $4\times 25=100$ cycles/sec or 200 bauds.
Shift: ± 200 cps or total 400 cps.
Deviation ratio: 200/100 or 400/200 $=2.0$.

Assuming sine-wave keying, the usual Bessel function analysis indicates that the bandwidth occupied—by sidebands down to 1 per cent of the unmodulated carrier level—is $2\times 4\times 100=800$ cps. This certainly can be restricted to $2\times 3\times 100=600$ cps, by suitable band-pass filtering at the transmitter, and still provide entirely satisfactory fidelity of envelope shape.

By allowing a nominal 600 cps bandwidth, major components necessary to FS reproduction of the third harmonic, of the fundamental keying frequency, will be passed.

A low-pass filter, for use after the discriminator, can provide suitable attenuation of any 400 cps transient beat between mark and space signals, if these overlap in time due to multipath propagation, and still provide the required fidelity of envelope shape.

It therefore appears that a rigorous and complete fm analysis would indicate that a deviation ratio of 2.0 is ample.

If keying is to be restricted to a sine-wave envelope, with no third harmonic component, a deviation ratio of unity (1.0) and total bandwidth of $2\times 2\times 100=400$ cps would appear to be sufficient. It is believed that, all factors considered, unity deviation ratio is a practical minimum for hf radio-circuit use.

III. Small Shift Values

Study of a tabulation giving sideband amplitudes, for various values of the deviation ratio, seems to indicate that there is little to be gained, and possibly something to be lost in performance, by going to values of deviation ratio appreciably less than unity (1.0). This is for hf radio circuits, especially during periods of marginal signals.

IV. Narrow-Band Advantages

The major and compelling reason for reducing total shift and bandwidth often is not to improve the signal/noise performance (see Section I above) but rather to eliminate interference. In commercial service over hf radio circuits, this may at times even justify a slight sacrifice in signal/noise ratio.

V. Optimum Bandwidth

Eq. (1) in the paper gives

$$bw = \frac{3}{4} \times \frac{1}{P}.$$

This actually is for a low-pass filter, or for half the total width of a band-pass filter.

The statement is made that: "These bandwidths were determined by the criterion for optimum signal/noise ratio; $bw = 3\times$keying speed in cycles."

It would appear that this statement is incomplete, and therefore possibly misleading, even for the case of ON/OFF keying to which it applies. For the fsk case, it may sometimes be a fair approximation or a convenient rule-of-thumb. For general fsk use, though, it hardly can be recommended as a proper, or a satisfactory, analytical and design criterion.

VI. Narrow-Band SSB/FSK Tests

In giving values of distortion caused by multi-path propagation, the text does not clearly state whether the figures apply to the case of a 60-wpm channel or to 4-channel multiplex keying at 150 baud rate.

Since one purpose of this SSB issue of PROCEEDINGS is to provide information for the use and guidance of regulatory bodies such as the FCC—and possibly the CCIR—it seems important that such information be as accurate, and as complete, as possible.

John B. Moore
New York, N.Y.

Author's Comment[2]

In reply to Mr. Moore's comments which are highly valued, I wish to submit the following:

1) It is agreed that a *given* slope-discriminator will exhibit a 6-db loss in signal-voltage output for each 2:1 reduction in frequency shift.

The thought in mentioning the possibility of a 6-db gain, however, was based on the premise that the discriminator for the narrow shift would be proportionately more sensitive.

The idea would be, through the use of higher Q elements and closer-to-center tuning to trade wide-band linearity for narrow-band sensitivity, such that the voltage output for ± 50 cycles would be equal to that formerly obtained with ± 200 cycles shift, for example.

The discriminator shown in Fig. 7 is one used in line-channeling equipment and does not represent the best which we think might be done in this direction.

[2] Received by the IRE, January 17, 1957.

With this discriminator, we were able to break even on $s{:}n$ with a ± 50 cycles shift and ± 125 cycle input band-pass filter as compared to ± 200 cycles shift using MARK-SPACE filters, each ± 125 cycles. The keying wave distortion was approximately the same in either case, so that while no actual gain was obtained, in this case, a bandwidth reduction was achieved.

If we attempted to use MARK-SPACE filter discrimination with ± 50 cycles shift, the filters would have to be so sharp as to greatly increase keying distortion when compared to slope-discrimination at the same keying speed. The slope-discriminator allows, we find, approximately twice the keying speed, with equivalent distortion and input bandwidths.

2) The statement: "Davey and Matte have shown that satisfactory separation of mark and space signals in the fsk system is possible for a deviation ratio as low as 4.7," was made by Jordan, *et al.*[3]

We agree that a unity deviation ratio is a good practical minimum for hf. It will probably be still some time before the majority of fsk transmissions achieve this reduction.

3) It is agreed that interference reduction may be a compelling reason for bandwidth reduction, even at the expense of a small $s{:}n$ loss.

4) Through the use of the ARQ system, to which you made significant contributions, such small losses, if they did occur, could largely be offset.

5) The formula stated is for the asymmetrical, low-pass filter, case. The band-pass filter, being DSB, would be twice this value. From this, the factors 1.5 and 3.0 times keying speed for the low-pass and band-pass filter bandwidths, respectively, are derived.

This appears to be a good criterion for fsk, as well as ook. In fsk where we have rigid amplitude limiting before discrimination it does not appear necessary to provide additional bandwidth for keying-wave-squaring as was necessary for ook.

As you can see, the unity deviation ratio which you mentioned as a practical minimum fits very nicely into this formula; for example:

ARQ Mux speed $=90$ dot-cycles/sec.
BW (band-pass filter): $3\times 90=270$ cps.
BW (low-pass filter): $1.5\times 90=135$ cps.
Frequency shift: $2\times 90=180$ cps.
Deviation ratio: 1.0.

6) Distortion on a 4-channel mux (150 bauds, non-ARQ) signal ran between 10 per cent and 20 per cent during the test described. This was during the normally useful periods of the frequency involved.

We appreciate your bringing to our attention your thoughts on these matters.

Christopher Buff
Brentwood, N.Y.

[3] D. B. Jordan, H. Greenberg, E. E. Eldredge, and W. Serniuk, "Multiple frequency shift teletype systems," Proc. IRE, vol. 43, pp. 1647–1655; November, 1955. See p. 1652, first column.

Discreet Writing*

A small cheer for Kulinyi, Levine, and Meyer,[1] who, on pages 1819 and 1821 of the SSB issue, spelled the word "discrete" correctly six times in two consecutive paragraphs. A record in this day. Hip, hip, and hats off.

WILLIAM L. SMITH
Spencerville, Md.

* Received by the IRE, January 14, 1957.

[1] R. A. Kulinyi, R. H. Levine, and H. F. Meyer, "The application of SSB to high-frequency military tactical vehicular radio sets," PROC. IRE, vol. 44, pp. 1810–1823; December, 1956.

SSB Performance as a Function of Carrier Strength*

Some important statements were made in the above paper by Firestone[1] that appear not to be substantiated by either mathematical analysis or measured data.

The statements appear under the headings "Receiver Desensitization" and "Intermodulation." The first statement in question under "Receiver Desensitization" reads as follows: "The strongest unwanted signal which needs to be rejected by a receiver depends greatly on the desensitization and intermodulation characteristics of the receiver." First, let us consider the desensitization question. Desensitization is only one of several types of interference which may effect the desired-signal performance of a receiver, thus, it should not be singled out as an offender until due consideration is given to other types of interference by some systematic system interference analysis. The several types of interference that may effect the desired-signal performance of a receiver operating in a system, and being interfered with by a single undesired transmitter, may be divided into the two following categories:

Category I—Interference caused by limited performance of the receiver.
1) Inadequate selectivity,
2) Signal desensitization,
3) Spurious and image responses.

Category II—Interference caused by extra band radiation of the interfering transmitter.
1) Modulation splatter,
2) Broad band noise radiation (transmitter noise),
3) Spurious and harmonic radiations.

Three types of interference are listed in each category. Observation and experience indicate that for passable design of transmitters and receivers, corresponding numbered types in each category will produce similar effects in system performance. Only the most careful interference evaluation can determine whether the transmitter or receiver is at fault.

One such method of interference evaluation relies on the use of high Q cavity filters, of known characteristics, which are capable of making the performance of either the receiver or the interfering transmitter more nearly perfect. A typical test setup for the measurement of receiver desensitization is shown in Fig. 1.

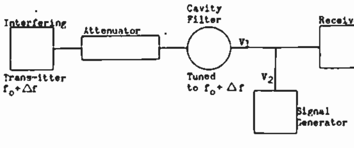

Fig. 1.

The cavity filter tuned to the frequency $f_0+\Delta f$ must have sufficient selectivity at the frequency f_0 to reduce any extra-band radiation from the interfering transmitter, near f_0, so that only the carrier frequency, $f_0+\Delta f$, causes desensitization. With this setup the receiver's signal desensitization figure may be determined as follows:

1) With the interfering transmitter turned off, adjust the signal generator on frequency f_0 to produce a standard s/n (usually 12 db). Record V_2 required to establish the standard s/n.

2) Turn on the interfering transmitter and adjust the attenuator to reduce the standard s/n by a given amount (usually 6 db). Record V_1, the level of the interfering signal at the input to the receiver.

3) The signal desensitization figure of the receiver for that particular separation, Δf, can then be calculated in db by the relationship:

$$20 \log_{10} \frac{V_1}{V_2}.$$

A typical test setup for the measurement of transmitter noise figure is shown in Fig. 2.

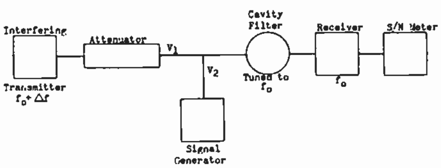

Fig. 2.

The essential difference between Fig. 1 and Fig. 2 is the location and adjustment of the cavity filter. The cavity filter when tuned to f_0 should have sufficient attenuation near frequency $f_0+\Delta f$ to reduce the level of signal desensitization so that only transmitter noise figure is being measured. The transmitter noise figure is measured by using the same steps as listed above for signal desensitization figure.

The key to accurate interference evaluation in both of these measurements is the cavity which separates the two types of interference so unquestionable guilt may be established. If the receiver signal desensitization figure in db is larger than the transmitter noise figure in db, then the guilt is most certainly with the transmitter. The degree of guilt with the transmitter or the receiver may vary over a wide range for different types of equipments or state of the art, but in no case can one be considered without proper evaluation of the other.

Modern design receivers which would be employed in the type of service under discussion, vhf band, provide a larger signal desensitization figure than the transmitter noise figure of transmitters for the same type of service for frequencies spacing greater than 1 per cent. Frequency spacings greater than 1 per cent normally produce types of interference not pertinent to this discussion. Typical signal desensitization figures for receivers employed in the vehicular communications service are from 95 db to 105 db for close spaced channels and up to 130 db for channels at 1 per cent separation. Typical transmitter noise figures for base station transmitters, in the same service, are from 75 db to 95 db for close spaced channels and up to 130 db for channels at 1 per cent separation. Based on these data, transmitter noise figures are up to 20 db less than receiver signal desensitization figures for close spaced channels, thus the strongest unwanted signal that needs to be rejected depends greatly on the transmitter noise figure and not receiver desensitization.

If the interfering signal is generated by a SSB transmitter, the author states that the amount of desensitization is determined by the average power radiated. This analysis may be true, however, the author again neglected interference due to transmitter noise figure. Recent measurements reported by J. S. Smith[2] revealed that the transmitter noise figure of a SSB transmitter operated near 30 mc was from 10 db to 15 db poorer than for a comparable fm transmitter. The transmitter noise figure of a SSB transmitter is relatively constant with changing modulation level, provided the transmitter remains keyed for the duration of each transmission. For systems employing lock-in oscillators it is normal that the transmitter remains keyed for the duration of each transmission. For SSB transmitters operating on frequencies above 30 mc it can be expected that transmitter noise figures may cause increased interference due to increased converter noise on higher frequencies.

The second problem in question that the author discusses is intermodulation. The statement is made that $f\pm\Delta f$ and $f\pm\Delta 2f$ type of intermodulation produces a signal proportional to $3A^2B$. This signal amplitude, listed at a previous point in the article as

$$\frac{3A^2Ba_3}{4}$$

is the largest of a number of third order intermodulation terms that are produced in a receiver. It is further stated that if A and B are each reduced by 16 db then the total decrease in the intermodulation term would be 48 db. This total reduction in the intermodulation term, if true, cannot be realized in a reduced carrier system if the reduced carrier is used for a locked-in oscillator. A SSB-receiver employing a locked-in oscillator operating from a reduced carrier must provide increased selectivity for the locked-in oscil-

* Received by the IRE, February 14, 1957.

[1] W. L. Firestone, PROC IRE vol. 44, pp 1839–1848; December, 1956.

[2] J. S. Smith, "Adjacent Channels and the Fourier Curse," PGVC Annual Natl. Conf.; November 29–30, 1956.

lator channel to obtain equal or greater s/n than in the voice channel. The intermodulation frequency would normally land in the narrower locked-in oscillator channel and degrade its operation. Normally, the s/n required for a locked-in oscillator channel is greater than the minimum usable s/n in the audio channel thus improvement in intermodulation for the example given is less than 48 db − 16 db, or 32 db.

In arriving at the conclusion that the intermodulation term is proportional to $3A^2B$, the author must have assumed that the multiplier a_3 included in a previous point in the article, is a constant.[3,4] Assuming a_3 is a constant disagrees with the first part of the third-order term as derived by Terman.[5] The complete expression for the first part of the third-order term pertinent to this discussion is as follows:

$$\frac{u}{3!g_m}\frac{\partial^2 g_m}{\partial E_g{}^2}\frac{3A^2B}{4}.$$

The second partial derivative

$$\frac{\partial^2 g_m}{\partial E_g{}^2}=\frac{24I}{E^3}.$$

If the plate load impedance is negligibly small, where I is the crest value of the third-harmonic component of the plate current, E_g is the grid bias at the operating point and E is the crest value of the signal voltage applied to the grid. If substitution is made for the second partial derivative, (1) becomes:

$$\frac{18uIA^2B}{3!g_mE^3}. \qquad (2)$$

Since E^3 is proportional to A^2B and u and g_m are assumed constants, the third-order intermodulation product is proportional to I and not A^2B.

Measurements made on amplifiers and converters operating at 150 mc seem to agree with (2). If A^2B is held constant then the intermodulation product is constant over a limited range, however, if A^2B varies then the intermodulation product varies directly with some second factor believed to be the crest value of the third harmonic component of the plate current. Typical measurements made with signals producing intermodulation products equivalent to about 1 μv show a decrease of about 10 db in the intermodulation product for a decrease of 6 db for both A and B. On the basis of these data the maximum reduction in intermodulation products that can be expected for the example given by the author is about 27 db or a net reduction of less than 11 db for lock-in oscillator operation. As mentioned above the lock-in oscillator channel normally requires an improved s/n over the minimum s/n for the voice channel, thus the improvement, if any, that could be expected in a SSB system with reduced carrier depends on design requirements for the lock-in oscillator channel.

Many of the conclusions listed at the end of the article are based on calculations that neglect the

$$\frac{\partial^2 g_m}{\partial E_g{}^2}$$

term in the odd-order intermodulation products; thus reevaluation will be necessary before creditability could be established.

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[3] This same assumed constant appears in: W. L. Firestone, "Evaluation of sideband noise and modulation splatter," 1955 IRE CONVENTION RECORD, part 8, pp. 22–28.

[4] This same assumed constant appears in: W. L. Firestone, A. Macdonald, and H. Magnuski, "Modulation splatter of vhf transmitters," *Proceedings of the Natl. Electronics Conf.*, vol. 4, pp. 264–273; February, 1955.

[5] F. E. Terman, "Radio Engineers Handbook," McGraw-Hill Book Co., Inc., New York, N. Y., 1st ed., p. 464.

Author's Comment[6]

In response to Mr. Shepherd's letter regarding my paper, I would like to make the following comments.

I believe that the statement which I made and which reads as follows: "The strongest unwanted signal which needs to be rejected by a receiver depends greatly on the desensitization and intermodulation characteristics of the receiver," is valid. To begin with, Mr. Shepherd points out that there are many other types of interference which should be considered. It is therefore clear the above statement which I made is being taken "out of context." It is only necessary to read the title of my paper to realize that I am varying only one parameter in my entire paper, and it is my purpose to evaluate the most important system characteristics which are affected by this change. For a given transmitter and receiver setup, changing the amount of carrier DOES NOT affect the selectivity of the receiver nor its spurious and image responses, and hence need not be considered as germane to the general discussion. Furthermore, changing the amount of carrier radiated should not significantly change the spurious and harmonic radiations or the broad-band noise radiation. This is relatively easy to see since for a given transmitter neither the type or number of stages, the plate dissipation, selectivity, nor the mode of operation change greatly as the carrier injection level is varied, hence these factors are not discussed.

While Shepherd's discussion of suitable methods of measuring desensitization and transmitter noise figure is very interesting, it seems to fit in more appropriately with some previous works which I have published[7,8] than the present SSB article. And while I can agree that typical transmitter noise figures are from 75 to 95 db for close spaced channels, I cannot agree that typical signal desensitization figures for receivers employed in the vehicular communication service are from 95 to 105 db for close spaced channels. It is my experience that typical figures are between 70 and 85 db for close spaced (40 kc) channels, and these figures are based on typical equipment manufactured by several of the leading companies in the field. If the spacing were decreased to say 10 kc as would be the case for SSB then these figures would be significantly worse. The conclusion is that transmitter noise figures are generally as good or better than the receiver signal desensitization.

In regard to my intermodulation comments I would like to point out that in two short paragraphs it is not possible to treat such an involved subject in much detail. It was therefore my purpose to point out the general direction of the improvement and give some indication of the amount of improvement that could be expected. It appears to me that Shepherd's comments boil down to two major points, namely that I assumed the coefficient a_3 was a constant, and that the improvement indicated cannot be realized in a lock-in oscillator system. I shall therefore endeavor to answer these points separately.

It is true that I have tacitly assumed that a_3 is a constant for the purposes of this discussion and have done so for several reasons. First, and most important is that a_3 may be considered to be constant for very small signal operation since the total signal swing is then over a very small portion of the transfer characteristic. Experience shows that in this small signal region the intermodulation is approximately proportional to A^2B. For example, a receiver which has 60 db of intermodulation protection and produces an on-channel interfering signal of one microvolt would require only 1000 μv on a 0.001 volt of adjacent and alternate channel signal at the first grid. Secondly if a_3 is not a constant, the extent to which it is not a constant depends upon: 1) the specific tube used, 2) the point of dc operation (all dc voltages), 3) the level of the incoming signal, and 4) the past life history of the tube.

Hence, even though in general a_3 which is proportional to

$$\frac{\partial^2 g_m}{\partial E g^2}$$

is not exactly a constant, I felt that a general discussion should not make an assumption on a factor which can vary so greatly from set to set and time to time.

And while I agree that the complete expression for the first part of the third order term is as follows

$$\frac{\mu}{3!g_m}\frac{\partial^2 g_m}{\partial E_g{}^2}\frac{3A^2B}{4};$$

and while the second order partial derivative may be expressed as

$$\frac{\partial^2 g_m}{\partial E_g{}^2}=\frac{24I_3}{E^3},$$

where I_3 is the crest value of the third harmonic component of the plate current, and E is the crest value of the signal voltage applied to the grid; the application of this expression is only valid when a single sine-wave voltage is applied. This expression is taken from Terman,[5] and I quote him: "These partial derivatives can be determined experimentally by applying to the grid of the tube a sine-wave voltage of known amplitude when the load impedance in the plate circuit is negligibly small." To extend this second partial derivative evaluation into the domain of 2 or more sine wave voltages which vary in amplitude is not valid without additional analysis. To illustrate this point consider I_3. I_3 is defined as

[6] Received by the IRE, February 28, 1957.

[7] W. L. Firestone, "Evaluation of sideband noise and modulation splatter," 1955 IRE CONVENTION RECORD, part 8, pp. 22–28.

[8] W. L. Firestone, A. Macdonald, and H. Magnuski, "Modulation splatter of vhf transmitters," *Proc. of the Natl. Electronics Conf.*, vol. 4, pp. 264–273; February, 1955.

the crest value of the third harmonic component of plate current. If two sine-waves are applied (adjacent and alternate channels as required to generate intermodulation) then what is I_3? Is it the crest value of the third harmonic component of plate current of the adjacent channel, or the alternate channel, or is it the crest value of the combination of these signals even though they are at different frequencies? Similarly E was based on a single sine wave signal, how do we now interpret it for 2 signals? What value of E would we apply to determine the 2nd partial if we are then going to use two signals E_1 and E_2 for intermodulation determination?

In any case evaluating the second order partial derivative by Terman's method yields a number for that partial derivative and for that level of operation and not a variable which is an inverse function of A^2B as implied.

If all of Shepherd's comments are correct then the intermodulation product varies as the crest value of the third harmonic component and not with the signal strengths of the intermodulation signals. This conclusion is contrary to existing information.[7–9]

Finally if Fig. 2 in my SSB paper is carefully examined it can be seen that the third order distortion is more accurately expressed as

$$\left(\frac{3b^2c}{4}\right)a_3 + \left(\frac{5b^4c}{4} + \frac{15b^2c^2}{8}\right)a_5$$
$$+ \left(\frac{105b^6c}{64} + \frac{105b^2c^5}{32} + \frac{105b^4c^3}{16}\right)a_1$$
$$+ \left(\frac{64b^8c}{32} + \frac{315b^2c^7}{64} + \frac{945b^6c^3}{64}\right.$$
$$\left. + \frac{315b^4c^5}{16}\right)a_9 + \text{higher order terms.}$$

Consequently to be completely accurate one should not evenconfine oneself to the a_3 term. It is my experience that in practice the a_5 to a_n terms can be very important in determining the magnitude of the third order term and in some special cases they might be more important than the a_3 term.

As a guide in intermodulation, assuming that a_3 is constant and that higher order terms are not too important is, in my opinion, a reasonable approach unless one wishes to investigate individual cases and go to great lengths to obtain very precise results in these cases.

As regards the statement that the lock-in oscillator system cannot take full advantage of the improvement indicated, Shepherd may be correct, although in any case significant advantage in improvement will result. However two factors stand out. First, if one looks at the very last page of my paper, it is clear that 6 different types of receiving systems for SSB were mentioned: 1) afc, 2) lock-in oscillator, 3) phase lock system, 4) active filter system, 5) narrow band amplifier and limiter system, and 6) locally injected oscillator when system stability justifies it. It is easy to enlarge this list, for example, by adding a standard AM detector, etc. To discuss in detail how much advantage each system could take of the indicated intermodulation improvement when reducing was beyond the intent of the paper, and why the lock-in oscillator system was singled out is not clear to me.

Referring to the statement made by Shepherd, "Normally, the s/n required for a locked-in oscillator channel is greater than the minimum usable s/n in the audio channel, thus improvement in intermodulation for the example given is less than 48 db − 16 db, or 32 db," I cannot justify why he subtracted 16 db from the improvement. Starting with a full carrier system and then reducing the carrier by 16 db would necessitate narrowing up the lock-in oscillator channel sufficiently such that its output resultant s/n is better than that in the audio channel. However, good design dictates this in the first place and now that the system is properly designed to receive a reduced carrier on channel signal, if we then reduce the intermodulating signals by 16 db apiece the improvement expected is up to 48 db. This reducing of the desired channel by 16 db does not worsen the intermodulation by 16 db as assumed above.

Finally, I cannot even agree with the statement that the s/n in the lock-in oscillator channel needs to be better than the s/n in the voice channel. If the lock-in oscillator channel includes a limiter as the last stage, all AM is stripped off of the lock-in oscillator signal. Experience with such a circuit shows that the s/n in the lock-in oscillator grid circuit may even be worse than the s/n in the voice channel, depending on the selectivity of lock-in oscillator filter, and still not degrade the communication in any way.

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[9] J. F. Byrne, "The selectivity and intermodulation problem in uhf communication equipment," Natl. Conf. on Airborne Electronics, Dayton, Ohio; May 12, 1952. See Fig. 10.

The Design of Wide-Band Phase Splitting Networks*

In the above paper[1] on wide-band phase-shift networks for single-sideband systems, I gave explicit formulas for the maximally flat approximation and the Tchebycheff approximation by means of elliptic functions. In the same year Darlington[2] and Orchard[3] published independently two very similar papers. Neither of these two papers treated the maximally flat case, whereas formulas for the maximally flat approximation were contained in my paper. The form of the mathematical relations in the three papers was so different that *Wireless Engineer* published a correspondence[4] in which I proved the equivalence of the results for Tchebycheff approximations in the three papers.

These factors, namely, the independence of the papers, the singular treatment of the maximally flat case in my paper, and the proof of equivalence in the later correspondence were indicated in an integrating article by Winkler.[5]

Since reference to my paper has been omitted in later studies of this field published in PROCEEDINGS,[6,7] including the recent SSB issue,[8] I feel that readers of the SSB issue should be informed of my article, as well as those of Darlington and Orchard.

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* Received by the IRE, February 15, 1957.

[1] W. Saraga, PROC. IRE, vol. 38, pp. 754–770; July, 1950.

[2] S. Darlington, "Realization of a constant phase difference," *Bell Sys. Tech. J.*, vol. 24, pp. 94–104; January, 1950.

[3] H. J. Orchard, "Synthesis of wide-band two-phase networks," *Wireless Eng.*, vol. 27, pp. 72–81; March, 1950.

[4] W. Saraga, "Wide band two-phase networks," *Wireless Eng.*, vol. 28, pp. 30–31; January, 1951.

[5] Stanley Winkler, "The approximation problem of network synthesis," IRE TRANS., vol. CT-1, pp. 5–20; September, 1954.

[6] "Radio progress during 1950," PROC. IRE, vol. 39, pp. 359–396; April, 1951. See p. 366.

[7] O. G. Villard, Jr., "Cascade connection of 90-degree phase-shift networks," PROC. IRE, vol. 40, pp. 334–337; March, 1952.

[8] D. E. Norgaard, "The phase-shift method of single-sideband signal generation," PROC. IRE, vol. 44, pp. 1718–1735; December, 1956. Also, "The phase-shift method of single-sideband reception," pp. 1735–1743.

Correspondence

A Junction Transistor for Kilowatt Pulses*

While it has long been recognized that the transistor is a very useful device for the fast switching of currents, applications have usually been at relatively low current levels. It is the purpose of this note to describe a transistor which has been developed to switch currents of 40 amperes in times of the order of a microsecond. Since the transistor can operate on voltages up to 30 volts, pulses with powers in the kilowatt range can be produced.

The theory of large signal switching behavior of transistors has been developed by Ebers, Moll, and Miller.[1-3] For our purposes the design requirements may be summarized: 1) high α at the operating current; 2) low extrinsic base resistance, and 3) high α cutoff frequency.

These objectives have been achieved using the design theory proposed by the present author,[4,5] in which the emitter is in the form of a thin bar, flanked by parallel bars making ohmic base connection. In the present transistor the configuration has been distorted to annular shape, since for this particular size, this allows a more economical use of germanium. The transistor element is shown in Fig. 1.

To increase emitter efficiency an alloy of 0.5 per cent gallium in indium was used for the emitter.[6] Traditional alloying methods were used, but a special effort was made to achieve a uniform thin base layer over the large area involved. This was done by reducing the thickness of the germanium and indium to about 0.005 inch and using very carefully machined graphite alloying jigs.

* Received by the IRE, November 13, 1956.

[1] J. J. Ebers and J. L. Moll, "Large-signal behavior of junction transistors," Proc. IRE, vol. 42, pp. 1761–1772; December, 1954.

[2] J. L. Moll, "Large signal transient response of junction transistors," Proc. IRE, vol. 42, pp. 1773–1784; December, 1954.

[3] J. J. Ebers and S. L. Miller, "Design of alloyed junction germanium transistors for high speed switching," *Bell Sys. Tech. J.*, vol. 34, pp. 761–781; July, 1955.

[4] N. H. Fletcher, "Some aspects of the design of power transistors," Proc. IRE, vol. 43, pp. 551–559; May, 1955.

[5] N. H. Fletcher, "Self-bias cutoff effect in power transistors," Proc. IRE, vol. 43, p. 1669; November, 1955.

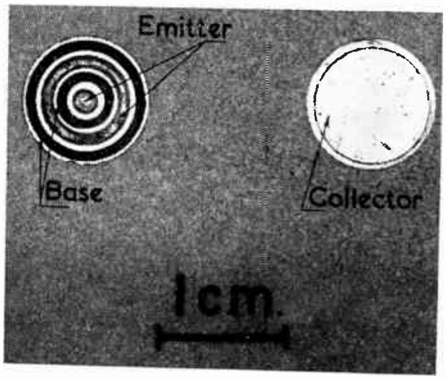

Fig. 1—The transistor element. Base rings show up black, emitters grey. The collector covers most of the back surface.

Relatively few transistors have been made, but the results reported below indicate what can reasonably be attained with this design. Collector currents as high as 45 a have been obtained with as little as 3-a base current, though average transistors require rather more drive than this. Rise times are fastest[2] when a constant current pulse is applied to the emitter in a grounded base configuration.

Rise time for a collector current of 40 a is as little as $\frac{1}{2}$ microsecond for some transistors, even without collector "bottoming," and overdrive could reduce this somewhat without appreciable storage. Rise times for other pulsing arrangements are much longer[2] (several microseconds typically) but greater power gains can be realized and overdrive is more economical. Extrinsic base resistances are less than 1 ohm so that input impedances are very low and power gain is quite high.

[6] L. D. Armstrong, C. L. Carlson, and M. Bentivegna, "PNP transistors using high-emitter-efficiency alloy materials," *RCA Rev.*, vol. 17, pp. 37–45; March, 1956.

The mechanical structure of the transistor is shown in Fig. 2. The collector junction has a low thermal resistance path to the chassis and the unit can dissipate considerable power, though it was designed principally for low duty cycle pulsing systems where dissipations of only a few watts are involved.

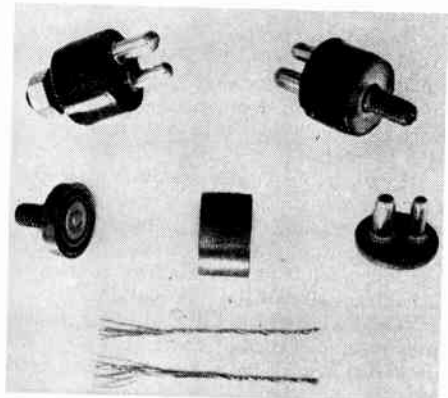

Fig. 2—An exploded view of the transistor, with two completed units above. Araldite type *D* is used to fill the case.

The author is specially indebted to H. Flood of this Laboratory who made the alloying jigs and other fittings as well as the transistors themselves.

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For a photograph and biography of Howard Boyet, see page 558 of the April, 1956 issue of Proceedings.

❖

Bobby J. Duncan (M'54) was born in Carrollton, Ga., on February 1, 1930. He received the B.S. degree in physics from Berry College, Rome, Ga., in 1950 and the M.S. degree in physics from Emory University, Atlanta, Ga. in 1951. While at Emory he performed research in microwave spectroscopy.

He continued research in this same field while performing graduate work toward a doctorate in physics at the University of Florida, Gainesville, Fla. until December 1952, at which time he joined the Sperry Gyroscope Company as a project engineer.

B. J. Duncan

At Sperry he initially performed research in applied microwave spectroscopy. Subsequently, he was associated with research and development projects on microwave ferrites, special radar systems, radar countermeasures techniques, and obstacle avoidance equipment. More recently he has worked almost exclusively in the field of microwave ferrite research and components applications. At present he is a senior engineer and group leader of the microwave ferrite research and advanced development group in the Applied Physics Section of the Microwave Electronics Division of Sperry.

❖

Lawrence Gould was born in Boston, Mass., on November 28, 1930. He received the B.S. and Ph.D. degrees in physics from

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L. Gould

In 1954 he was drafted into the Signal Corps and assigned to the Evans Signal Laboratory to study problems of microwave gas discharge devices. After a two year service period, he returned to Microwave Associates Inc. and is now directing a group working on microwave switching problems using ferrite and gas discharge devices.

He is a member of the American Physical Society and Sigma Xi.

❖

Alexander J. Grossman (M'45) was born on September 1, 1904, in New Rochelle, N. Y. He received the degree of Electrical Engineer from Rensselaer Polytechnic Institute in 1925. Since that time he has been a member of the technical staff of Bell Telephone Laboratories, engaged in the development of transmission networks. For several years, his main activity was the application of contributions in the field of network theory to practical problems and the development of the associated design techniques. At present, his interests include the development of networks for submarine cable systems, exploratory work in the field of transistor networks, and network theory. He is the author of a paper on electric wave filters published in the Pender-McIlwain Handbook.

A. J. Grossman

Mr. Grossman is a member of the Administrative Committee of the Professional Group on Circuit Theory.

❖

Nathan I. Hall (SM'47–F'55) was born in Elkins, W. Va., on October 24, 1910. He holds the degrees of B.S.E.E., M.S.E.E., E.E., and D.Sc.

N. I. Hall

Dr. Hall was a member of the West Virginia University faculty from 1934 to 1936. During this period, he did research work on high-speed electric motors and radio transmission. His investigations included ionospheric measurements and measurement of the velocity of radio ground waves. He has published numerous technical papers on these and other subjects.

During 1936–1937, Dr. Hall was a research assistant to Dr. F. E. Terman at Stanford University. At Stanford, he developed ionospheric measuring equipment and was the first to make ionospheric measurements west of the Mississippi River.

In 1937, Dr. Hall joined the technical staff of Bell Telephone Laboratories, where he spent the following ten years in the research and development of electronic telephone switching systems and radar systems. He has been granted a long list of patents in these and other fields.

Since 1947, Dr. Hall has been with the Hughes Aircraft Company in Los Angeles. He is a vice-president of the company and directs Hughes' Systems Development Laboratories. He is responsible for the development of guided missiles and other electronic weapon systems for the Department of Defense.

Dr. Hall is the recipient of Eta Kappa Nu's 1943 award as the nation's most outstanding young electrical engineer and is listed in *Who's Who in America*. He is a member of the Institute of Aeronautical Sciences, the American Physical Society, The Institute of Electrical Engineers, Eta Kappa Nu, Sigma Xi, and Tau Beta Pi.

❖

Joseph Hannwacker was born on March 8, 1928 in Brooklyn, N. Y. He received the B.S.E.E. degree in 1954 from Brooklyn Polytechnic Institute, Brooklyn, N. Y. He has performed graduate work in physics at Columbia University in New York.

J. Hannwacker

From January, 1955, to September, 1956, he was associated with the Sperry Gyroscope Company working in the field of applied microwave spectroscopy. He also worked extensively in the field of microwave ferrite research and component development. He is presently employed by the Polytechnic Research and Development Company.

❖

Arthur Karp (S'47–A'48–M'53–SM'56) was born in New York, N. Y. in 1928 and received the B.E.E. degree from the College of the City of New York in 1948 and the S.M. degree from the Massachusetts Institute of Technology in 1950. In 1950–51 he held a scholarship from the French Government and worked in the microwave vacuum tube research department at the Laboratoire Central de Télécommunications in Paris.

A. Karp

Mr. Karp entered the field of microwave vacuum tube research at the M.I.T. Research Laboratory of Electronics in September, 1948, studying traveling-wave tubes and the noise in electron beams. He joined the Electronics Research Department of the Bell Telephone Laboratories in October, 1951, engaging in research on electron tubes for millimeter wavelengths. Since October, 1956 he has been on leave-of-absence and is a member of Trinity College at the University of Cambridge, engaged in a doctoral research program at the Engineering Laboratory under the sponsorship of the British Admiralty.

Mr. Karp is a member of Tau Beta Pi and Eta Kappa Nu, and is an associate member of Sigma Xi.

❖

Bernard R. Linden was born in Newark, N. J., on February 6, 1926. He received the B.A. degree in physics from Cornell University in 1947, and the M.Sc. and Ph.D. degrees in physics from Ohio State University in 1949 and 1951, respectively. From 1951 to 1952 he was on a Fulbright Fellowship at the University of Amsterdam. From 1952 to the present he has been associated with the Tube Research Laboratories of the Allen B. Du Mont Laboratories, Inc. His main work has been in the field of physical electronics as applied to solid-state physics and electron-optical design.

B. R. Linden

Dr. Linden is a member of the American Physical Society. He is also a representative on the JETEC-4 Committee (photosensitive devices) of the Radio and Television Manufacturers Association.

❖

Kenneth E. Mortenson (S'46–A'50–M'55) was born in Melrose, Mass. on December 14, 1926. From 1944 to 1945 he attended Wesleyan University in the Navy V-12a program and was later transferred to the NROTC program at Rensselaer Polytechnic Institute, where he completed the requirements for the B.S. and B.E.E. degrees in 1947 and 1948, respectively.

K. E. Mortenson

Dr. Mortenson taught physics at R.P.I. in 1947–1948 and later became an instructor in electrical engineering while obtaining the M.E.E. degree in 1950. From 1949 to 1952 he participated in research projects, sponsored by the Signal Corps, which dealt with radiation and leakage and electromagnetic coupling devices. In 1952 he became a research associate, and in 1953 an assistant professor of electrical engineering at R.P.I. while completing the requirements for the Ph.D. degree in applied physics, which he received in 1954. From 1952 to 1956 he directed several research projects, including work on broadband interference measurements and transistor circuitry.

In 1956 Dr. Mortenson joined the General Electric Research Laboratory in Schenectady, N. Y., where he has been engaged in studies of the physical operation of transistors, particularly as an electric circuit component.

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❖

Gerald Schaffner (S'48–A'50–M'55) was born on May 14, 1927 in Chicago, Ill. He received the B.S. degree in electrical engineering and the M.S. degree in electrical engineering from Purdue University in 1949 and 1950, respectively. In 1956 he received the Ph.D. degree in electrical engineering from Northwestern University.

G. Schaffner

Dr. Schaffner was employed as a designer with the Thordarson Electric Manufacturing Company at Chicago, Ill. from 1950 to 1951. Since 1951 he has been with the Electronics Division of the Stewart-Warner Corporation in Chicago. At Stewart-Warner Dr. Schaffner is a research engineer working primarily with low power uhf beacons.

Dr. Schaffner is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

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For a photograph and biography of Harold Seidel, see page 1479 of the October, 1956 issue of Proceedings.

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Philip A. Snell (A'39–VA'39–M'48) was born on September 15, 1900, in Gloversville, N. Y. He graduated from Technical School in Chicago and continued his education, enrolling for special courses at Purdue Extension University in Fort Wayne, Ind.

He served as tube technician with Westinghouse Lamp Company from 1928 to 1935. He joined the research tube staff of Farnsworth Television in Philadelphia from 1935 to 1939. He then transferred to Fort Wayne to join the research staff of the new Farnsworth, Capehart Corporation from 1939 to 1948. Since 1948 he has been associated with the Allen B. Du Mont Laboratories, Inc. His work has been in the development of special type cathode-ray tubes and phototube devices.

P. A. Snell

Mr. Snell is a member of the American Museum of Natural History.

❖

Leonard Swern (A'49–M'53) was born on February 12, 1925 in New York, N. Y. He received the A.B. degree in physics from Columbia College in 1945. From 1947 to 1949, while a graduate student at Columbia, he worked at the Columbia Radiation Laboratory doing research and development work on millimeter wavelength techniques and components. He received the M.A. degree in physics from Columbia in 1948 and completed the course requirements for the Ph.D. degree in physics. He has also taken courses in applied mathematics at New York University.

L. Swern

Mr. Swern joined the Sperry Gyroscope Company in 1949 as a project engineer working on the design of microwave bolometers. In 1951 he was assigned to the microwave research group where he worked on projects involving the interactions between microwaves and matter. He became a senior engineer in 1954 and is now Engineering Section Head for Applied Physics in the Microwave Electronics Division of Sperry, where he is in charge of the microwave research and advanced development work on the properties of ferrites and other anisotropic media at microwave frequencies. He is a member of the American Physical Society.

Kiyo Tomiyasu (S'41–A'42–M'49–SM'52) was born in Las Vegas, Nev., on September 25, 1919. He received the B.S. degree in electrical engineering from the California Institute of Technology in 1940, and the M.S. degree in communication engineering from Columbia University in 1941. With a Low Scholarship, he studied at Stanford University and then entered Harvard University to continue graduate work on a Gordon McKay Scholarship. He served as a teaching fellow and research assistant at Harvard, and, after receiving the Ph.D. degree, he served as instructor.

Kiyo Tomiyasu

In September, 1949, Dr. Tomiyasu joined the Sperry Gyroscope Co. as project engineer and in 1952 was promoted to the position of engineering section head for microwave research in the Microwave Components Department.

Since August, 1955, he has been a member of the Technical Staff at the General Electric Microwave Laboratory, Palo Alto, Calif.

Dr. Tomiyasu is a member of the American Physical Society and Sigma Xi.

❖

J. Torkel Wallmark (A'48) was born in Stockholm, Sweden, on June 4, 1919. He received Civilingenjorsexamen in electrical engineering at the Royal Institute of Technology in Stockholm, in 1944, and Teknologie Licentiatexamen in 1947.

J. T. Wallmark

From 1944 to 1945 he worked as a tube designer with the A.B. Standard Radiofabrik, and from 1945 to 1947, he was with the Royal Institute of Technology, in Stockholm, Sweden, as a research engineer. In 1947 Mr. Wallmark was granted a fellowship by the American Scandinavian Foundation and spent a year with the RCA Laboratories, Princeton, N. J. At present, Mr. Wallmark is back at the Royal Institute of Technology in Sweden.

IRE Awards, 1957

Medal of Honor Award

RAYMOND A. HEISING

For his leadership in IRE affairs, for his contributions to the establishment of the permanent IRE Headquarters, and for originating the Professional Group system.

Founders Award

JULIUS A. STRATTON

For his inspiring leadership and outstanding contributions to the development of radio engineering as a teacher, physicist, engineer, author and administrator.

Morris Liebmann Memorial Prize

OSWALD G. VILLARD, JR.

For his contributions in the fields of meteor astronomy and ionosphere physics which led to the solution of outstanding problems in radio propagation.

Browder J. Thompson Memorial Prize

DUDLEY A. BUCK

For his paper entitled "The Cryotron—A Superconductive Computer Component," which appeared in the April, 1956 issue of the PROCEEDINGS OF THE IRE.

Vladimir K. Zworykin Television Prize

DONALD RICHMAN

For contributions to the theory of synchronization, particularly that of color subcarrier reference oscillator synchronization in color television.

Harry Diamond Memorial Award

GEORG GOUBAU

For his many contributions in ionospheric research and circuit theory and for his discovery of the surface wave transmission principle.

Joint Winners of the W. R. G. Baker Award

REYMOND J. KIRCHER

For his paper entitled "Properties of Junction Transistors," which appeared in the July-August, 1955 issue of the IRE TRANSACTIONS on Audio.

ROBERT L. TRENT

For his paper entitled "Design Principles of Junction Transistor Audio Amplifiers," which appeared in the September-October, 1955 issue of the IRE TRANSACTIONS on Audio.

D. RAYMOND FEWER

For his paper entitled "Design Principles for Junction Transistor Audio Power Amplifiers," which appeared in the November-December, 1955 issue of the IRE TRANSACTIONS on Audio.

New Fellows

L. G. Abraham

For contributions to the engineering of broad band coaxial transmission systems.

W. J. Albersheim

For contributions in the fields of sound reproduction and military electronics.

F. L. Ankenbrandt

For leadership in areas of military air navigation and communications.

Jack Avins

For contributions to the development of television receivers.

Laurence Batchelder

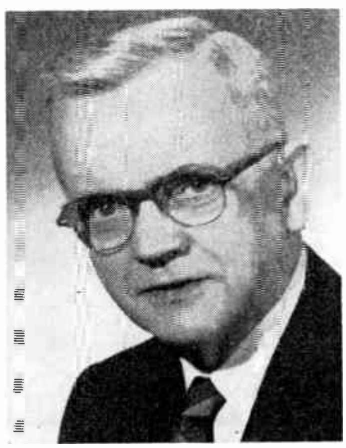
For contributions to the design and development of sonar equipment.

A. A. Barco

For contributions to radio and television receiver circuits.

W. T. Born

For applications of electronic techniques to geophysical exploration.



G. P. Bosomworth

For contributions to the use of electronics in the rubber industry.

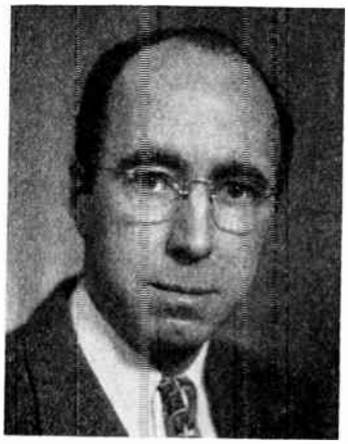
J. D. Cobine

For contributions to engineering education and gaseous electronics research.

New Fellows

G. C. Comstock

For contributions to the development and application of navigational radar.



Sidney Darlington

For contributions to network theory and to guidance and control systems.

Rinaldo DeCola

For contributions to the fields of military electronics and to television receivers.

Burgess Dempster

For pioneering in loudspeaker production, and contributions to engineering management.

Milton Dishal

For contributions in the field of circuits using modern network theory.

F. H. Drake

For contributions to the field of airborne communications equipment.

W. A. Edson

For contributions in the fields of education and microwave electronics.

G. S. Field

For contributions to ultrasonics and to the defense research program of the Royal Canadian Navy.

L. E. Flory

For contributions in the fields of industrial television and medical electronics.

New Fellows

SIDNEY FRANKEL

For contributions to the field of circuit techniques.

C. A. FRISCHE

For contributions in the field of electronic servo controls and related devices.

W. N. GOODWIN, JR.

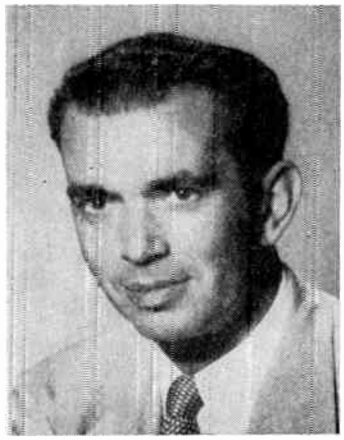
For contributions in the field of electrical measuring instruments.

GEORG GOUBAU

For contributions to the field of microwave transmission.

GERALD GROSS

For contributions to international regulation of telecommunications.



W. J. GRUEN

For contributions to the improvement of television receivers.

R. N. HALL

For contributions in the field of semiconductor devices.

P. G. HANSEL

For contributions in the fields of radio navigation and direction finding.

H. R. HEGBAR

For contributions to the development of analog computers and their applications.

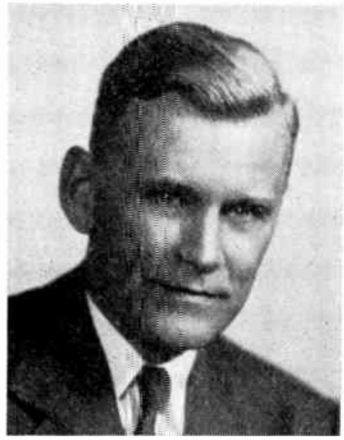
New Fellows

JAMES HILLIER

For contributions in the field of electron optics, particularly electron microscopy.

W. S. HINMAN, JR.

For contributions in the fields of meteorographic instrumentation, aircraft navigation and ordnance electronics.



GUNNAR HOK

For contributions to electronic science and education.

S. G. L. HORNER

For contributions to radio communications in Canadian Northern and Arctic regions.

H. W. HOUCK

For pioneering contributions in the field of broadcast reception.

W. H. HUGGINS

For contributions in the field of circuit theory.

J. E. KEISTER

For contributions in the field of television transmitters and military electronic equipment.

G. R. KILGORE

For pioneering work in high frequency electron tubes, and for leadership in the field of military electronic devices.

T. H. KINMAN

For contributions in high frequency research and semiconductor devices.

New Fellows

W. J. Kleen

For contributions to electron tube theory and microwave techniques.

Issac Koga

For contributions in quartz crystal techniques and engineering education.

L. M. Leeds

For contributions to the design of television equipment for studio and industrial applications.

Jesse Marsten

For contributions to the design and manufacture of electronic components.

D. W. Martin

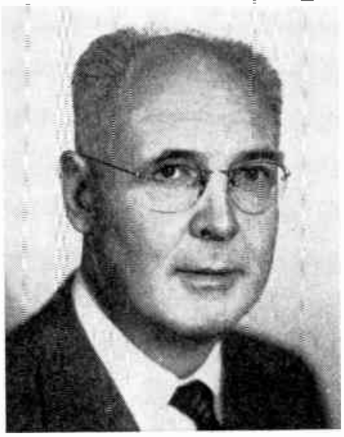
For contributions in the field of acoustics.

R. E. Mathes

For contributions to the development of terminal apparatus for point-to-point radio service.

J. W. Mauchly

For pioneering contributions to the field of electronic computers.



J. O. McNally

For contributions in the field of electron tubes.

Eugene Mittelmann

For pioneering in the field of industrial electronics.

New Fellows

H. K. Morgan

For engineering contributions in the fields of air communication and navigation.

W. J. Morlock

For pioneering work in sound systems, and for contributions to engineering management.

R. M. Morris

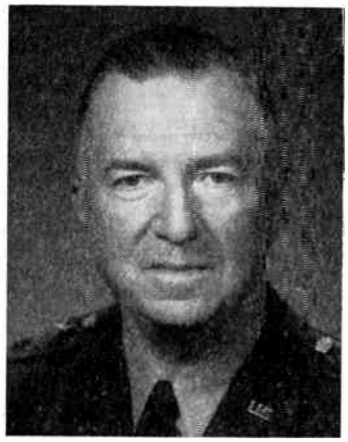
For contributions in the field of radio and television broadcasting.

F. L. Moseley

For contributions to the development of aircraft navigation systems and electronic instruments.

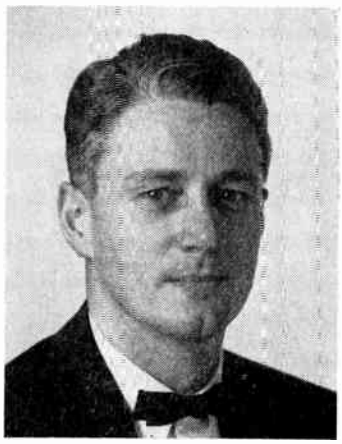
H. Q. North

For practical developments in the field of semiconductors.



J. D. O'Connell

For distinguished leadership in the field of military electronics.



L. E. Packard

For contributions in the field of electronic instrumentation.

Leon Podolsky

For contributions in the field of electronic component engineering.

J. L. Potter

For contributions as an engineer and educator.

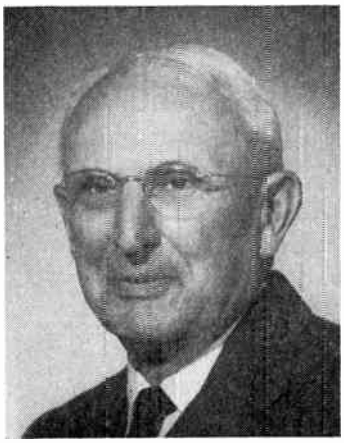
New Fellows

R. M. RYDER

For contributions to the development of microwave tubes and applications of transistors.

R. E. SAMUELSON

For leadership in research and development in the field of radio communication.



O. S. SCHAIRER

For distinguished service in fostering and administering electronic research.

E. H. SCHREIBER

For contributions to radio and television broadcasting.

M. H. SCHRENK

For pioneering achievements in naval aviation electronics.

O. H. SCHUCK

For contributions in the field of instrumentation and control.

J. C. W. SCOTT

For contributions in the field of ionospheric propagation.

L. D. SMULLIN

For contributions in the field of microwave tubes.

R. A. SYKES

For contributions to the development of quartz crystal units for filter networks and frequency control.

New Fellows

L. E. Thompson

For contributions to microwave communication systems and development of special purpose radio receivers.

G. S. Turner

For achievements in telecommunications and in their international regulation.

C. D. Tuska

For pioneering services to radio communications.

O. G. Villard, Jr.

For contributions to knowledge of the ionosphere and its role in the propagation of radio waves.

C. C. Wang

For basic contributions in the field of microwave tubes.

A. H. Waynick

For contributions in radio transmission, ionosphere research and engineering education.

Joseph Weil

For contributions to engineering education.

W. D. White

For achievements in the fields of air traffic control, information theory, countermeasures and radar.

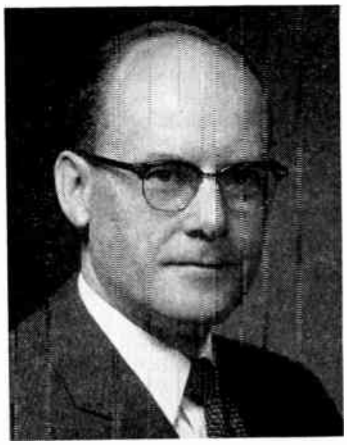
F. C. Williams

For contributions to the theory of noise in vacuum tubes, and in the fields of radar and digital computers.

New Fellows

S. B. Williams

For contributions to telephone switching systems and to computers.



A. K. Wing, Jr.

For contributions to the advance of vacuum tube techniques.

C. F. Wolcott, Jr.

For contributions as an engineer and executive, and for active participation in Institute affairs.

IRE News and Radio Notes

Television Conference Program for April 26–27 Is Revealed

The IRE Cincinnati Section in cooperation with the Professional Groups on Broadcast and Television Receivers, and Broadcast Transmission Systems is sponsoring the Eleventh Annual Technical Conference on Television at the Cincinnati Engineering Society Building, Cincinnati, Ohio, April 26–27, 1957.

The first day's schedule will feature a display of exhibits and a banquet, at which the speaker will be G. H. Brown, Chief Engineer, Commercial Electronics Products, Radio Corporation of America. The second day's program will consist of exhibits and technical sessions during which the following papers will be presented: *Practical Aspects of TV Tuner Design*, C. D. Nestlerode, Dumont Labs., Inc.; *A Constant Input-Impedance RF Amplifier for VHF TV Receiver*, H. B. Yin and H. M. Wasson, RCA; *A Transistorized Carrier System for Transmission of Television Signals*, L. G. Schimps, Bell Labs.; *Color TV Recording on Black-and-White Lenticular Film*, J. M. Brumbaugh, E. D. Goodale and R. D. Kell, RCA; *Transistor Receiver Video Amplifiers*, M. C. Kidd, RCA; *Transistor Design for Picture IF Stages*, R. J. Turner and P. E. Hermann, Philco Corp.; *Color Signal Distortion in Envelope Type of Second Detectors*, B. D. Loughlin, Hazeltine Research Corp.; *A Transistorized Horizontal Deflection System*, H. C. Goodrich, RCA; and *A New Approach to Horizontal Deflection Tube Testing*, G. M. Lankard, Sylvania Elec. Prod., Inc.

Information concerning advance registration or hotel reservations can be obtained from C. B. Shaw, Jr., Hangar Three, Lunken Airport, Cincinnati 26, Ohio.

IRE Annexes Five Chapters

The following chapters were officially approved by the IRE Executive Committee on February 6: PG on Aeronautical and Navigational Electronics, Joint New York, Long Island and Northern New Jersey Sections; PG's on Antennas and Propagation, Akron and Syracuse Sections; PG on Microwave Theory and Techniques, Syracuse Section; and PG on Vehicular Communications, Dallas Section.

On the same day, the establishment of the Nashville Subsection of the Huntsville Section was approved.

Conference Dates Released

The National Electronics Conference has just released its meeting dates for the next seven years. They are as follows: October 7–9, 1957; October 13–15, 1958; October 12–14, 1959; October 10–12, 1960; October 9–11, 1961; October 8–10, 1962; and October 7–9, 1963.

The schedule of the National Aeronautical and Navigational Electronics Conference for the next three years calls for meetings on May 13–15, 1957, May 12–14, 1958, and May 11–13, 1959.

Philadelphia IRE-AIEE Section Outlines Sessions on Nuclear Radiation for April and May

The Philadelphia Section of the IRE and the Educational Committee of the Philadelphia AIEE Section are jointly sponsoring the first of a series of planned annual symposiums. This year's symposium consists of a series of six lectures on design for nuclear radiation by speakers from Naval Research Laboratory, Atomic Energy Commission, Battelle Institute and General Electric Company. Meetings are held at the Philadelphia Electric Company's auditorium, Ninth and Sansom Sts., Philadelphia, Pa., at 8 P.M.

Herbert Rabin of the Naval Research Laboratory will speak on testing and measurement techniques, facilities for irradiation of materials, components and equipment, and methods of monitoring temperature radiation spectrum at the session scheduled for April 5. Paul Schall of Battelle Memorial Institute will discuss radiation damage and shielding effects on materials at the next session, slated for April 18. On April 25, a speaker from the General Electric Company

Calendar of Coming Events

British Radio & Electronic Component Show, Grosvenor House and Park Lane House, London, England, Apr. 8–11

Industrial Electronics Conference, Ill., Inst. of Tech., Chicago, Ill., April 9–10

First National Nuclear Instrumentation Conference, Atlanta, Ga., Apr. 10–12

Ninth Southwestern Regional Conference & Show, Shamrock-Hilton Hotel, Houston, Tex., April 11–13

National Simulation Conference, Shamrock-Hilton Hotel, Houston, Tex., April 11–13

PGTRC National Telemetering Symposium, Philadelphia, Pa., April 14–16

Special Technical Conference on Solid State Dielectric and Magnetic Devices, Catholic Univ., Wash., D. C., April 22–23

Symposium on Role of Solid State Devices in Electric Circuits. Engrg. Society Bldg., New York City, April 23–25

Region Seven Technical Conference & Trade Show, San Diego, Calif., April 24–26

Eleventh Annual Spring Television Conference, Engrg. Society Bldg., Cincinnati, Ohio, April 26–27

81st SMPTE Convention, Shoreham Hotel, Washington, D. C., April 29–May 3

Electronic Components Conference, Morrison Hotel, Chicago, Ill., May 1–3

Symposium on Image Formation and Measurement with Electronic Techniques, Morse Audit., Boston Museum of Science, May 2

Symposium on Microwave Ferrites and Devices & Applications, Western Union Auditorium, New York City, May 9–10

National Aero. and Nav. Electronics Conference, Dayton, Ohio, May 13–15

Fifth Annual Semiconductor Symposium of the Electrochemical Society, Statler Hotel, New York City, May 13–16

ACM 11th Annual Convention, Masonic Temple, Detroit, Mich., May 22–24

IRE-URSI Spring Meeting, Hotel Willard, Wash., D. C., May 22–25

PGPT First Annual Conference on Production Techniques, Willard Hotel, Washington, D. C., June 6–7

PGMIL First National Meeting, Sheraton-Park Hotel, Washington, D. C., June 17–19

ACM Twelfth National Meeting, Univ. of Houston, Houston, Tex., June 19–21

International Symposium on Physical Problems of Color Television, Paris, France, July 2–6

WESCON, Fairmont Hotel and Cow Palace, San Francisco, Calif., Aug. 20–23

URSI Twelfth General Assembly, Boulder, Colo., Aug. 22–Sept. 5

Special Technical Conference on Magnetic Amplifiers, Penn Sheraton Hotel, Pittsburgh, Pa., Sept. 4–6

Industrial Electronics Symposium, Morrison Hotel, Chicago, Ill., Sept. 24–25

National Electronics Conference, Hotel Sherman, Chicago, Ill., Oct. 7–9

will outline radiation effects on electrical equipment and devices. Radiation effects on electronic components will be the topic of J. C. Pigg of the Atomic Energy Commission on May 2. A speaker from the Air Research Center will deliver an address on "How to Design for the Nuclear Environment" at the concluding session on May 9.

Advance registration for all six lectures is $3.50 for IRE and AIEE members, and $4.00 for all others. Registration at the door is $4.50 for all lectures and $1.00 for individual meetings. All registration arrangements should be made with W. Kassimir, Atlantic Refining Co., 260 S. Broad St., Philadelphia 1, Pa.

WESCON Establishes New San Francisco Bay Branch Office

A new San Francisco Bay area office has been established by the Western Electronic Show and Convention, according to Don Larson, business manager. The location is 60 West 41st Avenue, San Mateo, Calif. This office will be headquarters for the 1957 WESCON, to be held at the San Francisco Cow Palace, August 20–23.

The new office will also be Bay Area headquarters for two sponsoring WESCON organizations, the West Coast Electronic Manufacturers' Association and the San Francisco IRE Section. A staff will maintain records on the electronics industry in the San Francisco Bay area, and the activities of these organizations which pool efforts in the presentation of the annual WESCON. The new office will also serve as a meeting place for group gatherings of the industry, with a complete conference area available.

The main office of WESCON remains in Los Angeles at 342 North La Brea Avenue.

IRE Plans Fall Publication of WESCON Convention Record

A plan to publish each year all available papers presented at the WESCON Electronic Show and Convention has been adopted by the respective boards of directors of the IRE and WESCON. The plan will go into effect with the 1957 WESCON, to be held August 20–23 in San Francisco, Calif

The new publication, to be called "IRE WESCON Convention Record," will be handled by IRE headquarters and will be issued during the fall. To avoid confusion with the Convention Record issued for the IRE National Convention, the latter has been renamed "IRE National Convention Record."

Details on prices and ordering procedures will be announced at a later date.

ISA Hosts First Nuclear Instrumentation Conference

Dr. W. F. Libby, of the United States Atomic Energy Commission, will address the first National Nuclear Instrumentation Conference, in Atlanta, Georgia, April 10–12. The conference is sponsored by the Instrument Society of America and its Southeastern Sections.

Dean Joseph Weil, of Univ. of Fla., is planning the conference program in his capacity of Director of ISA's Nuclear Industry Division. The three-day conference program will include technical sessions on reactor instrumentation (two sessions), industrial nuclear instrumentation, basic problems in nuclear instrumentation, health physics, radiation instrumentation, university nuclear instrumentation programs, and nuclear instrumentation in the medical field. Models of the Brookhaven medical reactor and the Oak Ridge graphite reactor will also be on display.

For additional conference information, write H. S. Kindler, Director of Technical Programs, Instrument Society of America, 313 Sixth Avenue, Pittsburgh 22, Pa. Exhibit information and space are available from F. J. Tabery, Exhibit Manager, 3443 South Hill St., Los Angeles 7, Calif.

Proceedings Now Available

Proceedings of the recent Automation Coding Symposium, held at The Franklin Institute, Philadelphia, Pa., will be published as a monograph under the auspices of the *Journal of The Franklin Institute*. Free copies will be sent to all symposium registrants, and others can purchase copies at $3.00 per copy.

Proceedings of the Conference on Magnetic Amplifiers, held at Syracuse, N. Y. April 5–6, 1956, are also available from IRE Headquarters at $4.00 per copy. The conference was held under the joint sponsorship of the IRE Professional Group on Industrial Electronics, the AIEE and the ISA.

WESCON Papers Deadline Set for May 1

Authors wishing to present papers at the 1957 WESCON Convention to be held in San Francisco on August 20–23 should send 100–200 word abstracts, together with complete texts or additional detailed summaries, to the Technical Program Chairman, D. A. Watkins, Stanford Electronics Laboratories, Stanford University, Stanford, California, by May 1 for consideration by the Technical Program Committee. Authors will be notified whether or not their papers have been accepted by June 1.

RETMA Sets Design Symposium

The Radio-Electronics-Television Manufacturers Association will sponsor a second symposium on applied reliability at the Hotel Syracuse, Syracuse, N. Y., June 10–11.

Problems to be discussed in technical sessions will be those of mechanical design, selection and use of components, proof of mature design, and case histories of reliable and unreliable designs. An evening panel session will discuss "Industry vs. Military Responsibility on Contract and Specification Control for Reliability."

Advance registration arrangements can be made with the RETMA Engineering Office, Room 650, 11 W. 42 St., New York 36, N. Y. at $4.00 per person.

courses are intended for engineers who find it necessary or who wish to obtain a basic understanding of the field, but who cannot spare more than a few days for this purpose.

The courses are built around the principles and application of measurement, communication and control. Course I will consist of the fundamentals in each of these fields and will include some basic work in nonlinear systems. Course II will take up applications of the fundamentals to more advanced problems. There will be four hours of lecture each morning and three hours of laboratory demonstration each afternoon. Extensive use will be made of computing, instrumentation and servo laboratories on the campus. The role of analog computing methods will be emphasized.

April 15 is the closing date for registration. Further information may be obtained by writing to Prof. L. L. Rauch, Room 1521, East Engineering Building, University of Michigan, Ann Arbor, Mich.

Technical Writers' Institute Is Scheduled for June 10–14

The fifth annual Technical Writers' Institute will be held at Rensselaer Polytechnic Institute, Troy, N. Y., June 10 through June 14. It is designed for those who supervise technical writing in business, industry, and the professions.

The workshop will include sessions on manuals and instruction books, reports, technical promotion, training programs, industrial films, and graphic and illustrative aids.

Among a special group of lecturers, in addition to the Rensselaer staff, will be D. D. Starr, Technical Information, Engineering Division, Chrysler Corp.; Charles Troupe, Service Publications, The Glenn L. Martin Co.; D. H. Green, Publications Engineering, Collins Radio Co.; H. J. Constantin, Illustrations for Publication, Sperry Gyroscope Co.; H. T. Sharp, Assistant Editor, Chemical Engineering Magazine; J. C. Hoffman, Product Information, Special Defense Projects, General Electric Co.; and F. C. Regan, Manager of Advertising for Abrasives, Behr-Manning Corp.

Additional information can be obtained by writing J. R. Gould, Director, Technical Writers' Institute, Rensselaer Polytechnic Institute, Troy, N. Y.

F. W. Brown Named to International Radio Conference

F. W. Brown (A'54), Director of the Boulder Laboratories of the National Bureau of Standards, has been appointed by the chief of the telecommunications division of the U. S. Department of State to an important committee chairmanship of a forthcoming international conference. As chairman of the committee on technical questions, Dr. Brown will be concerned with securing information on such things as definitions, designations and classes of emissions, frequency tolerances, interference and radio propagation.

He will head the seventy-one-member committee on technical questions of the International Radio Conference scheduled to meet in Geneva, Switzerland, in 1959.

Boston Section Co-Sponsors Symposium on Images May 2

The New England Section of the Optical Society of America and the IRE Boston Section are planning a symposium and an exhibition of apparatus dealing with image formation and measurement with electronic techniques. The symposium will be held in the Morse Auditorium, Boston Museum of Science, May 2. The exhibition will also be held at Science Park, open to the public the evening of May 1, and to only scientists and their guests May 2. Speakers and topics already scheduled are: G. R. Harrison, *Electro-Optics in a New Era;* Otto Schade, *Instrumentation for Measuring the Optical Fine-Wave Spectrum of Image Forming Devices;* S. S. West, *The Flying Spot Microscope;* and B. F. Burke, *Radio Telescopes.* John Patterson will also give a talk and demonstration of the planetarium associated with the Boston Museum.

Committee members of this symposium are: Frederick Brech, K. C. Black, Walter Driscoll, D. J. Lovell, and F. D. Smith. Advance registrations can be made with Mr. Brech, 26 Farwell St., Newtonville, Mass. There is no registration fee.

Military Affiliate Radio System Announces Speakers

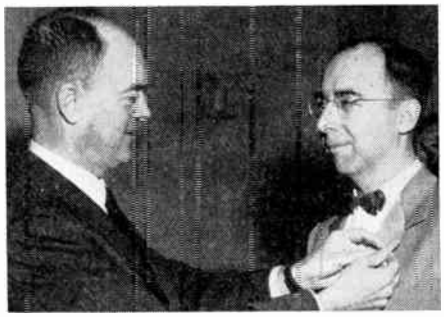
At 2 p.m. (EST) April 7, the Air Force MARS Eastern Technical Net will present J. P. Costas who will read a paper on his double sideband amplitude modulated suppressed carrier transmission system. An half-hour question-and-answer session will follow. Other speakers scheduled are: M. G. Crosby on product detectors for April 14; Frank Lester on vhf and uhf techniques for April 21; and William Kaufman on control theory for April 28.

The Eastern Technical Net, a part of the Air Force Military Affiliate Radio System educational program, is conducted by J. H. McCoy AF2IYX, over the air at 7635 kc. A Southwest Net has Sunday afternoon programs 1–3 p.m. (CST) at 7305 kc (soon to move to 7460 kc), and a Western Net will soon be scheduled at 7832.5 kc.

Peter Waterman Wins Highest Civilian Award From Navy

The Navy's highest civilian award was presented to Peter Waterman (S'42–A'45–M'55) of Washington, D. C. recently for his leadership and scientific accomplishments which have materially aided the nation's defense against the danger of enemy attacks.

Mr. Waterman, an electrical engineer at the Naval Research Laboratory, received the Distinguished Civilian Service Award from Assistant Secretary of the Navy for Air Garrison Norton.



Garrison Norton, *left*, and Peter Waterman, *right*.

A native of Hartford, Connecticut, Mr. Waterman received his Bachelor of Science degree in electrical engineering from the University of Vermont in 1943. He joined the NRL upon graduating and headed the servomechanisms group which was engaged in work on fire control servomechanisms.

In October 1945, Mr. Waterman was awarded the Navy's second highest civilian award, the Meritorious Civilian Service Award, for directing the work "which made important contributions to the improvement of Navy gun fire control equipment."

Prior to this he had been made section head of the director development section in the fire control division. This section was later renamed the equipment research section and because of its important projects and growth in personnel, it was made a branch in 1948 with Mr. Waterman as its head.

Mr. Waterman is the author of numerous papers in the field of missile guidance, and is a member of the American Physical Society and the A.I.E.E.

A conference on solid state devices, sponsored by the IRE, AIEE, Aeronautical Research Laboratory, Wright Air Development Center, and Catholic University of America, will be held on the latter's campus April 22–23. Pictured here is the local committee in charge (*left to right*): J. C. Michalowicz, chairman; J. H. Kilcoyne Jr., publicity; C. F. Pulvari, papers; G. E. McDuffie, Jr., registration; R. H. Elbourn, finance; and I. L. Cooter, hotel and transportation.

Technical Committee Notes

The **Antennas and Waveguides** Committee met at IRE Headquarters on January 30 with Chairman Henry Jasik presiding. The entire meeting was devoted to the review of comments received on the Proposed Standards on Antennas and Waveguides: Waveguide and Waveguide Component Measurements. After consideration this document was approved for submission to the Standards Committee.

Chairman Iden Kerney presided at a meeting of the **Audio Techniques** Committee held at IRE Headquarters on January 22. The major portion of the meeting was devoted to the review of the Proposed Standard on Audio Techniques: Definitions of Terms being prepared by Subcommittee 3.1 on Audio Definitions under the chairmanship of D. S. Dewire.

The **Electron Tubes** Committee met at IRE Headquarters on February 15 with Chairman P. A. Redhead presiding. The chairman reported that the Standards Committee had approved the revisions in the Gas-Filled Radiation-Counter Tubes Definitions and the Introduction to the Standard. This now completes all work on definitions prepared by this committee. The Standards on Electron Tubes: Definitions of Terms, 1957, which consists of approximately 530 terms, will be published shortly in the IRE PROCEEDINGS.

The committee discussed, amended and unanimously approved the section on Storage Cathode-Ray Tubes of the Proposed Standards on Electron Tubes: Methods of Test.

Vice-Chairman M. J. E. Golay presided at a meeting of the **Information Theory and Modulation Systems** Committee held at IRE Headquarters on January 17, 1957. The entire meeting was devoted to discussion of definitions of terms, which is now in preparation in the committee.

The **Nuclear Techniques** Committee met at the National Bureau of Standards in Washington, D. C. on January 17 with Chairman G. A. Morton presiding. The entire meeting was devoted to discussion of definitions of terms, now in preparation in the committee.

Chairman R. M. Showers presided at a meeting of the **Radio Frequency Interference** Committee held at IRE Headquarters on January 17, 1957. The major portion of the meeting was devoted to the discussion of. and the drafting of comments, on IEC proposal 12-1 (Central Office) 12 concerned with radiation.

The **Standards** Committee met at IRE Headquarters on February 14 with Chairman M. W. Baldwin, Jr. presiding. The revisions in the Standard on Electron Tubes: Gas-Filled Radiation-Counter Tube Definitions were discussed, amended and unanimously approved.

There was a partial review of the Proposed Standard on Navigation Aids: Direction Finder Measurements.

The Introduction to the Standards on Electron Tubes: Definitions of Terms, 1957 was discussed, amended and unanimously approved.

Books

Schaltungstheorie und Messtechnik des Dezimeter- und Zentimeter-Wellengebietes by A. Weissfloch

Published (1954) by Birkhäuser Verlag, Basel. Switzerland, and Stuttgart, Germany. 305 pages+3 index pages. Illus. $9\frac{1}{4}\times6\frac{1}{4}$.

In the broad field of microwave theory and measurement Dr. Weissfloch's book is in many respects unique. While its scope is roughly comparable with that of other books, its approach is different. Its primary mathematical tool is the geometry of the bilinear transformation. The general usefulness of graphical methods based on the properties of this transformation is familiarly illustrated in the form of circle diagrams and impedance charts such as the Smith chart. But this is only one aspect of the systematic account presented in this book of geometrical procedures, based on the properties of linear transformations and applied to the theory of four-poles and uniform transmission lines.

The first chapter begins with a careful and well-illustrated introduction into the mapping properties of the bilinear transformation. These are then used to develop in detail what the author aptly calls circular-geometrical four-pole theory. This begins with the lossless T-section, continues with the lossless general four-pole and the lossy four-pole. The second chapter begins with a review of the properties of transmission lines and waveguides; it includes the distributions of current and voltage, impedance measurement, and the transfer of power. After this introduction the general problem of four-poles connected between two sections of homogeneous transmission line or waveguide is considered in detail. The well-known Weissfloch transformer theorem is a central part of this discussion. The chapter concludes with many applications of the transformer theorem to various discontinuities along lines and guides. Chapter 3 is concerned with more complicated junctions in homogeneous lines and guides including two four-poles spaced along a line, six-poles, magic tees, and directional couplers. The last chapter consists of a detailed study of the general problem of matching and compensation.

In one respect, the book is incomplete since it does not include a discussion of Deschamps' graphical method for the analysis of waveguide junctions. This omission is presumably a consequence of the fact that the manuscript was completed before the publication of the Deschamps method.

Dr. Weissfloch's book is well organized and clearly written. The material is carefully arranged and integrated, and very well illustrated. The novel approach to the general problem of microwave theory and measurement is refreshing and illuminating. And it is presented in a manner that never leaves the reader in doubt that the author is an expert in the field.

R. W. P. KING
Gordon McKay Professor
Cruft Laboratory
Harvard University
Cambridge, Mass.

1957 IRE NATIONAL CONVENTION RECORD

All available papers presented at the 1957 IRE National Convention will appear in the IRE NATIONAL CONVENTION RECORD to be published in July. The IRE NATIONAL CONVENTION RECORD will be issued in ten Parts, with each Part devoted to related subjects. The papers for each session are listed on pages 373–406 of the March issue.

Instructions on Ordering

1. If you are a member of a Professional Group and have paid the group assessment by April 30, you will automatically receive, free of charge, that Part of the IRE NATIONAL CONVENTION RECORD pertaining to the field of interest of your group, as indicated in the chart below.

2. If you are not a member of an IRE Professional Group, IRE NATIONAL CONVENTION RECORD Parts may be purchased at the prices listed in the chart below. Orders must be accompanied by remittance, and to assure prompt delivery, should be sent immediately to The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y.

IRE NATIONAL CONVENTION RECORD

Part	Title	Free to Paid Members of Following IRE Professional Groups	Prices for Members (M) College & Pub. Libraries & Sub. Agencies (L) Non-Members (NM)		
			M	L	NM
1	Microwave; Antennas & Propagation **Sessions: 3, 14, 21, 32, 39, 48, 55**	Antennas & Propagation Microwave Theory & Techniques	$3.25	$7.80	$ 9.75
2	Circuit Theory; Information Theory **Sessions: 7, 15, 22, 29, 42, 50**	Circuit Theory Information Theory	3.50	8.40	10.50
3	Electron Devices; Receivers **Sessions: 8, 16, 23, 31, 33, 38**	Electron Devices Broadcast & TV Receivers	2.75	6.60	8.25
4	Computers; Automatic Control; Medical Electronics **Sessions: 1, 9, 18, 26, 37, 41, 49**	Electronic Computers Medical Electronics Automatic Control	4.00	9.60	12.00
5	Instrumentation; Telemetry **Sessions: 34, 46, 47, 53, 54**	Instrumentation Telemetry & Remote Control	2.50	6.00	7.50
6	Components; Production; Industrial Electronics **Sessions: 25, 40, 43, 44, 51**	Component Parts Production Techniques Industrial Electronics	2.00	4.80	6.00
7	Audio; Broadcast **Sessions: 11, 19, 20, 27, 35**	Audio Broadcast Transmission Systems	2.50	6.00	7.50
8	Aeronautical; Communications; Military Electronics **Sessions: 2, 5, 6, 10, 17, 24**	Aeronautical & Navigational Electronics Communications Systems Military Electronics Vehicular Communications	3.50	8.40	10.50
9	Ultrasonics; Nuclear Science **Sessions: 4, 12, 28, 36**	Ultrasonics Engineering Nuclear Science	1.25	3.00	3.75
10	Quality Control; Engineering Management **Sessions: 13, 30, 45, 52**	Reliability & Quality Control Engineering Management	1.75	4.20	5.25
	Complete Set (10 Parts)		$27.00	$64.80	$81.00

1957 National Symposium on Telemetering

SPONSORED BY THE IRE PROFESSIONAL GROUP ON TELEMETRY & REMOTE CONTROL

SHERATON HOTEL

PHILADELPHIA, PENNSYLVANIA

APRIL 14–16, 1957

The National Symposium on Telemetering and its exhibits, sponsored by the IRE Professional Group on Telemetry and Remote Control, will be held at the Sheraton Hotel, Philadelphia, Pa., April 14–16.

A. S. Westneart, Jr. of Applied Science Corp., papers chairman, stated that thirty-two papers on telemetering theory systems and applications, remote control theory systems and applications, and system elements for telemetering and remote control will be presented.

Committee chairmen are: L. P. Clark, National Chairman; C. W. Kropp, Vice-Chairman; Gordon Jacobs, Finance; Fred Fanella, Exhibits; T. R. Gregory, Facilities; Henry Kaplan, Program; and D. M. Jones, Publicity.

Advance registration arrangements should be made with H. W. Royce, The Glenn L. Martin Co., Baltimore 3, Md.

APRIL 14

Exhibits open.

APRIL 15

9:00 A.M.

Session I

Remote Control Systems

Phase Angle Analogs in Out-of-Sight Control Instrumentation, C. L. Parish, Chance Vought Aircraft, Inc.

A Wide-Band Microwave Link for Telemetering, R. E. Glass, Sandia Corp.

A Low-Level, High Speed Sampling System, J. P. Francis, Magnavox.

A Remote Control System for Airborne Test Vehicles, Lyman Nickel, North American Aviation, Inc.

Petroleum Production Telemetering and Remote Control Systems, J. C. Stilley, Arabian American Oil Company.

Session II

Telemetry Systems I

Separation of Frequency Modulated Carriers Using Cascaded Phase Locked Oscillators, G. W. Preston, General Atronics Corp.

Flight Data System, G. E. Sandgren, Victor Adding Machine Co.

Restrictive Bandwidth PDM Systems, A. S. Westneat, Applied Science Corp. of Princeton.

Extension of FM/FM Capabilities, H. O. Jeske, Sandia Corp.

Telemetering System for X-17 Missile, R. M. Powell, Lockheed Missile Systems.

2:00 P.M.

Session III

Transistor Applications

Transistor Circuits Applied to Telemetering, J. H. Smith, Texas Instruments Inc.

A Low-Level Electronic Sub-Commutator, J. M. Walter, Jr. and J. H. Searcy, Radiation, Inc.

Solid State Pulse Width Modulator, Henry Kaplan, Burroughs, Inc.

High-Speed, High-Accuracy Multiplexing of Analog Signals for Digital Systems, R. E. Marquand, Radiation, Inc.

Low-Level Transistor Amplifier for DC Measurements in Telemetering, T. E. Smith and H. F. Harris, Texas Instruments, Inc.

Progress Report on a Solid State FM-FM Telemetering System, E. Y. Politi, Lockheed Missile Systems Division.

Session IV

Data Processing I

Operation of Airborne Telemetering, E. Shanahan, Martin-Baltimore.

Data Processing, Analog or Digital, A. S. Westneat, Applied Science Corp. of Princeton.

Application of Telemetry to Flight Testing, J. A. See, Boeing Airplane Co.

Automatic Reduction of Telemetering Oscillograph Data, Donald Segel, Litton Industries, and Graham Tyson, Northrop Aircraft.

A Digital Method for FM Telemetering Measurements, John Humphries, Dynac, Inc.

Space Ship Telemetering, Hans Scharla-Neilsen, Radiation, Inc.

APRIL 16

9:00 A.M.

Session V

Data Processing II

A Direct Computer Controlled Data Editing System, B. M. Gordon, Epsco, Inc.

Handling PCM on the Ground—Some Problems, Some Solutions, T. Hagan, Epsco, Inc.

Telemetry Magnetic Tape Recorder/Reproducer, R. E. Hadady, and S. Gilman, Consolidated Electrodynamics Corp.

Specification and Design of Multi-Channel Sampling Devices Relative to Telemetering System Requirements, John Brinster and E. B. Garretson, General Devices, Inc.

An Automatic Telemetry Reduction System, E. T. Hatcher, RCA Service Co.

Session VI

Components

A Rugged RF Power Amplifier for Use in the 200 MC Telemetry Band, D. D. McRae, Radiation, Inc.

Completely Transistorized Strain Gage Oscillator, W. H. Foster, Electronic Engineering Co. of Calif.

A New Transistor-Magnetic F/M Discriminator, G. H. Barnes and R. M. Tillman, Burroughs Corp.

An Electronic Commutator Using Transistors, Peter Slavin, Canadair Ltd.

Transistorized Pulse Width Keyer, J. A. Riedel, Applied Science Corp. of Princeton.

2:00 P.M.

Session VII

Theory

Notation and Characteristics of 2-Level Codes, George Birkel, Jr., Radiation, Inc.

The Bandwidth Requirements of PDM, Frank Rock, Applied Science Corp. of Princeton.

Transmission of Information over PCM Equipments, F. Mansfield Young, Epsco, Inc.

Theoretical Considerations of Practical Data Transmission Systems, F. Mansfield Young, Epsco, Inc.

Noise and Bandwidth in PDM/FM Telemetering, K. M. Uglow, Consulting Engineer.

Session VIII

Telemetering Systems II

PCM Data Collecting and Recording System Designed for Airborne Use, J. P. Knight, Radiation, Inc.

Coding for Suppression of Noise and Interference in Airborne PCM Telemetering Systems, H. F. Harmuth, General Electric Advanced Electronics Center.

Time Interval Telemetering System, Ned Wilde, Armour Research Foundation.

Telemetering Receiving System at the Air Force Missile Test Center, H. A. Roloff, RCA Service.

PDM-PAM Conversion System, R. L. Kuehn and Walter T. Johnston, R. M. Parsons.

Low Level Commutation System for Telemetry Applications, F. Shandelman, A. Hartung, and H. Golden, Teledynamics Inc.

Seventh Region IRE Conference

San Diego, California, April 24–26, 1957

The IRE Region Seven is sponsoring a regional conference at the Conference and Federal Buildings, Balboa Park, San Diego, California, April 24–26. Over two hundred technical exhibits will also be on display.

There will be no pre-registration by mail. Those planning to attend should send hotel reservations to the San Diego Convention and Tourist Bureau, 924 Second St., San Diego, Calif. by April 10.

April 24
9:45–12:00 a.m.

Microwaves

Chairman: R. G. Stegen, Canoga Corp.

Strip-Line Directional Couplers, J. K. Shimizu, Stanford Research Institute.

A Logarithmic Impedance Chart, Jack Wills, Canoga Corp.

The Microwave Interferometer as a Versatile Experimental Tool, F. M. Millican, Convair.

Some Design Factors for the Magnetron-Isolator Combination, R. A. Krogh, Litton Industries.

2:00–4:30 p.m.

Computer Applications

Chairman: D. M. Finnigan, Stanford Research Institute.

Designing Data-Processing Systems: Matching Component Configurations to Requirements, Michael Montalbano, Kaiser Steel Corp.

Trends in Computer Applications Throughout the World, E. S. Calhoun, Stanford Research Institute.

Pan American's 705 Data-Processing Center in Action, J. S. Woodbridge, Pan American World Airways.

Electronic Computers and Management Control, G. Kozmetsky, Litton Industries.

Electron Devices

Chairman: J. E. Keister, General Electric Co.

High-Speed Silicon Junction Diode, T. La Chapelle and C. Levi, Pacific Semi-Conductors.

A New High Frequency NPN Silicon Transistor, A. B. Phillips and A. M. Intrator, General Electric Co.

High Voltage Silicon Rectifiers, J. W. Thornhill, Texas Instruments.

7:00–9:00 p.m.

Professional Management

Chairman: D. G. Burgar, Cubic Corp.

Management of a Navy Laboratory, Capt. J. M. Phelps, USN, Naval Electronics Lab.

Growth from Performance, David Packard, Hewlett-Packard Co.

Balanced Management Concept, C. B. Thornton, Litton Industries.

Student Papers

Chairman: C. Moe, San Diego State College.

Titles and authors of papers not available at time of publication.

April 25
9:00–12:00 a.m.

Antennas and Propagation

Chairman: J. B. Smyth, Smyth Research Associates.

Meteoric Ionization and its Applications, Von R. Eschleman, Stanford Univ.

Antenna Array Design Techniques, R. W. Clapp, Hughes Aircraft.

Tropospheric Propagation, J. B. Smyth, Smyth Research Associates.

Data Handling and Automation

Chairman: L. M. Silva, Beckman Instruments.

Systems Engineering, Dr. Chien, Beckman Instruments Corp.

Applications of Computers and Control Optimization in Industrial Plants, Dr. Williams, Monsanto.

Digital Computer Control, Geoffrey Post, Litton Industries.

Critical Evaluation of Analog and Digital Computers in Process Control Systems, Dr. Stout, Ramo-Wooldridge Corp.

Application of Linear Program to the Oil Industry, Dr. Garvin, California Research.

1:30–4:30 p.m.

Recent Developments in Loudspeakers

Chairman: L. L. Beranek, Bolt, Beranek and Newman, Inc.

Electrostatic Speakers, R. B. Goldman, Philco Corp.

Electrostatic Loudspeaker Design, Arthur Janszen, Janszen Labs.

Loudspeakers for the Generation of High Intensity Sound, John Hilliard, Altec Lansing.

Radiation Impedance of Acoustical Arrays, Gordon Martin, Naval Elec. Lab.

Electronic Aids to Air Navigation

Chairman: D. M. Stuart, Civil Aeronautics Administration.

VORTAC, Peter Caporale, CAA.

Requirements for Future Air Traffic Control Systems, G. B. Litchford, Special Assistant to the President.

System Considerations in the Design of an Air Traffic Control Radar Beacon System, W. N. Pike, Air Navigation and Development Board.

Self-Contained Navigation Aids and the Common System of Air Traffic Control, Nathaniel Braverman, WADC.

Improvements in Air Traffic Control, F. S. McKnight, CAA.

Collision Avoidance Systems, G. R. Schneider, Collins Radio Corp.

April 26
9:00–11:30 a.m.

Electron Tubes

Chairman: N. Moore, Litton Industries.

Ceramics in Microwave Tubes, C. E. Ward, Varian Associates.

Ceramic Tube Design Considerations for High Temperature Operation, R. E. Moe, General Electric Co.

The L-3028 Family of Magnetrons—Rugged Tubes of Many Applications, H. W. Smith, Litton Industries.

Electronic Phaseshifters and Switches at UHF, R. H. Geiger, White Elec. Devices.

A High Voltage Beam Switch Tube, M. E. Levin, Eitel McCullough, Inc.

Circuit Application of High Temperature Components, A. H. Dicke and C. E. Doyle, Wright Air Development Center.

International Geophysical Year

Chairman: R. A. Helliwell, Stanford Univ.

Interplanetary Medium, R. M. Bracewell.

International Geophysical Year Scatter Sounding Network, A. M. Peterson.

Whistlers During International Geophysical Year, R. A. Helliwell, Stanford Univ.

2:00–5:00 p.m.

Instrumentation

Chairman: M. L. Klein, North American Aviation, Inc.

LITTON 20 Digital Differential Analyzer, R. W. Rutishauser, Litton Industries.

Millisadic and its Application to the Processing of Commutated PDM and PAM Data, William Kneen, Consolidated Electro-Dynamics Corp.

Sound Pressure Instrumentation of Rocket Engines, J. R. Wood, North American Aviation, Inc.

A Special Purpose Arithmetic Unit for an In-Line Processor, M. J. Mendelson, Norden-Ketay Corp.

Nuclear Activation and Damage

Chairman: James Crawford, Oak Ridge National Labs.

Nuclear Radiation in Electronic Design, E. G. K. Schwarz, Convair.

Effect of Reactor Radiation on the Electrical Properties of Electronic Components, P. S. Miglicco, Convair.

The Effects of Nuclear Radiation on Semiconductor Electronic Components—A Preliminary Study, M. A. Zavier, Inland Testing Labs.

Effects of Nuclear Radiation on Electronic Components, R. D. Shelton.

Radiation Effects in Semiconductor Components, J. C. Pigg, Oak Ridge National Labs.

Effect of Neutron and Gamma-Ray Irradiation on the Dielectric Constant and Loss Tangent of Some Plastic Materials, R. A. Weeks, Oak Ridge National Labs.

1957 Electronic Components Symposium

MORRISON HOTEL, CHICAGO, ILLINOIS, MAY 1, 2, 3, 1957

The 1957 Electronic Components Symposium, co-sponsored by four leading electronic organizations, will be held May 1–3 at the Morrison Hotel, Chicago, Ill.

More than one thousand persons are expected to attend the three-day meeting dealing with the latest developments in electronic components.

The symposium is sponsored annually by the IRE, American Institute of Electrical Engineers, Radio-Electronics-Television Manufacturers' Association, and West Coast Electronic Manufacturers' Association.

The meeting also has the active participation of agencies of the Department of Defense and the National Bureau of Standards.

Papers will be given in eight areas: high temperature components, radiation effects, component reliability, passive components, active components, instrumentation and measurements, materials development, and general component needs.

Symposium committee chairmen are: R. M. Soria, Amphenol Electronics Corp., General Chairman; V. H. Disney, Armour Research Foundation of Illinois Institute of Technology, Technical Program; J. S. Powers, Bell and Howell Company, Arrangements; J. H. Enenbach, Illinois Bell Telephone Company, Finance; R. R. Jenness, Northwestern University, Proceedings, and V. J. Danilov, Illinois Institute of Technology, Publicity.

Information concerning the symposium can be obtained by writing to J. S. Powers, Electronic Components Symposium, 84 E. Randolph St., Chicago 1, Ill.

WEDNESDAY, MAY 1

9:30 a.m.

Session I: Introductory Session

Chairman: R. M. Soria, Amphenol Electronics Corp.

Introductory Remarks by R. M. Soria, Conference Chairman.

Military Requirements in Electronics, Brig. Gen. E. F. Cook, Commanding General, Signal Corps Engineering Laboratories.

Advantages Which Unified Specs Can Bring to National Defense, R. Soward, Convair.

Future Electronic Component Requirements for the Air Force, Col. W. Donics, USAF Asst. Chief, Communications and Electronics Div., Deputy Commander, Research and Development, Air Research and Development Command.

Choose a Non-Standard Part Wisely!, L. D. Harris, General Electric Co.

Components for Weapons Systems, R. H. Griest, Hughes Aircraft Co.

Some Aspects of Canadian Component Development, F. Simpson, Chief of Components Research Section, Defense Research Telecommunications Establishment, Ottawa, Canada.

12:30 p.m.

Luncheon

Needs for Engineering Training in Universities, R. A. Ramey, Manager, New Products Engineering Dept., Westinghouse Electric Corp.

2:45 p.m.

Session II: Components I

Chairman: J. J. Drvostep, Sperry Gyroscope Co.

New Developments in Piezoelectric Ceramic I-F Bandpass Filters, D. Elders and E. Gikow, Signal Corps Engineering Laboratories.

Properties of Ferroelectric Devices as Current Regulating and Frequency Determining Elements, C. Rosenberg, Bell Telephone Laboratories, Inc.

High Speed Magnetic Switches for Memory Matrices, D. R. Erb, Westinghouse Electric Corp.

A High-Frequency Ferrite Delay Line for Phase Modulation, C. F. Spitzer, General Electric Co.

A Miniaturized Quartz Crystal Unit for the Frequency Range 2 KC to 16 KC, D. M. Ruggles, Bell Telephone Laboratories, Inc.

Low Loss Ultrasonic Quartz Delay Lines with Barium Titanate Transducers, C. A. Bieling, Bell Telephone Laboratories, Inc.

7:30 p.m.

Session III: Nuclear and Environment Studies

Chairman: F. E. Wenger, Air Research and Development Command.

The Effects of Nuclear Radiation on Electronic Components, Dr. R. Shelton, Admiral Corp.

An Investigation of the Effects of Nuclear Radiation on Transistors, A. J. Schwartz and D. B. Kret, Radio Corporation of America.

Improved Reliability Through Electronic Part Development and Failure Rate Studies, J. Gruol, Signal Corps Engineering Laboratories.

Observations of Component-Part Debugging in Complex Electronic Equipment, J. A. Connor and F. A. Hartshorne, Radio Corporation of America.

Hermetically Sealed, Miniaturized Rotary Switch, L. G. Brodrick, P. R. Mallory and Co., Inc.

Trends in Shock, Vibration and Sound Energy Simulation, C. A. Golueke, Wright Air Development Center.

THURSDAY, MAY 2

9:00 a.m.

Session IV: Components II

Chairman: K. V. Newton, Bendix Aviation Corp.

Wire Type Solid Electrolyte Tantalum Capacitors, R. J. Millard, K. N. Lambert, and D. B. Peck, Sprague Electric Co.

A Solid Electrolyte Battery, B. F. Wagner, General Electric Co.

Complementary Graded Base Switching Transistors, D. E. Bode and R. E. Swanson, IBM Research Center.

Miniature Tuners for Transistor Circuits, Capt. C. K. Greene, Wright Air Development Center.

A High Quality Modular RF Transformer Package for Printed Circuit or Conventional Wiring Systems, D. M. Lisbin, Westinghouse Electric Corp.

Pulse Transformer Design Chart, R. Lee, Westinghouse Electric Corp.

12 Noon

Luncheon

History, Present Status, and Future Predictions of Needs and Components, P. S. Darnell, Director, Military Apparatus Division, Bell Telephone Laboratories, Inc.

2:30 p.m.

Session V: High Temperature Investigations and Development

Chairman: J. A. Osborn, Westinghouse Electric Corp.

High Temperature Radiation Resistance Resistors, A. O. Liermann and C. W. Heath, General Electric Co.

Ultra High Temperature—Fixed Resistors, E. Hauth, International Resistance Corp., H. Packer, P. B. Mallory Co., Inc., and E. Miller, Wright Air Development Center.

A High K Ceramic Capacitor for 200° C Application, C. A. Shaw, Onondaga Pottery Co.

Ultra High Temperature (500° C) Miniaturized Power Transformer and Inductor Materials, J. F. Rippin, Jr., Wright Air Development Center, and G. Walter, General Electric Co.

High Temperature—Magnetic Amplifiers, M. Frank, Wayne Engineering Research Institute, and K. A. Jellison, Wright Air Development Center.

High Temperature Sub-Assembly Techniques, C. N. Hood, General Electric Co.

FRIDAY, MAY 3

9:00 a.m.

Session VI: Instrumentation and Measurements

Chairman: G. Shapiro, National Bureau of Standards.

Automatic Data Taking Device for Transistors, J. Alman, Remington Rand Univac, H. Cary, Battelle Memorial Institute, and V. Walter, Inland Testing Laboratories.

Automatic Production Testing of Printed Wire Modules, E. D. Davis and H. S. Dordick, Radio Corporation of America.

An Automatic Data-Recording System, G. H. Jenkinson and J. E. Drennen, Battelle Memorial Institute.

A Procedure for Determining the Equivalent Circuit Elements Representing Ceramic Transducers Used in Delay Lines, A. H. Meitzler, Bell Telephone Laboratories, Inc.

The ASTRAMATIC System for Automatic Production Testing, Data Recording, *and Statistical Analysis*, E. Hoo, Electronic Control Systems, Inc.

Automation of Precision Potentiometer Functional Conformity Measurements, A. Blaustein, Fairchild Controls Corp.

12 Noon

Luncheon

Electronic and Air Power Progress, Lt. Gen. C. S. Irvine, Deputy Chief of Staff, Material (AFMDC) Headquarters, U. S. Air Force.

2:30 p.m.

Session VII: Materials

Chairman: W. S. Franklin, J. E. Fast and Co.

Evaporated Magnetic Films, D. Moore, Wright Air Development Center.

Organosilicon Compounds as Insulation for Miniature Power Transformers for High Temperature Operation, F. T. Parr, Westinghouse Electric Corp.

Protection of Electronic Components During High Temperature Transients Using Heat Storage Materials, D. M. Trones, Minneapolis-Honeywell Regulator Co.

The Limitations of Potting Compounds in Aircraft Connector and Cable Assemblies, V. D. Elarde, Amphenol Electronics Corp.

Progress Report on the Development of Low-Loss, High Temperature, Coaxial Cables, W. F. Croft and E. T. Pfund, United Electrodynamics, and Capt. B. Suverkrop, Wright Air Development Center.

Copper Clad Fluoroply Multilayer Laminates, L. B. Allen, D. E. McElroy, and S. J. Stein, International Resistance Co.

1957 Annual PGMTT Meeting

The 1957 Annual PGMTT Meeting, under the aegis of the IRE Professional Group on Microwave Theory & Techniques, and the IRE New York, Northern New Jersey and Long Island Sections, will take place at Western Union Auditorium, 60 Hudson St., N. Y., N. Y., May 9–10. The theme of the meeting is "Microwave Ferrites and Related Devices, and Their Applications."

Samuel Weisbaum is in charge of the technical program. Moe Wind, facilities; J. W. Kearney, treasurer, and T. N. Anderson, publication and organizing, are the other members of the meeting committee. Members of the steering committee include B. J. Duncan, R. MacVeety, Jack Melchor, R. D. Wengenroth, W. W. Mumford, H. E. D. Scovil, E. N. Torgow, and S. W. Rosenthal.

All papers presented at the meeting will be submitted to the TRANSACTIONS of the PGMTT for consideration.

Attendance at this meeting is limited so early registration is urged. Registration fees are $6.00 for IRE members and $7.00 for non-members. Cocktails and dinner reservations are an extra $7.50 per person. Checks should be made payable to the 1957 Annual PGMTT Meeting and mailed to K. S. Packard, Airborne Instruments Labs., 160 Old Country Rd., Mineola, L. I., N. Y.

THURSDAY, MAY 9

8:00 a.m. to 9:00 a.m.

Registration.

9:00 a.m. to 12:30 p.m.

SESSION I

Moderator: A. A. Oliner, Microwave Research Institute.

Opening Remarks: H. F. Engelmann, National Chairman, PGMTT.

The State of the Microwave Ferrite Art, B. Lax, M.I.T. Lincoln Labs.

Non-Reciprocal Electromagnetic Wave Propagation in Ionized Gaseous Media, Louis Goldstein, University of Illinois.

Solid State MASER, H. E. D. Scovil, Bell Tel Labs.

12:30 p.m. to 2:00 p.m.

Lunch.

2:00 p.m. to 5:00 p.m.

SESSION II

Moderator: J. H. Rowen, Bell Tel. Labs.

Some Techniques of Microwave Generation and Amplification Using Electron Spin States in Solids, D. I. Bolef and P. F. Chester, Westinghouse Research Laboratories.

Nickel Cobalt Ferrite Line-Widths as a Function of Temperature, J. E. Pippin, Gordon McKay Laboratory, Harvard University.

Ferrimagnetic Garnet Line Widths as Function of Temperature, Wolfe Rodrique and J. E. Pippin, Gordon McKay Laboratory, Harvard University.

Bridging Effects in Resonance Isolator, W. J. Crowe, Bell Telephone Labs.

Exact Solution for a Cylindrical Cavity Resonator Containing a Rod of Gyromagnetic Material, H. E. Bussey and L. A. Steinert, National Bureau of Standards.

Ferrite Directional Coupler, D. C. Stinson, Lockheed Aircraft Corporation, Missile Systems Division.

Reflections in a Ferrite Filled Waveguide, C. B. Sharpe and D. S. Heim, Electronic Defense Group, Engineer Research Inst., University of Michigan.

Resonant Properties of Ring Circuits with Ferrites, F. J. Fischer, Columbus, Ohio.

6:00 p.m.

Cocktails.

7:00 p.m.

Annual PGMTT dinner at which the presentation of the PGMTT award for best paper will be made.

FRIDAY, MAY 10

9:00 a.m. to 12:30 p.m.

SESSION III

Moderator: Ernest Wantuch, AIRTRON.

Longitudinally Magnetized Ferrite Loaded Coaxial Components, H. Seidel, Bell Telephone Labs.

Reciprocal Ferrite Devices in TEM mode Transmission Lines, D. Fleri and B. J. Duncan, Sperry Gyroscope Company.

Non-Reciprocal Ferrite Devices in TEM Mode Transmission Lines, R. Mangiaracina and B. J. Duncan, Sperry Gyroscope.

Non-Mechanical Beam Steering by Scattering from Ferrites, M. S. Wheeler, Westinghouse Electric Corp., Air Arm Division.

An Electronic Scan Using a Ferrite Aperture Luneberg Lens System, D. B. Medved, Convair.

A Microwave Ferrite Frequency Separator, H. Rapaport, RCA, Surface Communications Systems.

Higher-Order Mode Propagation in Ferrite Devices and Wide-Band Tunable Ferrite Microwave Filters, R. F. Soohoo, Cascade Research Corporation.

Ferrite Loaded Circularly Polarized Microwave Cavity Filters, W. L. Whirry and C. F. Nelson, Research Laboratories, Hughes Aircraft.

2:00 p.m. to 3:20 p.m.

SESSION IV

Moderator: Howard Scharfmann, Raytheon Manufacturing Co.

Errors in Measuring Differential Phase Shift, L. M. Silber, Polytechnic Institute of Brooklyn.

Ferrite Loaded Cavity Resonator, G. S. Heller and M. M. Campbell, M.I.T. Lincoln Laboratory.

Resonance Isolator at 1300 MC/SEC, G. S. Heller and G. W. Catuna, M.I.T. Lincoln Lab.

Approximate Solutions for Cavities and Waveguides Containing Ferrites, Walter Hauser, M.I.T. Lincoln Lab.

3:30 p.m. to 5:00 p.m.

SESSION V

ROUND TABLE DISCUSSION: "Design Limitations of Microwave Ferrite Devices."

Moderator: C. L. Hogan, Harvard University.

Panel Members: H. Seidel, Bell Tel. Labs., G. S. Heller, M.I.T., Lincoln Laboratory, R. C. LeCraw, D. O. F. L., J. O. Artman, Harvard University, P. H. Vartanian, Sylvania, H. Carlin, Microwave Research Institute, and D. L. Fresch, Trans-Tech Inc.

Ninth Annual National Conference on Aeronautical Electronics

DAYTON BILTMORE HOTEL, DAYTON, OHIO

MAY 13–15, 1957

Monday morning

May 13

EQUIPMENT APPLICATIONS I

Main Ballroom—Dayton Biltmore Hotel

Moderator: Michael Glass, Hughes Aircraft Company.

Operation of an Electronic Component Parts Application Unit, J. P. Francis, Glenn L. Martin Co.

Establishment and Results of a Comprehensive Component Reliability Program, C. G. Walance, Hughes Aircraft Co.

Analysis of Electronic Parts Application, William Barron, Bell Aircarft Corp.

Component Application Engineering at RCA, R. H. Baker, RCA.

Management Support of Components Applications Organizations, J. H. Allen, Bendix Aviation Corp.

COMPONENT PARTS—VACUUM TUBES

Junior Ballroom—Dayton Biltmore Hotel

Moderator: Walter Knecht, ARDC.

The Wamoscope, a New Micro-Wave Display Device, R. G. E. Hutter and D. E. George, Sylvania Electric Products, Inc.

A 300-Watt Stacked Ceramic Tetrode for Airborne Transmitters, W. B. Foote, Eitel-McCullough, Inc.

Vacuum Tubes for 500° C Envelope Temperature and High Vibration Applications, J. W. Wyman, Bendix Aviation Corp.

The Amplitron, A New Type Micro-Wave Amplifier Tube for High Power, Broad Band Equipment Application, W. C. Brown, Raytheon Mfg. Co.

Half-Tone Display Storage Tube with Magnetic Deflection, M. E. Craig, RCA Tube Division.

C. W. UHF Traveling Wave Power Amplifier of Extended Band Width, Walter Harmon, General Electric Co.

NAVIGATION I

Engineers Club—Auditorium

Moderator: D. G. C. Luck, RCA.

Application of Automatic Dead Reckoning Equipment to Current Problems of Air Navigation, Henry R. Walcott, Eclipse Pioneer Division, Bendix Aviation Corp.

The Nature of Doppler Velocity Measurement, Dr. F. B. Berger, General Precision Lab.

Precision Azimuth Reference Systems for Aerial Navigation, A. J. Shapiro, Kearfoot Co., Inc.

Inertial Navigation Performance Characteristics, Robert W. Wedan, Minneapolis-Honeywell Co.

Doppler Navigation, William J. Tull, General Precision Labs.

Self-Contained Navigation and the Common System, Nathaniel Braverman, C & N Laboratory, WADC.

Afternoon

EQUIPMENT APPLICATIONS II

Main Ballroom—Dayton Biltmore Hotel

Moderator: A. M. Okun, Bell Aircraft Corp.

Evolution of a Coordinate Indexing System for Use in Parts Application, A. J. Chippendale, Convair Division, General Dynamics.

Connector Design and Development to Meet Advanced Application Requirements, T. A. Thompson, Douglas Aircraft Co.

An Airborne Atomic Frequency Standard, J. J. Bagnall and J. H. Holloway, National Co.

A One-Kilowatt Airborne Radio-Frequency Power Amplifier, J. B. Humfeld, Hughes Aircraft Co.

Depot Test Equipment Concepts, D. B. Dobson, RCA.

Some Design Factors for the Magnetron-Isolator Combination, B. A. Krogh, Litton Industries.

COMMUNICATIONS

Junior Ballroom—Dayton Biltmore Hotel

Moderator: L. B. Hallman, WADC.

Some New Horizons in Communications, G. H. Scheer, Jr., WADC.

The Synchronous Detector, R. J. Lutze, General Electric Co.

Kineplex, N. L. Doelz, Collins Radio Co.

LABIL—A Light Aircraft Binary Link, W. F. Walker, Stromberg-Carlson.

Some Aspects of Digital Transmission of Data, Siegfried Reiger, Air Force Cambridge Research Center, Lawrence G. Hanscom Field.

ELECTRONIC EQUIPMENT

Engineers Club—Auditorium

Moderator: To be announced later.

A 90 DB Logarithmic Response Video Pulse Amplifier, C. E. Wilson and A. H. Zefting, Stromberg-Carlson.

Compensation Network Design as Applied to Transistor Feedback Amplifiers, T. E. Smith, Texas Instruments, Inc.

The Design of Transistor Intercommunication Systems for Military Aircraft, G. S. Rambie, Jr., Texas Instruments, Inc.

A High Performance Transistor-Regulated Power Supply, T. A. Weil and B. Erdman, Raytheon Manufacturing Co.

Biased Chokes for Improved Swinging Choke Action, T. A. Weil, Raytheon Manufacturing Co.

The Valentine Antenna, E. M. Turner, WADC.

Tuesday morning

May 14

NAVIGATION II

Engineers Club—Auditorium

Moderator: T. A. Kouchnerkavich, Civil Aeronautics Authority.

Some Aspects of VORTAC, Sven Doddington, Federal Telecommunication Laboratory, Inc.

Radio Position Fixing by Low Frequency Composite Wave Measurement, Ben Alexander, Federal Telecommunication Laboratories, Inc.

DELRAC—A Long-Range Aid to Navigation, D. H. Toller-Bond, Decca Navigator System, Inc.

Light Weight Digital Computers, John Mayer, Weapons Guidance Laboratory, WADC.

Radio Positions Fixing by Low Frequency Groundwave Phase, Winslow Palmer, Sperry Gyroscope Co.

MANAGEMENT I—RESEARCH

Junior Ballroom—Dayton Biltmore Hotel

Moderator: O. H. Winn, General Electric Co.

Planning of Research Work at Westinghouse Electric, Clarence Zener, Westinghouse Research Labs.

Organization of Research Projects, W. O. Bowie, Sylvania Electric Research Labs.

Achieving Teamwork in Research Projects, R. Kompfner, Bell Telephone Labs.

Measurement of Research Results, L. R. Fink, General Electric Co.

The Climate of University Research, C. W. Gartlein, Cornell Univ.

ENVIRONMENT I

Main Ballroom—Dayton Biltmore Hotel

Moderator: Walter Robinson, Consulting Engineer.

Integration of Crew and Equipment Cooling in Supersonic Bomber Design, A. E. Hitsman, Boeing Airplane Co.

Cooling Airflow Control Systems for Airborne Electronic Equipment Designed for Efficient Use of Refrigerated Air, L. H. Schreiber and H. R. Wesson, Convair.

Factors Influencing the Selection of Liquid Rather Than Air for Cooling an Airborne Electronic Component, K. J. Fawcett, Melpar, Inc.

High-Reliability Thermal Design for Commercial Avionics, H. M. Passman, Collins Radio Co.

A Summary of Cooling Design Data for Airborne Electronic Equipment, C. D. Jones, Ohio State Univ.

Afternoon

Engineers Club—Auditorium

FORUM PANEL

Moderator: G. L. Haller, General Manager, Defense Electronics Division, General Electric Co.

Forum Subject: Wanted—New Ideas in Airborne Electronics.

Panelists: J. E. Arnold, M.I.T.; Lt. General T. S. Power, Commander, Air Research and Development Command, USAF; B. D. Thomas, Director, Battelle Memorial Institute; Rear Adm. R. E. Bennett, Chief, Office of Naval Research; H. L. Hoffman, President, Hoffman Radio Corp.; and a representative of a university research organization, as yet unnamed.

Wednesday Morning May 15

AIR SAFETY

Engineers Club—Auditorium

Moderator: Lester Glantz, Bulova Research and Development Laboratories, Inc.

The Challenge of Air Safety, Jerome Lederer, Flight Safety Foundation, Inc.

Safety Provided by the Future Air Traffic Control System, G. C. Dewey, G. C. Dewey & Co., Inc.

Words of Caution Regarding Any Air Safety Program, K. C. Black, Raytheon Manufacturing Co.

Air Force Considerations in Collision Avoidance, Albert Segen, C & N Laboratory, Wright Air Development Center.

Operational Requirements for Data Line, (*Tentative Title*), Brig. Gen. M. W. Arnold, Air Transport Association of America.

Problems in Airborne Communications, (*Tentative Title*), Capt. J. D. Smith, Airline Pilots Association.

COMPONENT PARTS—MISCELLANEOUS

English Room—Dayton Biltmore Hotel

Moderator: H. V. Noble, WADC.

Effects of Nuclear Radiation on Electronic Components and Systems, J. R. Milliron, Electronic Components Laboratory, WADC.

High Temperature (500°) Capacitors, Roger L. Foust, Electronic Components Laboratory, WADC.

Some Considerations in the Measurement of Capacitor Insulation Resistance, F. W. Graham, General Electric Co.

High Temperature Film Resistors, E. M. Griest, Corning Glass Works.

Failure Rate Measurements by Means of Accelerated Tests, I. K. Munson, RCA.

Pulse Transformer, W. A. Ernst Westinghouse Electric Corp.

MANAGEMENT II—DEVELOPMENT

Main Ballroom—Dayton Biltmore Hotel

Moderator: Louis DeRosa, Federal Telecommunications Labs.

Papers in this general area of interest are to be delivered by the following persons: Thomas Meloy, Melpar, Inc.; Robert Crago, International Business Machines; J. E. Boyd, Georgia Institute of Technology; J. F. Byrne, Motorola Research Lab.

Titles for the papers to be delivered by the above author-speakers will be determined later.

ENVIRONMENT II

Junior Ballroom—Dayton Biltmore Hotel

Moderator: J. P. Welsh, Cornell Aeronautical Lab.

Temperature Control Design of the Airborne Bombing and Navigation System An/ASB-4, Beal Marks, J. J. Student, R. M. Dailey and R. W. Hook, International Business Machines Corp.

Improved Airborne Equipment Design with Forced-Air and Liquid Cooling, Walter Robinson, Consulting Engineer.

Heat Exchange System Design for Airborne Electronic Equipment, F. P. Benning and J. P. Jacob, United Aircraft Products, Inc.

Design of Liquid-Filled Containers for High-Voltage Equipment, T. P. Jordan, Sylvania Electric Products, Inc.

Effect of Submerged Liquid Cooling on the Electronic Performance of Several Common Types of Electronic Circuits, K. R. Vincent and G. W. Millsap, Convair.

Afternoon

MANAGEMENT III—PRODUCTION

Main Ballroom—Dayton Biltmore Hotel

Moderator: Werner Auerbacher, Emerson Radio and Phonograph Corp.

Engineering Management of a Production Project, Jack Giles, General Electric Co.

Low Volume Production of a Complex System, Irving Anthony, Emerson Radio and Phonograph Corp.

Reproduction Engineering Starts with Research and Development, Leon Himmel, Federal Telecommunication Labs.

Management Problems in Engineering—Production Relations for a Crash Program, H. K. Hudson, Raytheon Manufacturing Co.

COMPONENT PARTS—SEMICONDUCTORS

Junior Ballroom—Dayton Biltmore Hotel

Moderator: R. M. Ryder, Bell Telephone Labs.

Medium Power Silicon Transistor, R. W. Aldrich and M. Waldner, General Electric Co.

Theoretical Discussion of Radiation Effects on Transistors, J. J. Loferski, RCA Labs.

A Silicon Unijunction Transistor, Stanley Brown, General Electric Co.

High Frequency Germanium Transistors, N. H. Ditrick, RCA.

High Voltage Silicon Rectifiers, J. W. Thornhill and Lt. J. S. LaRue, Texas Instruments, Inc.

A 5 to 10 MC Ten Watt Transistor, Author-speaker from Bell Telephone Labs.

ENVIRONMENT III

Engineers Club—Auditorium

Moderator: F. E. Carroll, United Aircraft Products, Inc.

Cooling and Its Application to Infra-Red Detectors, K. W. Harper, General Electric Co.

Methods of Determining the Thermal Condition of Electron Tubes, J. P. Welsh, Cornell Aeronautical Lab.

A Method of Packaging Transistorized Printed Circuit Board Assemblies for Efficient Use of Refrigerated Air, H. Kamei, North American Aviation, Inc. and A. R. Tice, G-V Controls, Inc.

Expendable-Liquid Cooling of Missile Electronic Equipment, W. W. Hagner, Johns Hopkins Univ.

Forced Air Cooling for Power Transistors, Melvin Mark, Consulting Engineer.

Abstracts of IRE Transactions

The following issues of "Transactions" have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
Aeronautical & Navigational Electronics	Vol. ANE-3, No. 4	$ .85	$1.25	$2.55
Antennas & Propagation	Vol. AP-4, No. 4	2.10	3.15	6.30
Audio	Vol. AU-4, No. 6	.80	1.20	2.40
Automatic Control	PGAC-2	1.95	2.90	5.85
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Information Theory	Vol. IT-2, No. 4	1.85	2.75	5.55
Microwave Theory & Techniques	Vol. MTT-5, No. 1	1.75	2.60	5.25
Reliability & Quality Control	PGRQC-9	2.40	3.60	7.20

*Public libraries and colleges may purchase copies at IRE Member rates.

Aeronautical & Navigational Electronics

Vol. ANE-3, No. 4, December, 1956

Chairman's Report—J. L. Dennis (p. 140)

Integrated Instrument System—C. F. Fragola and C. J. Hecker (p. 141)

Through the years the art of aircraft control has progressed from the use of fundamental human observations that restricted operations to contact conditions, to modern flight instrumentation that permits the precise control of modern high speed, high performance aircraft under very restricted ceiling and visibility weather conditions. The story of improved flight instrumentation in this paper includes the evolutionary trend toward more stringent control requirements which has pointed up the limitations of pure human response and resulted in a philosophy that a system capable of greater accuracy must, in addition to providing a fine degree of instrumentation, take human characteristics into consideration.

Through a discussion of the symbolic and pictorial approaches to providing the ideal flight instrumentation, the present trend toward combining the two approaches is explained and the features of a production system incorporating these modern concepts is described.

A Visual Communication System—V. I. Weihe, B. R. Boymel, and S. C. Feild, Jr. (p. 145)

It has long been realized that manual voice-radio and wire telephone communication facilities cannot meet the future requirements of aviation for air traffic control communication services. The operational requirements for an Air Traffic Control Signaling System (ATCSS) are discussed; the relative applicability of a number of existing and proposed communication systems are shown using an arbitrary but consistent rating method. Block diagrams are presented which cover the laboratory experimental equipment of the Melpar Visual Communication System (VCS). A typical sequence of events and their durations for ATCSS service is given. The main conclusion is that a direct viewed bright storage tube display method should be employed in the ATCSS airborne unit, rather than an electromechanical (or "roller-wheel") type of indicator system.

Azimuth Errors of the TACAN System—De Witt T. Latimer, Jr. (p. 150)

The major advantages and disadvantages of adding the ninth harmonic to the fundamental bearing signal of the TACAN system are presented. The effects of this addition are discussed in terms of the improved system accuracy, particularly under conditions of adverse siting.

TACAN azimuth error data that were collected in ready-to-use form by recently developed methods are presented. These data, which allow separation of the site error from the total system error, indicate the azimuth accuracy of the TACAN equipment.

Pencil and Paper Calculation of Noise Level in Superheterodyne Radar Receivers—D. W. Haney (p. 157)

One method for determination of noise level in radar receivers involves the use of a calibrated signal generator for input voltage and use of a quadratic detector instead of the second detector of the radar receiver.

By a related method, which is described in this paper, the rms noise level may be determined from the second detector output as a function of receiver input for any common type of second detector even though its response is not exactly quadratic.

This method is justified mathematically, and in a practical application the noise level of a 600-mc receiver is determined as 11.2 microvolts (referred to input).

Automatic Testing Is Good Business—L. E. McCabe (p. 161)

Automatic testing can be economically applied in the military electronic equipment business even though it is characterized by small production quantities, complexity of product, and rapid evolution of design changes.

It has been the experience of the Light Military Electronic Equipment Department (LMEED) of General Electric that it is possible to determine an automatic test program of long range significance and still see immediate benefits, if the program is tailored to a sound understanding of the needs of the particular business involved.

This article discusses the approach to automatic testing at LMEED, some of the equipment concepts involved, and the plans to further the program toward the goal of full automation of the test operation.

PGANE News (p. 165)

Contributors (p. 166)

Suggestions to Authors (p. 168)

Cumulative Index to IRE Transactions on Aeronautical and Navigational Electronics—1951–1956 (following p. 168)

Antennas & Propagation

Vol. AP-4, No. 4, October, 1956

News and Views (p. 587)

Circularly-Polarized Biconical Horns—C. Goatley and F. D. Green (p. 592)

A new technique is described for obtaining circular polarization from a biconical horn. This polarization is obtained through use of a spatial array of thin conducting elements between the cone faces; orientation of the elements varies with distance from the feed. The method has been used successfully at S band and X band.

This technique was applied to horns which have maximum radiation intensity in the plane normal to the cone axes as well as to other horns with beams tilted out of this plane. The beam tilt was obtained by an appropriate choice of cone configurations.

Phase Centers of Microwave Antennas—David Carter (p. 597)

This paper is concerned with the location of the phase centers of microwave antennas. The inadequacy of conventional aperture theory for the accurate description of phase centers is discussed. Formulas are developed and, for numerical indications, calculations are made for paraboloidal reflectors of different f/D ratios and a class of primary patterns which provide an approximate representation of a great many common feeds. The results are presented in graphical form to provide useful design information and show the dependence of principal E- and H-plane phase center location on feed and dish parameters. Contrary to the prediction of aperture theory, it is shown that the phase centers of axially symmetric antennas are not in the aperture plane, but they are dispersed about it.

A New Method for the Measurement of the Average Dielectric Constant of the Underground Medium on Site—M. A. H. El-Said (p. 601)

An electromagnetic interference pattern in the far-distance condition is utilized to determine the average dielectric constant of the underground propagational medium. The method depends upon the determination of the surface wave velocity by means of measuring the first self-resonance frequency of a dipole wire laid on the earth's surface.

The Image Method of Beam Shaping—P. T. Hutchison (p. 604)

The image scheme of beam shaping for use at microwave frequencies is described. Cosecant-squared radiation patterns are obtained with parabolic reflectors no larger than those required to give pencil beams of commensurate beam widths. Radiation patterns calculated using diffraction theory are compared with measured patterns of a paraboloidal dish fed by a horn feed and one image horn. Experimental patterns are included to show the effects of variation of all parameters. A qualitative analysis of a paraboloid fed by a horn and several images is shown to agree with measured results.

Loop Antenna Measurements—P. A. Kennedy (p. 610)

Experimental measurements on three loop antenna configurations are presented. The technique for obtaining impedance and current distributions using a single-wire transmission line over an image plane is described with particular attention given to the difficulties encountered. The results are reproduced in graphical form, and for loops where theoretical results are available, curves comparing theory and experiment are presented.

Systematic Errors Caused by the Scanning of Antenna Arrays: Phase Shifters in the Branch Lines—L. A. Kurtz and R. S. Elliott (p. 619)

By choosing a suitable equivalent circuit representation for an array-type scanning antenna with the phase shifters in its branch lines, a general expression is derived for the radiation pattern in terms of the active impedances of the radiating elements, the incremental phase shift between elements, and the desired aperture distribution. If the active impedances of the radiating elements vary with beam position as a result of mutual coupling, or if the active impedances are constant, but different from the characteristic impedances of the branch lines, then an infinite series is required to represent the radiation pattern. The first term of the series is the desired pattern and the remainder can be defined as the systematic error. Individual terms of the series represent beams with relative angular positions which correspond to multiples of the interelement phase shift and with relative magnitudes which are dependent upon the deviation from a match of the active impedance of the radiators. The systematic error, if uncontrolled, can prevent the achievement of low sidelobe level. Experimental evidence is presented in support of the analysis.

A High-Performance Conically-Scanning *X*-Band Antenna of Novel Design—J. G. McCann and R. J. Stegen (p. 628)

A conically-scanning antenna is described which consists of a scanner mechanism, rear feed, and 8-foot diameter paraboloid producing a pencil beam having equal *E*- and *H*-plane beamwidths. The polarization may be quickly changed from linear to circular to vertical polarization. The sense of the polarization is readily changed from left-hand to right-hand. The replacement of one section of transmission line by a half-wave plate changes the unit from a nutating polarization scanner to a twice-scan-speed rotating polarization scanner. The scanner incorporates some novel features such as a symmetrical rear feed having equal *E*- and *H*-plane patterns, high-power quarter-wave and half-wave plates, and an orthogonal mode absorber.

This paper describes the techniques which were employed to obtain the above mentioned performance. The problems encountered and their solutions are described.

Slot Admittance Data at K_a Band—M. G. Chernin (p. 632)

Admittance data on transverse edge slots in RG-96/U waveguide can be obtained by a technique called the moving lossy short technique. By this technique the radiation attenuation of a test section of identical slots can be determined. It is then possible to specify both the slot inclination angle and the slot depth of cut required to yield any conductance within the range of measurements. These data were used to design an experimental 30-slot array with a 25-db Taylor aperture distribution. This array was successful, yielding sidelobes near −23 db. Subsequently, the same data were used in designing an 8-foot array with 432 edge slots having the same aperture distribution. This array had a half-power beamwidth of 14 minutes and sidelobes of the order of −24 db. These results compared favorably with design objectives.

Radiation by Disks and Conical Structures—A. Leitner and C. P. Wells (p. 637)

The Lebedev integral transform is applied to a class of boundary value problems in the theory of diffraction and antennas, including circular disks, apertures and hollow conical structures. It is found that the conventional Wiener-Hopf technique, together with this transform, does not explicitly solve these problems. Instead, one is led to an infinite system of linear equations for the representation of the unknown transform function.

Diffraction of Microwaves by Tandem Slits—L. R. Alldredge (p. 640)

The diffraction of a plane electromagnetic wave by two identical slits in tandem is investigated for normal incidence with the polarization parallel to the edges of the slits.

Theory shows that the scattering cross-section coefficient is proportional to the imaginary part of the far field forward scattering factor. The stationary form of the scattering cross section is developed in terms of the incident field and unknown currents on the edges of the conductors forming the slits. Calculations using the Kirchhoff-type approximation in this stationary form for a tandem slit separation of 0.157λ are in good agreement with the experimental values for slit widths greater than 0.5λ.

Similar calculations for zero tandem slit separation, corresponding to a single slit, and for slit widths greater than 0.3λ are in good agreement with those of the exact theory of Morse and Ruberstein, and as determined experimentally.

The infinitely long slits are approximated experimentally by use of a parallel plate system described earlier by Row. The experimental results show an interesting resonance effect as the slit width changes.

Transmission Characteristics of Inclined Wire Gratings—O. J. Snow (p. 650)

Small diameter parallel wires were imbedded in thin plastic sheets and located closely before an antenna dish receiving plane wave *X*-band energy. Polarization was parallel to the wires, and the grating interval was varied between a fifth-wavelength and a whole wavelength for different panels. Received intensity was measured for varied grating tilt angles about an axis lying parallel to the wires and near the center of each panel. Sharp and intense transmission dips were observed for tilt angles at which the parasitic reradiation maxima lay in or near the end-fire direction. The shapes and angular positions of the transmission vs tilt angle curves are approximated by a tentative theory which assumes that the input impedance of the grating is independent of tilt angle and that the apparently absorbed power is proportional to the areas under plots of antenna array patterns. A more precise theory which includes the effect of varying input impedance was required to predict approximate amplitudes as well as sharp transmission dips of smaller magnitude.

On Resonance in Infinite Gratings of Cylinders—S. N. Karp and J. Radlow (p. 654)

The diffraction by a grating is examined (for spacing large compared to wavelength and dimension of grating element) for wavelengths in the neighborhood of the "Rayleigh" wavelengths. The shape of the elements, and their size in wavelengths is unrestricted. The results, including the effect of interaction, are expressed in terms of quantities relating to single scattering. Some properties of certain determinants formed from single scattered amplitudes are derived. The results are compared with those obtained by other authors, using various restrictions on the parameters.

Line-of-Sight Wave Propagation in a Randomly Inhomogeneous Medium—B. M. Fannin (p. 661)

Theoretical calculations have been made, using the single-scattering approximation, for propagation in a randomly inhomogeneous medium in which the deviations of refractive index from the mean are small. The statistical quantities considered were the variance, correlation function, and power spectrum for the phase and relative amplitude of the field at a point and their differences at two points. The emphasis in this paper is in indicating the transition from the ray treatment results to the scattering cross section results. The correlation function for the refractive index is taken to be time as well as space dependent so that the power spectrum can be computed from the original formulation.

Partially Reflecting Sheet Arrays—Giswalt Von Trentini (p. 666)

Multiple reflections of electromagnetic waves between two planes are studied, and the increase in directivity that results by placing a partially reflecting sheet in front of an antenna with a reflecting screen is investigated at a wavelength of 3.2 cm. The construction and performance of various models of such arrays is discussed. Thus, for example, a "reflex-cavity antenna" with an outer diameter of 1.88λ and an over-all length of only 0.65λ is described which has half-power beamwidths of 34° and 41° in the *E* and *H* planes, respectively, and a gain of approximately 14 db. It is shown that larger systems produce considerably greater directivity but that their efficiency is poor.

Contributors (p. 672)

Index to IRE Transactions on Antennas and Propagation—Volume AP-4, 1956 (following p. 674)

Audio

Vol. AU-4, No. 6, November–December, 1956

PGA News (p. 139)

Terminated Horn Enclosures—W. E. Glenn p. 143)

The characteristics of a finite exponential horn terminated in a physically realizable impedance have been calculated on an IBM 650 computer. Some of the results of the calculations and some experimental results of tests on such a terminated horn are presented.

An Experimental 9000-Watt Airborne Sound System—D. W. Martin, A. Meyer, R. K. Duncan, and E. C. Broxon (p. 146)

An experimental 9000-watt speech announcing system AN/AIC-11(XA-1) was developed for installation in a B-26 aircraft. The system was used for studies of direct communication through the atmosphere to ground personnel from aircraft operating at relatively high altitude. The equipment consisted of a turbine generator type of auxiliary power unit; three 3000-w amplifiers, each driving a separate twin-horn loudspeaker; signal preparation, control, and monitoring units; a loudspeaker mounting frame which rotates the loudspeakers and supports two of the twin horns outboard from the fuselage; and magnetic tape recorders.

On the Phasing of Microphones—B. B. Bauer (p. 155)

Correct phasing of microphones is most important when two or more microphones are connected simultaneously to a single transmission system. The phasing of all gradient and of some phase-shift microphones is reversed for rewardly arriving sound waves. In this paper the phase-frequency characteristics of most common microphones are described; methods for predicting or ascertaining the phasing of microphones are given; and a system is proposed for experimentally de-the resulting horizontal motion is accomplished by simple modification to the horizontal centering circuit of the camera. This device should allow a television station to get much longer life out of certain I. O. tubes and, just as important, to eliminate objectionable burn-in to the television viewer.

The Use of a Mobile Television Monitoring Unit in an Enforcement Program—R. L. Day (p. 19)

The Mobile Television Monitoring Unit operated by the Federal Communications Commission at the present time is being used as an instrument of co-operative enforcement. Discrepancies are called to the attention of the station's technical people, working with them in a positive effort to improve the situation with the mutual goal of providing better TV service for the general public. TV station inspection is combined with technical monitoring observations and cooperative corrective action has usually been taken when discrepancies have been noted.

As a result 86 TV stations have benefited from the activity of this unit during the past year.

The specialized monitoring equipment, mounted in relay racks in a $1\frac{1}{2}$ ton package delivery truck includes interpolation oscillators, electronic counter, black and white and color picture monitors, high quality oscilloscopes and spectrum analyzers.

Measurement of Service Area for Television Broadcasting—R. S. Kirby (p. 23)

It is proposed that the present definition of television service in terms of iso-probability contours be abandoned. A new definition of service area, first proposed by Norton and Gainen in 1950, is recommended in its place. This provides a much more useful measure of service and makes the estimating techniques more tractable.

A method of estimating the service area is described. This method consists of sampling the field strength at specified random locations along circular routes around the transmitter using portable field-strength measuring equipment and an antenna height of thirty feet. The estimate of the service area expressed in square miles is arrived at by a simple integration process based on the probability distributions of field strength levels as a function of distance from the transmitter.

Sawtooth Testing of Audio Amplifiers—R. C. Hitchcock (p. 31)

A sawtooth wave contains all harmonics, and a single oscilloscope picture shows three things about an audio amplifier; flatness of frequency response, transient stability, overload.

A Magnetic Tape Recording System for Video Signals—R. H. Snyder (p. 35)

The magnetic recording of the composite monochrome video signal by a machine which is of size and cost compatible with the other apparatus associated with television broadcasting, providing a generous uninterrupted program length, on a tape reel of practical dimensions, is an accomplished technological fact. Some of the parameters which describe the video signal and the magnetic recording medium are explored in this paper, and the means by which certain happy coincidences have been exploited, are explained. It is stated that the practical operating conditions of the recorder are the result of judicious selections among the compromises imposed by considerations of performance, economy, and the nature of the medium. It is suggested that the development is neither "an order of magnitude extrapolation from recent technology," as it has been described by one observer, nor "merely the next logical extension of the magnetic recording art," as it has been described by another, but is, instead, a nearly ideal example of the process by which creative team engineering, working within well defined limitations, extends the usefulness of existing knowledge.

Emergency Standby Facilities for the Aural Television Transmitter—Benjamin Wolfe (p. 40)

This paper presents a simple method for using the visual carrier of a television broadcast station for the purposes of transmitting both the aural and visual signals. The method is intended for use during emergency operation when the sound portion of the TV transmitter is "off the air" and the visual portion of the transmitter is in working order. This system is not recommended as a substitute for a standby transmitter. While various methods of transmitting the sound on the picture carrier are known, a simple and economical method of multiplexing is desirable for a TV broadcasting station. Consideration was given to a system which would require no adjustment of the main visual transmitter should the normal sound transmitter fail. The signal to noise ratio of the multiplexing system to be described is not as favorable as the conventional method of transmission.

An Economical Guide to Station Planning—D. M. Weise (p. 46)

History of the Directional Antenna in the Standard Broadcast Band for Purpose of Protecting Service Area of Distant Stations—R. M. Wilmotte (p. 51)

IRE Professional Group on Broadcast Transmission Systems Membership Directory (as of 20 November 1956) (p. 56)

Component Parts

Vol. CP-3, No. 3, December, 1956

Rodolfo M. Soria (p. 80)

The Application of Large Capacitors for Use in Energy Storage Banks—D. F. Warner (p. 81)

In this paper, the explosively increasing demand for energy storage capacitor applications is discussed with particular reference to the various special characteristics required and how they affect the cost, mechanical configuration, circuitry, and protection requirements. Of particular importance at this time, there is included a discussion of means of accomplishing lower inductance and "Q" factor improvement over conventional design. A brief discussion of capacitor fundamentals contributing to the control of discharge waves is presented to illustrate the need for precise determination of circuit characteristics and electrical parameters.

Several application problems are discussed, illustrating all of the factors pertinent to the design requirements, and how to obtain them. Since most of these installations represent considerable investment and frequently the need for high reliability, means for protection both against voltage overshoot and isolation in the event of unit failure is also discussed.

Thus, the information presented establishes the criteria for selection of large banks of capacitors under the most favorable economic and technical conditions, along with an assurance of reliable performance without excessive unused safety factor dollars.

Synchro and Resolver Performance Definitions—L. A. Knox (p. 88)

A consistent set of fundamental synchro and resolver definitions is presented with some application notes. The realizable performance of these units in a system is indicated.

These definitions or Proposed Standards have been subdivided into three parts, as follows:

I. Precision resolver definitions,
II. Synchro control transmitter definitions,
III. Synchro performance definitions.

Each subject will be treated regarding definitions, schematic diagrams, errors, and pertinent data.

Factors Affecting Attenuation of Solid Dielectric Coaxial Cables Above 3000 Megacycles—J. R. Hannon (p. 99)

Computations of contributory dielectric, inner conductor stranding and plating, outer conductor coverage and leakage, contact resistance, braid pressure and plating losses, and comparison with laboratory data indicates that the major emphasis to obtain low-loss cables, for use at frequencies above 3 kmc, rests with the outer conductor construction.

Miniaturized High-Altitude, High-Temperature Connectors—C. H. Stuart and R. F. Dorrell (p. 105)

A new series of miniaturized multicontact connectors for high-temperature high-altitude conditions has been developed. The degree of performance achieved represents a considerable advance in the connector art. Both nonshielded and shielded contact designs have been incorporated in these connectors.

Three exceptional features under extreme environmental conditions characterize these connectors: at altitudes in the order of 70,000 feet and temperatures of $-85°$F, the connectors permit operation at 500 volts ac 60 and 400 cps and withstand test voltages of 1000 volts ac 60 cps. Connectors are moisture resistant continuously during a ninety day humidity exposure maintaining a minimum insulation resistance of 200 megohms. Operating temperatures for these connectors are $+500°$F to $-85°$F. Design features and test data under the environmental conditions are presented.

Some Basic Physical Properties of Silicon and How They Relate to Rectifier Design and Application—G. Finn and R. Parsons (p. 110)

The saturation range and the avalanche range of the reverse characteristic of a silicon rectifier and how these regions vary qualitatively with temperature and bulk characteristics of the silicon used are discussed. Also some reasons why these characteristics may vary from theoretical values are given.

The forward current is discussed from the standpoint of the resistive component and conductivity modulation. The effects of temperature, device geometry, and bulk characteristics of both of these components are shown.

In a general fashion, some problems concerning operating life and shelf life of packaged rectifiers are given.

Impregnation of Toroids for High-Temperature Service—E. O. Deimel (p. 113)

Two materials have been used successfully for vacuum impregnation of magnetic amplifiers and similar devices for continuous operation at 325°F or higher. Complete impregnation, even through layers of interwinding tape, has been achieved.

One material is an undiluted silicone rubber capable of withstanding 500°F continuously. The other is a rigid, filled epoxy resin using a nonvolatile hardener, which will withstand 325°F continuously. With proper design, both materials will withstand high accelerations in vibration between $-65°$F and their maximum operating temperature.

Reliable Precision Wirewound Resistor Design—J. S. Galbraith (p. 116)

The precision wirewound resistor has undergone fairly extensive design changes within the last four or five years.

The old "open bobbin" style resistor has been completely replaced by the molded thermosetting plastic enclosed resistor.

The new design is capable of performing at extremely low and high temperatures, and will withstand rugged environmental testing, such as salt water immersion, humidity testing with polarizing voltage, and repeated temperature cycling. Improvements in design and in wire insulation have made it possible to operate these new units at higher internal temperatures, thus increasing their power handling capability.

The protection afforded by the encapsulated housing makes the use of fine wire feasible in military resistors with consequent increase in maximum resistance value.

This paper compares the old and new resistors, and discusses some of the factors which influenced the new precision resistor design.

Multipurpose Evaporated Metal Film Resistors—S. J. Stein (p. 119)

Evaporated metal film resistors have been prepared on ceramic bases which have properties that compare favorably with wirewound resistors. These resistors have been tailor-made for several different types of application. A molded style is available for general purpose applications. A coated version in nine different wattage ratings is aimed at power resistor applications. For high-temperature or high-precision requirements, a special hermetically-sealed variety is undergoing field evaluation. Controlled resistance-temperature coefficients having positive, negative, or near-zero values can be obtained. These resistors have much lower inductance values than conventional wirewound resistors. In addition, they are smaller in size and have a weight advantage particularly for the higher resistance ranges. They offer potential advantages when used for miniaturization or airborne equipment. Performance characteristics for the various types are presented and compared to their wirewound analogs and the existing military specifications.

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Electronic Computers

Vol. EC-5, No. 4, December, 1956

Change of Editorship (p. 183)

A New Type of Ferroelectric Shift Register—J. R. Anderson (p. 184)

Ferroelectric shift registers having completely independent parallel or serial inputs and outputs have been designed and constructed. The principal components of these shift registers are single crystals of barium titanate and silicon junction diodes. Two ferroelectric units and two to three silicon junction diodes are required for each stage of the shift register. Practical operating speeds for 10-stage shift registers with transistor drives are at present from 0 to 5 kc. The small size of the ferroelectric units and the low power consumption in this speed range make the ferroelectric shift register attractive for many digital circuit applications.

Junction Transistor Switching Circuits for High-Speed Digital Computer Applications—G. J. Prom and R. L. Crosby (p. 192)

This paper describes junction transistor switching circuits capable of reliable operation at a clock rate of one megacycle. These circuits, consisting of a flip-flop, a gated pulse amplifier, and diode gates, consume a minimum of power and operate over a temperature range of $-55°C$ to $+85°C$ with complete transistor interchangeability. Applications of these circuits to binary counters, shift registers, and accumulators are also presented.

A Multipurpose Electronic Switch for Analog Computer Simulation and Autocorrelation Applications—N. D. Diamantides (p. 197)

A system of four diodes in a series-parallel connection is combined with dc operational amplifiers in order to accomplish a variety of computational operations. The diode circuit is equivalent to a SPST switch survey or a voltage pulse. When inserted in series with the input of an amplifier or an analog memory, the switch makes possible waveform sampling or waveform quantizing of the input voltage. Other functions, such as the fast discharge of an integrator, are also achieved.

A very significant application of the diode switch in computer circuitry is its use in combination with a multiplier and a storer (a bank of integrators) in order to obtain autocorrelation or cross-correlation of messages after they have been translated into voltages. A commutator or a ring counter is employed to provide the switching pulse. The correlator has the advantage of generating the correlation function concurrently with the message without necessitating previous recording and repeated playback.

Representation of Nonlinear Functions by Means of Operational Amplifiers—R. M. Howe (p. 203)

The representation of a wide variety of nonlinear functions by means of the interconnection of unstabilized operational amplifiers is discussed. The nonlinear functions described include rectification, saturation functions, coulomb friction, dead space, and starting friction. The use of operational amplifiers alone to produce square waves and triangular waves, as well as gating operations, is also discussed. These latter circuits are combined to give a time division multiplier using only standard operational amplifiers as components. Accuracy capabilities for all of these nonlinear operations are the order of 0.01 to 1 per cent.

An Error Analysis of Electronic Analog Computers—V. A. Marsocci (p. 207)

Due to the physical unrealizability of electronic adding and integrating circuits with ideal characteristics, errors will be introduced in the solution of differential equations obtained by the use of electrical analog computers. Numerical errors in the solution will be introduced by fluctuations in the value of plate and of grid supply voltages, changes in the values of circuit components, and changes in the values of the vacuum tube constants. In addition, the limited frequency response of the machine components will cause the computer to solve a characteristic equation of a higher order than the original characteristic equation whose solution is desired. The error in the solution manifests itself as a shifting in the roots of the original characteristic equation as well as the production of some extra roots. The effect of this change in the root position as well as the presence of the extra roots is experienced in the curve of the solution as a function of the independent variable. In a paper on the accuracy of differential analyzers, Macnee has derived an expression which gives the value of the characteristic root shift. The use of this expression is accurate only for certain types of ordinary differential equations.

In this paper a new expression for the value of the root shift is derived. The analysis preceding the new root-shift expression is developed in such a manner as to include the Macnee analysis as a special case.

Pulse Generator and High-Speed Memory Circuit—Z. Bay and N. T. Grisamore (p. 214)

Circuits for the recycling of pulses by means of a driving tube and an electromagnetic delay line have been developed. The necessary characteristic for the driving tube is shown and the effects of the delay line on the amplitude and width of pulses with respect to recycling operation are explained. Two modes of operation of these circuits are possible. One mode of operation allows any number of pulses in a recycling period, the number being limited only by the time space on the delay line. The other mode restricts the number of pulses in a recycling period to a particular value.

Experimental circuits are shown which have been used as generators of pulses as short as 5 millimicroseconds at frequencies as high as 50 mc. Other circuits are shown which can be used as memory circuits for the storage of a number of these short pulses.

The IBM 705 EDPM Memory System—R. E. Merwin (p. 219)

The IBM 705 memory system utilizes magnetic cores both as a storage element and also in a matrix address selection system. The magnetic core has been established as a memory element for large data processing machines. The core compares very favorably with other means of storage with respect to such factors as speed, reliability, size, cost, life, and simplicity of associated electronic circuitry.

The memory consists of a main 20,000 character unit and a 512 character storage unit. Both are three-dimension coincident current systems with the larger containing 35 planes of 4000 cores each and the other consisting of seven planes of 512 cores each. The basic memory cycle is 9 μs long when operating with the input-output units or on internal transfer of data. When operating with the central processing unit a 17-μs cycle is required. Data may be transferred within memory in five character blocks, and the five character instructions are transmitted to the control unit in one-memory cycle. Transfers between memory and the input-output and arithmetic units is serial by character.

Use of the magnetic core matrix switch greatly reduces the electronic equipment required to drive the memory. Simplified circuitry requiring no adjustments eliminates any maintenance time required for making routine adjustments. Indefinite life of the core eliminates any replacement problem of the basic storage element itself.

Reliability of an Air Defense Computing System: Component Development—H. F. Heath, Jr. (p. 224)

This paper presents the general aspects of the component development program for the AN/FSQ-7 Air Defense Computer. The requirements of the system for high reliability and long life necessitate proper selection of the type of component, proper specification of the component to the manufacturer, and proper component application by the circuit design engineer. The component development program has made free use of ideas from the computer industry and the component industry.

Reliability of an Air Defense Computing System: Circuit Design—R. E. Nienburg p. 227)

Extreme reliability resulting in no unscheduled down-time and a low ratio of scheduled maintenance to operate time was the objective of the AN/FSQ-7 design program. The circuit design philosophy of this program is presented. In addition, an approach is given whereby the concept of marginal checking is applied to determine quantitatively 1) the relative reliability of computer circuits and 2) that a margin of safety consistent with the circuit design philosophy existed. An appendix is included setting forth actual examples in a qualitative manner.

Reliability of an Air Defense Computing System: Marginal Checking and Maintenance Programming—M. M. Astrahan and L. R. Walters (p. 233)

Marginal checking by varying supply voltages for some time has been a means of preventive maintenance for electronic systems. Some important innovations have been employed in the marginal checking system of the AN/FSQ-7 air defense computer to give a more effective high-speed preventive maintenance technique. Completely automatic preventive maintenance testing is discussed incorporating program control of the marginal checking system.

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Symposium—The Design of Machines to Simulate the Behavior of the Human Brain (p. 240)

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Information Theory

Vol. IT-2, No. 4, December, 1956

Michael J. Di Toro, Jr. (p. 100)

Applications of Information Theory—M. J. Di Toro, Jr. (p. 101)

On the Shannon Theory of Information Transmission in the Case of Continuous Signals—A. N. Kolmogorov (p. 102)

On Noise Stability of a System with Error-Correcting Codes—V. I. Siforov (p. 109)

The problem posed in this paper is to give a relation connecting the noise stability of a communication system in which error-correcting codes are used to the parameters of these codes. The amount x of code combinations is found, which differ from each other by not less than a given number of elements for a given arrangement of all the possible combinations. It is proved that the quantity x depends on the arrangement of the primary combinations. Inequalities are obtained for the greatest amount of combinations x_m, for which any two differ by not less than d elements for small and large values n of the common number of elements in each code combination. It is established that the probability of distortion, p_n, of a code group in a system with correcting codes satisfies the inequality $p_n<f(p, n, d)$, where p is the probability of distortion of one element. The shape of the $f(p, n, d)$ function is found for large values of n. Graphs of this function are constructed.

Two Inequalities Implied by Unique Decipherability—Brockway McMillan (p. 115)

Consider a list of b words, each word being a string of letters from a given fixed alphabet of a letters. If every string of words drawn from this list, when written out in letters without additional space marks to separate the words, is uniquely decipherable, then

$$a^{-l_1} + a^{-l_2} + \cdots + a^{-l_b} \leq 1, \qquad (1)$$

where l_i, $1 \leq i \leq b$, is the length of the ith word in the list. This result extends a remark of J. L. Doob, who derived the same inequality for lists of a more restricted kind. A consequence of (1) and work of Shannon is that this more restricted kind of list suffices in the search for codes with specified amounts of redundancy.

A Note on the Maximum Flow Through a Network—P. Elias, A. Feinstein, and C. E. Shannon (p. 117)

This note discusses the problem of maximizing the rate of flow from one terminal to another, through a network which consists of a number of branches, each of which has a limited capacity. The main result is a theorem: The maximum possible flow from left to right through a network is equal to the minimum value among all simple cut-sets. This theorem is applied to solve a more general problem, in which a number of input nodes and a number of output nodes are used.

Rectification of Two Signals in Random Noise—L. L. Campbell (p. 119)

The spectrum of the output of a half-wave rectifier is derived for an input which is the sum of random noise and two sinusoidal signals of different frequencies. The method used is the characteristic function method described by Rice. The components of the output spectrum are given as infinite series of hypergeometric functions. If both the input signals are small compared with the noise, it is shown that the ratio of the output signal power at the difference frequency to the output noise power is proportional to the product of the input signal-to-noise power ratios at the two frequencies. If one of the input signals is very large compared with the noise, it is shown that the other signal and the noise are translated in frequency without alteration of the signal-to-noise ratio. A correction factor is obtained for the case where the large signal is not quite large enough. Finally, the output signal-to-noise ratio of a single-sideband detector is calculated as a function of the input signal-to-noise ratio, when the sideband amplitude is one-half the carrier amplitude.

Optimum Detection of Random Signals in Noise, with Application to Scatter-Multipath Communication, I— Robert Price (p. 125)

Solutions are obtained in open form for the optimum, probability-computing detector of either Gaussian signals, or known signals transmitted via scatter-paths, where the signals have been further perturbed by additive white Gaussian noise. The optimum receiver operates on the received waveforms with filter-functions and biasing constants determined by pairs of inhomogeneous and homogeneous integral equations, respectively.

General solution in closed form has not been obtained, but it is possible to draw a few broad conclusions, among them that the filter-functions can be physically realizable. Approximate solution (for the optimum scatter-path receiver) at small signal-to-noise ratios yields a block diagram having interesting implications. For a single-scatter-path, the optimum receiver may be interpreted as the combination of a correlator with an optimum estimator of the Wiener type. Certain special cases in which complete solution is possible have been investigated in detail, and appropriate curves are presented.

The role and performance of the probability-computing detector in an optimum decision-making receiver, for the types of channels considered, is deferred to a companion paper.

A Coincidence Procedure for Signal Detection—Mischa Schwartz (p. 135)

A coincidence method of detecting signal in the presence of noise is compared to the statistically optimum Neyman-Pearson procedure utilizing signal integration and threshold detection. In this coincidence procedure a specified number of the fixed group of successive pulses are required to exceed a voltage threshold level. The analysis is carried out for the case of constant-amplitude signals only and the results indicate that the best possible coincidence method requires about 1.4 db more power than the Neyman-Pearson method.

Some General Aspects of the Sampling Theorem—D. L. Jagerman and L. J. Fogel (p. 139)

The sampling theorem is recognized as an interpolation formula. Starting from the Lagrange Polynomial, this theorem is developed under conditions which are of broader applicability than those usually stated. Such a view point indicates the essential unity of temporal and frequency domain application. It will also be shown that the theorem is applicable as an exact interpolation formula throughout the complex plane. The basic theorem is extended to include sampling of the first derivative of the function. The concept of band-limited functions is introduced through use of Fourier-Stieltjes representations. This is then shown to be subsumed under the general class of functions which is considered in connection with the interpolation theorems developed. This approach, as presented, readily leads to the establishment of many sampling theorems. It is hoped that this paper will aid the formulation of particularly applicable sampling theorems for specific problems.

The Axis-Crossing Intervals of Random Functions—J. A. McFadden (p. 146)

For an arbitrary random process $\xi(t)$ there exists a function $x(t)$ which may be obtained by infinite clipping. The axis crossings of $x(t)$ are identical with those of $\xi(t)$. This paper relates the probability density $P(\tau)$ of axis-crossing intervals to $r(\tau)$, the autocorrelation function of $x(t)$, *i.e.*, the autocorrelation *after clipping*. It is shown that the expected number of zeros per unit time is proportional to $r'(0+)$, *i.e.*, the right-hand derivative of $r(\tau)$ at $\tau=0$. Next a theorem is proved, stating that $P(\tau)=0$ over a finite range $0\leq\tau<T$ if and only if $r(\tau)$ is linear in $|\tau|$ over the corresponding range of $|\tau|$. If $r(\tau)$ is nearly linear for small τ, then the initial behavior of $P(\tau)$ is like $r''(\tau)$. These results are illustrated by some simple, random square-wave models and by a comparison with Rice's results for Gaussian noise.

Determination of Redundancies in a Set of Patterns—Arthur Glovazky (p. 151)

A set of black-and-white patterns can be identified by successive sampling of the individual "cells" which constitute these patterns. If the number of patterns is P and the number of cells is C, it is possible to find, for any specified sampling sequence at least $C-P+1$ cells which may be omitted without obstructing unique identification.

Such redundant cells can be found by two methods: Construction of a "code mobile," and compilation of a "code schedule." The "mobile" is useful inasmuch as its topological characteristics can be correlated with the information capabilities and the inherent redundancy of the given identification process. The "schedule," on the other hand, is the numerical means by which practical cases can be rapidly and systematically solved.

Besides revealing the redundancies in a given set, both the "mobile" and the "schedule" may serve as useful tools in evaluating and designing sampling programs.

Correction (p. 154)

PGIT News (p. 154)

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Microwave Theory & Techniques

Vol. MTT-5, No. 1, January, 1957

J. R. Whinnery (p. 2)

The Next Problem in Engineering Education—J. R. Whinnery (p. 3)

Broad-Band Balanced Duplexers—C. W. Jones (p. 4)

Balanced duplexer circuits are described and a comparison is made between the two principal configurations employing gaseous switching devices. The balanced tr duplexer is limited in power-handling ability, while the balanced atr duplexer has slightly greater received-signal insertion loss. An analysis is made of the reflecting properties of an atr array, and the practical upper limit of the number of array elements is determined.

Calculation of the Parameters of Ridge Waveguides—Tsung-Shan Chen (p. 12)

In this paper an algebraic expression which constitutes an approximation to Cohn's transcendental equation is given for the determination of the dominant-mode cutoff wavelength of ridge waveguides. A modified derivation of Mihran's equation for calculating the characteristic impedance of ridge waveguides is discussed. Based upon these formulas, nomographs are constructed to permit the determination of these parameters with sufficient accuracy when the waveguide and the ridge dimensions vary. Experimental verification of the calculated cutoff wavelength is included.

Excitation of Higher Order Modes in Spherical Cavities—R. N. Ghose (p. 18)

An analysis for determining approximately the optimum position of the exciting source inside a spherical cavity for exciting any TE or TM mode is presented. For any TE or TM mode the orientation of the exciting probe or loop is determined by maximizing the surface integral of $\bar{H}$ or line integral of $\bar{A}$ which is proportional to the excitation coefficient for the corresponding mode. Specific examples of mode discrimination by proper orientation of the exciting source are also included in the paper. Besides, graphs of the surface integral of $\bar{H}$ and the line integral of $\bar{A}$ for various modes are presented to indicate the variation of mutual inductance for any mode, for different positions of the exciting source.

Strip Line Hybrid Junction—H. G. Pascalar (p. 23)

The equivalent circuit of a strip line network is shown to display the properties of a hybrid junction. An application is illustrated by design of a balanced mixer and the presentation of the resultant measured data.

Losses in Dielectric Image Lines—D. D. King and S. P. Schlesinger (p. 31)

The dipole mode in a dielectric rod permits an image system in which half the dielectric and its surrounding field are replaced by a metal sheet. If the field is allowed to extend many wavelengths outside the rod, the resulting line has very low losses. The contribution of the image surface to line loss is calculated, and shown to be generally less than the dielectric loss. Radiation from obstacles along the line is also discussed. Such obstacles in closed single-mode waveguides are useful for matching purposes. Although matching elements are easily constructed for the image line, radiation loss proves difficult to control.

General Synthesis of Quarter-Wave Impedance Transformers—H. J. Riblet (p. 36)

This paper presents the general synthesis of a radio frequency impedance transformer of n quarter-wave steps, given an "insertion loss function" of permissible form. This procedure parallels that of Darlington for lumped constant filters by providing the connection between Collin's canonical form for the insertion loss function and Richards' demonstration that a reactance function may always be realized as a cascade of equal length impedance transformers terminated in either a short or open circuit. In particular, it is shown that insertion loss functions of the form selected by Collin are always realizable with positive characteristic impedances, and that the synthesis procedure, for maximally flat and Tchebycheff performance, involves the solution, at most, of quadratic equations. In addition, this procedure permits the proof of Collin's conjecture that, for his insertion loss function, the resulting reflection coefficients are symmetrical. Finally, closed expressions are given for the coefficients of the input impedance of a given n section transformer in terms of the n characteristic impedances and vice versa.

An Analysis of the Diode Mixer Consisting of Nonlinear Capacitance and Conductance and Ohmic Spreading Resistance—A. C. Macpherson (p. 43)

A method is presented for calculating the mixer admittance matrix Y' which results when an ohmic impedance is connected in series with a diode mixer described by an admittance matrix Y. There are no restrictions on the frequency dependence of the ohmic impedance nor on the number of harmonic sidebands considered. The equations are worked out in detail for the "low Q" case in which signal, image, and intermediate frequencies are considered, and it is shown that Y' in this case is "nearly low Q." As a result of this analysis the usual criterion for good high-frequency mixing, *i.e.*, that the product of the spreading resistance and the barrier capacitance be small compared with unity, is criticized and a new figure of merit is proposed.

Explicit formulas have been derived for calculating the elements of Y' when Y represents the parallel combination of a nonlinear conductance and capacitance. In general, these formulas are cumbersome, but three special cases have been considered in detail.

Case 1: Zero spreading resistance and equal admittances connected to image and signal terminals. Results: a) The conversion gain is independent of the contact area. b) Regions of negative IF conductance are always associated with arbitrarily high gain.

Case 2: High-frequency, small spreading resistance, image shorted across nonlinear conductance and capacitance. Results: a) The conversion loss and the IF admittance can be given by closed equations. b) The IF conductance can be negative. c) Regions of negative IF conductance are bounded by regions of arbitrarily small IF conductance. d) The conversion loss can *decrease* with increasing frequency. e) Low conversion loss is accompanied by narrow bandwidth.

Case 3: The spreading resistance is zero and the image is shorted. Results: a) Above a certain frequency negative IF conductance is obtained and arbitrarily low conversion loss is possible. b) The situation is quite similar to that of Case 1.

Measurements of mixer performance at the "available terminals" are discussed and the failure of the "phenomenological theory of mixing" as a basis for making such measurements is emphasized.

Resonance Properties of Ring Circuits—F. J. Tischer (p. 51)

The ring guide or ring circuit, a microwave device consisting of a waveguide having the ends connected to form an annular ring, has properties similar to those of ordinary resonant cavities. Wave propagation within the ring guide, its interaction with a waveguide to which it is coupled, and its resonant circuit properties are investigated in this report. The properties of a prototype circuit consisting of a ring guide of rectangular cross section were found to agree with theory.

Frequency Stabilization of a Microwave Oscillator with an External Cavity—Irving Goldstein (p. 57)

This paper describes a procedure by which a cavity stabilizer may be designed for a microwave oscillator. Formulas are derived for the following essential design parameters: 1) stabilization factor; 2) stabilization range; 3) vswr of the stabilizer circuit with cavity tuned; 4) vswr of the stabilizer circuit with the cavity detuned; and 5) insertion loss of the cavity assembly.

The validity of designing with the derived relations has been experimentally confirmed.

Cooling of Microwave Crystal Mixers and Antennas—G. C. Messenger (p. 62)

The development of low-noise mixer crystals has reached the point where the noise figure is approaching fundamental, theoretical limits. The desire for still greater sensitivity has led to the consideration of other possible means for noise reduction. This paper will discuss two possibilities: physically cooling the mixer crystal, and using an antenna directed toward background noise which is lower than room temperature. The improvement which can be realized increases rapidly as the room-temperature noise figure is reduced.

Measurement and Control of Microwave Frequencies by Lower Radio Frequencies—R. C. Mackey and W. D. Hershberger (p. 64)

From the fields of nuclear and paramagnetic resonance comes a relation between precession frequency and magnetic field strength for nuclei and unpaired electrons. The relation is such that $f_n = K_n H$ for nuclei and $f_e = K_e H$ for electrons. Thus if the frequency of one oscillator is set for f_n and the frequency of another oscillator is adjusted so that simultaneous nuclear and electronic resonance occurs in the same magnetic field, the frequency ratio of the oscillators is given by the ratio of K_e to K_n. Values of K_e and K_n have been tabulated for many substances and therefore allow frequency comparisons to be made. For example, protons in mineral oil and electrons in hydrogen have a precession frequency ratio of 658.228; hence for an f_e in x band, f_n is about 14 mc when the magnetic field is 3300 Gauss. Changing the value of H causes the frequencies to move up or down the frequency scale but their ratio is always constant. By this method microwave frequencies may be measured with equipment of a much lower frequency range. The precision of measurement is limited by the widths of the nuclear and electronic resonance curves and runs between one part in 10^4 to 10^5. This frequency measurement method may be made the basis of automatic control of microwave frequencies by quartz crystals or very stable lower frequency oscillators. An experimental model of such a system has been constructed and operated.

Discontinuities in a Rectangular Waveguide Partially Filled with Dielectric—C. M. Angulo (p. 68)

The modal spectrum for a rectangular waveguide with a dielectric slab at the bottom of the guide is obtained following the characteristic Green's Function method developed by Marcuvitz. Then a four-terminal network is found as equivalent to the junction of the partially filled waveguide and an empty rectangular waveguide.

An integral equation is written for the electric field at the plane of the junction and variational expressions are derived for the parameters of the four-terminal network connecting the transmission line equivalent to the partially filled waveguide to the transmission line equivalent to the empty guide.

A reasonable guess for the electric field at the discontinuity gives approximate values for the parameters of the four-terminal network. These values agree with experiment.

The parameters of the network are plotted vs frequency and thickness of the slab.

Correspondence (p. 75)

Contributors (p. 75)

Reliability & Quality Control

PGRQC-9, January, 1957

(Papers Presented at WESCON, Los Angeles, Calif., Aug. 21–24, 1956)

Organizing for Reliability—A. M. Okun and J. Cohen (p. 1)

This paper covers only one part of the broad field of reliability. The "what," "why" and "where" of reliability are left to the many other excellent papers. Here, the "how" of reliability is discussed—how to organize to help achieve reliability. This type of organization is not peculiar to the electronics industry nor is it the solution for all reliability problems. It is a good practical method of organization where, by exercising the correct control, the desired result can be achieved.

The Price of Reliability in Airborne Electronic Equipment—A. H. Wulfsberg (p. 9)

The need for increased reliability in both civil and military electronic equipment has recently received much attention and publicity. It appears to be generally accepted that reliability must first of all be designed into the equipment, but it must also be recognized that the design engineer is faced with a host of other design factors such as operating functions, performance, maintainability, cost, size, weight and environmental conditions.

The effect of each of these factors on reliability will be discussed in the hope of promoting better understanding of the problems which face both the equipment user and designer.

Also to be discussed are the actual costs of designing reliable equipment in terms of dollars, time and talent, and factors involved in the motivation of management and engineering personnel to produce reliable equipment design.

An Evaluation of the Cost of Missile Unreliability and the Influence of Field Checkout—A. L. Stanly and J. Tampico (p. 17)

A procedure is presented for determining the optimum amount of field checkout equipment for a guided missile weapons system. The reduction in number of missiles required for a given probability of kill is established for each incremental increase in reliability, taking into account the variables which affect probability of kill at the time of launching. As a result, relationship is derived for the rate of reduction in cost of the weapons system with increasing missile reliability. This savings is then compared with the costs of the various amounts of field checkout equipment which, by rejecting unreliable missiles, results in increasing reliability in flight. The savings from increased flight reliability are then compared with the cost of achieving the increased flight reliability to establish the optimum amount of field checkout equipment.

Calculations of the Risk of Component Applications in Electronic Systems—J. A. Connor (p. 30)

A panoramic insight into the prediction of electronic-system reliability from component characteristics will be given by demonstrating the means whereby simple and economically-practical computations can be made to determine the "risk" factors. A range of specific circuitry and environmental conditions will be chosen as test cases. The statistical adequacy of certain computational procedures will be asserted along with the promotion of a scheme for appraising complete complex systems. The criteria for determining acceptable costs for such reliability evaluations will be described and shown to be compatible with the bounds of practical reliability economics.

Reliability as a Responsibility of Engineering Management—C. J. Savant and H. S. Hansen (p. 45)

Existing reliability literature stresses the necessity for a total systems approach to achieve improvement of reliability in electronic equipment. This is necessary to diminish the adverse *interaction* effects which occur when supposedly reliable subsystems are assembled into systems.

Similar adverse *interaction* effects can occur in organizational units of the people who perform operations required at each stage in the life cycle of electronic equipment.

This paper briefly discusses the responsibilities, the direction and the controls by which management can help to improve equipment reliability by improving the interaction effects between every organizational unit, or combination of units, which contribute to any stage in the growth of electronic equipment.

The Unreliable Universal Component—M. A. Acheson (p. 49)

Perhaps the most universally applied components are in the field of electronics, and especially in the field of electron tubes. The electron tube is often used to serve such a multitude of diverse usages that it is expected to be an universal component. That tubes are often required to meet as many as a dozen sets of unrelated standards is a problem that has developed from the rapid growth of military and industrial electronics. The significance of this situation as it affects reliability is analyzed and discussed in some detail.

Guided Missile Tube Reliability—Alfred Blattel (p. 55)

This paper will deal with the methods employed to achieve the highest possible electrical and mechanical reliability with complete emphasis on quality. The following items will be described: tube design to give (a) maximum mechanical strength under the severest known conditions, (b) to eliminate skill from the mount operation; aspects of parts manufacture and their quality control; new assembly jigs designed to prevent damage to parts (in particular micas) and minimize handling of parts and mounts; methods employed to reduce dust and particles both on parts and finished mounts; a new approach to welding and a new method of using inert gases to prevent oxidation during welding; new cleaning devices designed to improve emission and cleanliness; and sealing-in and exhausting methods for maximum efficiency.

Prediction of Tube Failure Rate Variations—M. P. Feyerherm (p. 65)

The prediction of a reliability figure for an electronic equipment requires that due consideration be given to the various stresses and environments associated with the electron tubes. In order to handle large numbers of tubes, it is necessary to derive certain broad rules and formulas which when applied to "basic" failure rates will give rates representative of the assumed special conditions. In spite of the fragmentary and varied nature of available data, it has been possible to formulate rules and equations which have been demonstrated by experience to yield satisfactory results.

New Testing Concepts for the Advancement of Electro-Mechanical Component Reliability—H. Grumet (p. 72)

With the present needs for ever-increasing reliability of components and systems, we believe that Rototest Laboratories' new cine-radiographic processes present to the design engineer an advanced testing concept for improvement of component reliability.

We feel that application of *X*-ray motion pictures to the fields of electrical, electronic and electro-mechanical design is analogous to the advent of the microscope in the field of medicine. Before the extensive use of the microscope doctors could diagnose a disease by the symptoms, but could not prescribe a cure, since the cause was unknown. Once the microscope revealed the nature of the difficulty, immediate progress was made in preventing and eliminating the various illnesses. In the field of qualification, reliability and quality control testing, we in turn are dealing with "sick" components and in our opinion the proper and extensive application of cine-radiographic techniques will enable us not only to know when a component has failed, but why it failed—thereby permitting the design engineer to remedy the situation in the shortest possible time.

While we do not look upon *X*-ray motion pictures as a complete cure-all for our present reliability problems, we do foresee this new concept adding another dimension to present testing techniques and opening one new road on which the design engineer can travel toward greater reliability.

White-Noise Vibration Test for Electronic Tubes—J. D. Robbins (p. 86)

A "white noise" vibration test for vibration evaluation of electron tubes over a wide range of frequencies is discussed. This test method was developed as a possible solution to two problems encountered in vibration testing, (1) the need for a test which has some relation to the environment experienced by the tube in actual use, and (2) the requirement that the test be completely reproducible with the test equipment being stable and capable of reproduction.

"White noise" vibration is explained theoretically and is compared with other methods. A practical test method is described which has energy distributed equally per octave, covers a bandwidth of 100 to 5000 cps, and has a 15 g peak value. Details are presented on the white noise generator, vibration test equipment, and on the methods of reading the tube noise output. Actual testing specifications for subminiature tubes are explained.

Environmental Effects on Vacuum-Tube Life—H. C. Pleak and A. V. Baldwin (p. 93)

Experimental data are presented in graphical form which will give design engineers a practical basis for estimating tube life under other than normal operating conditions. A discussion of the relative merits of mean life and per cent life is given and the conclusion reached that neither meets the requirements of the equipment designer. A new concept called "time for first failure" is introduced and suggested as a possible substitute when enough data have been collected to understand how it varies by type and by operating condition.

New Filamentary Tubes of High Reliability—Ross Wood (p. 102)

Two filamentary tube types are described. Type CK-6611 is for IF service and type CK-6612 for RF service at 100 megacycles and above. Operating from 1.25 volt A and 30 volt B supplies, unusual operating characteristics are obtained, thus making possible a significant reduction in size and weight of portable equipment.

Design features are discussed which make this performance possible, while at the same time affording a high degree of reliability. Particular attention is given to the factors resulting in long life, freedom from filament burnout on over-voltage, and ability to withstand severe shock and vibration.

A Comprehensive Quality Control Program Designed to Improve Subminiature Tube Reliability—H. Hoyle and H. Davis (p. 105)

The rapidly expanding use of electron tubes in critical military and commercial applications has forced the tube manufacturer to re-examine his manufacturing techniques. These techniques have to be extended and modified to assure conformance with the high degree of performance reliability required. Analysis of field information indicated that failures could be broken into "catastrophic" and "wear-out" type failures. It became apparent that means would have to be developed to eliminate the generation of these types of defects, both actual and potential, at their source, the manufacturing operation. To accomplish this objective an extensive quality control program was designed, organized and placed in operation. It is the purpose of this paper to describe the operation of the plan, to follow its progress and illustrate how the reduction of the actual and potential defects has improved subminiature tube reliability.

A New 300-Watt Stacked Ceramic Tetrode of High Reliability—W. B. Foote (p. 113)

The tube to be described is small (no larger than the 4X150A), rugged, has an integral finned cooler, and can be operated at full ratings to 500 mc. The all metal-ceramic construction permits rigorous processing to provide increased cleanliness and superior outgassing. An unique basing and socketing arrangement plus rigid internal construction allows the tube to withstand 50G of long duration shock in any plane with no support other than afforded by the socket. Noise output under 20G vibration is very low. A history, construction and method of assembly are reviewed.

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.

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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 *The Electronic and Radio Engineer*, incorporating *Wireless Engineer* and included in the March, 1956 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534(47) 658
Recent Research in Ultrasonics and Physical Acoustics in the U.S.S.R.—R. T. Beyer. (*Nuovo Cim.*, vol. 4, Supplement, pp. 31–64; 1956. In English.) A review with 314 references to Russian literature as well as 43 other references.

534.2–8 659
The Absorption of Ultrasonic Waves in a Number of Pure Liquids over the Frequency Range 100 to 200 Mc/s—E. L. Heasell and J. Lamb. (*Proc. Phys. Soc.*, vol. 69, pp. 869–877; September 1, 1956.) Measurement apparatus and results are described. Values of α/f^2 are given for 94 liquids, where α is the absorption coefficient relating to the excess pressure and f is the frequency. The mechanisms responsible for the attenuation are discussed.

524.2–8–14 660
Ultrasonic Relaxation Theory for Liquids—J. H. Andreae and J. Lamb. (*Proc. Phys. Soc.*, vol. 69, pp. 814–822; August 1, 1956.) Analysis is presented expressing the relations between the absorption of the sound per wavelength, the velocity of the sound, and the thermodynamic parameters of the liquid.

534.213.4–8 661
Approximate Formulae for some Frequently Occurring Combinations of Sound Conduits—C. Kleesattel. (*Acustica*, vol. 6, pp. 288–294; 1956. In German.) Combinations of $\lambda/2$ and/or $\lambda/4$ elements for ultrasonic purposes are discussed. Formulas are derived for the forces and velocities at the end faces as functions of impedance, internal friction and tuning. Application of the formulas to electroacoustic transducers is indicated.

534.232:621.395.6:621.372.5 662
Equivalent Quadripole Networks for Electromechanical Transducers: Part 2—A. Lenk. (*Acustica*, vol. 6, pp. 303–316; 1956. In German.) For previous work see 2181 of 1955 (Reichardt and Lenk).

534.232:621.395.6:621.372.5 663
Transducers and their Equivalent Circuits. Application to Microphones—N. Rouche. (*Acustica*, vol. 6, pp. 317–323; 1956. In French.) A classification similar to that of Fischer (*e.g.*, 953 of 1954) is established on a more general basis by examining all the possible forms of the transduction equation compatible with the principle of conservation of energy.

534.232–8:621.318.134 664
Magnetostrictive Ultrasonic Transducers made of Ferrites—Y. Kikuchi, H. Shimizu, and M. Terajima. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, vol. 7, pp. 9–15; June, 1955.) An experimental investigation of underwater transducers for the frequency range 70–80 kc is reported. Three different methods of construction are suggested. Electroacoustic efficiencies of 60–93 per cent are obtained. The investigation of the ferrites for these transducers is described in separate papers by Kikuchi *et al.* (843 below).

534.24+[538.566:535.43 665
Scattering Theorems for Bounded Periodic Structures—Twersky. (See 746.)

534.24–14 666
Reflection of Sound by a Thin Rod in Water—L. M. Lyamshev and S. N. Rudakov. (*C.R. Acad. Sci. U.R.S.S.*, vol. 110, pp. 48–51; September 1, 1956. In Russian.) Experimental results with copper, steel, and aluminium rods, of a thickness small compared with the wavelength in water, indicate that the nonspecular reflections are due to flexural and longitudinal waves in the rod. The critical angle for nonspecular reflection is given by $\sin\theta = c/c_x$, where c is the velocity of sound in the liquid and c_x the velocity of the waves in the rod.

534.4:534.7 667
Various Methods of Representing Sound Spectra—L. Cremer and L. Schreiber. (*Frequenz*, vol. 10, pp. 201–213; July, 1956.) The relative advantages of various methods are discussed, particularly with regard to their suitability for representing line or continuous spectra. For objectively derived spectra, graphs based on filter characteristics are proposed. Following consideration of the distribution of nerve impulses along the basilar membrane, it is recommended that subjective measurements be represented by loudness/frequency-group graphs which alllow for masking effects. See also 1604 of 1956 (Zwicker and Feldtkeller).

534.75 668
The Form of Vibrations of the Impulse- or Noise-Excited Basilar Membrane, as Measured on an Electrical Model of the Inner Ear—H. Bauch. (*Frequenz*, vol. 10, pp. 222–234; July, 1956.) Measurements made using a network consisting of 65 T-sections, based on anatomical data and the calculations of Zwislocki (1052 of 1951), gave results in close agreement with anatomical measurements.

534.836.087.4 669
The Measurement of Noise in the Presence of Level Fluctuations—G. Bobbert and R. Martin. (*Z. Ver. dtsch. Ing.* vol. 98, pp. 997–1002; July 1, 1956.) The apparatus described determines the average noise level over a given period. A coupled recorder and counter mechanism automatically counts and registers the numerical value of the level at successive short time intervals. Its application in the analysis of traffic noise is detailed.

534.84:534.6 670
Proposal for the Definition and Measurement of Intelligibility on a Subjective Basis—H. Niese. (*Hochfrequenztech. u. Elektroakust.*, vol. 65, pp. 4–15; July, 1956.) Difficulties involved in the pulse echo technique proposed by Thiele (311 of 1954) are discussed. In the new method proposed, a determination is again made of the ratio between the useful and disturbing components of the sound, the analysis of the echo oscillogram into the two components being made in accordance with two empirically established time functions depending on the integration effected by the ear and on the subjective perception of noise. Details of suitable test apparatus are given.

534.846.6 671
Some Experiments in a Room and its Acoustic Model—A. F. B. Nickson and R. W. Muncey. (*Acustica*, vol. 6, pp. 295–302; 1956.) Experiments made on a room about $14\times5\times3$ m^3 over an interval of about an octave around 200 cps, and on a quarter-scale model over a correspondingly scaled frequency range, confirm that the accuracy of the model measurements is satisfactory for objective acoustic tests.

ANTENNAS AND TRANSMISSION LINES

621.372.2 672
Dispersive Properties of Multifilar Helices—N. N. Smirnov. (*C.R. Acad. Sci. U.R.S.S.*, vol. 110, pp. 212–215; September 11, 1956. In Russian.) The characteristics of a bifilar helix are analyzed by the method of space harmonics used previously (3276 of 1956) and the extension of the analysis to m-wire helices is outlined. Dispersion characteristics of a bifilar helix are presented graphically for 1) in-phase, and 2) antiphase excitation.

621.372.8 673
Approximate Method of Calculating a Slightly Irregular Waveguide—A. G. Sveshni-

kov. (*C.R. Acad. Sci. U.R.S.S.*, vol. 110, pp. 197–199; September 11, 1956. In Russian.) The solution of Maxwell's equations for propagation in a slightly irregular cylindrical waveguide is obtained in the form $E=E^0+\epsilon_0E^1+\cdots$, with a similar expression for the magnetic vector, where $_x{}^0$ is a function of the shape of the waveguide. The system of coordinates used is similar to that of Jouguet (3786 of 1947).

621.372.8 674
A Clear Representation of Processes in the Propagation of Discontinuous Signals in Waveguides—A. Rubinowicz. (*Z. Angew. Math. Phys.*, vol. 7, pp. 316–325; July 25, 1956.) The field is analyzed in terms of the primary radiation and the components contributed by successive reflections from the walls. Relations between the forerunner and the main wave are elucidated.

621.372.8 675
Waveguide Hybrid Circuits and their Use in Radar Systems—J. W. Sutherland. (*Electronic Engng.*, vol. 28, pp. 464–469; November, 1956.) "The principal types of waveguide hybrid are described and their properties are compared. Balanced duplexers, balanced mixers and waveguide switching, adding, and subtracting circuits are discussed."

621.372.8 676
Symmetrical Dielectric Junctions in Waveguides with Circular Cross-Section for H_{01} Waves—B. Z. Katsenelenbaum. (*Radiotekhnika i Elektronika*, vol. 1, pp. 339–343; March, 1956.) The method of analysis used previously (3485 of 1955) is applied to the case of an H_{01} wave incident on a long transition between two waveguides having equal cross sections but filled with different dielectrics. The general formula for the reflection coefficient is derived.

621.372.8.002.2 677
Waveguide Surface Finish and Attenuation—J. Allison and F. A. Benson. (*Electronic Eng.*, vol. 28, pp. 482–487, 548–550; November and December, 1956.) The surface properties of drawn, mechanically lined, sprayed, cast, electroplated, electropolished, chemically polished and electroformed waveguides are discussed. Attenuation values and roughness factors are tabulated. 36 references.

621.396.67:621.372.54 678
On the Wide-Band Matching of a Dipole Antenna and Yagi Antenna—R. Sato and K. Nagai. (*Sci. Rep. Res. Inst., Tohoku Univ., Ser. B.*, vol. 7, pp. 23–44; June, 1955.) A dipole antenna is shown to be equivalent to a high-pass filter having a certain resistance across its output terminals; it is matched when this terminal resistance coincides with the image impedance of the preceding filter. Application of standard network theory to the problem is illustrated by examples.

621.396.67.029.62 679
Radiotelephony Antennas for the 4-Metre Band—W. Seefried. (*NachrTech.*, vol. 6, pp. 319–325; July, 1956.) A survey of various dipole types for fixed and mobile stations, including a description of a modified folded dipole for use on locomotives.

621.396.674.3 680
The Radiation Field and Impedance of Aerials—K. Fränz. (*Arch. Elekt. Übertragung*, vol. 10, pp. 269–273; July, 1956.) A relation between the inductance and the over-all frequency characteristic of the radiation resistance of dipole antennas of any shape is obtained, and it is proved that the reactive component of radiated power is concentrated in the vicinity of the antenna. The calculation of a long-wave dipole of maximum damping is given, and the impedance of a free antenna is derived as a limiting case of the reactance of an antenna enclosed in a cavity. Formulas are included for the determination of conductor arrangements to satisfy the boundary conditions of given fields.

621.396.677.3 681
Calculating the Efficiency of Antenna Arrays—G. Mather. (*Tele-Tech and Electronic Ind.*, vol. 15, pp. 102–103, 208; June, 1956.) The efficiency is estimated by comparing the rms current of the array with the current of an omnidirectional radiator of equivalent height and power input.

621.396.677.7.012.12 682
Radiation Patterns of a Dielectric-Coated Axially-Slotted Cylinder—R. A. Hurd. (*Canad. J. Phys.*, vol. 34, pp. 638–642; July, 1956.) Expressions for the field are derived theoretically, and azimuthal radiation characteristics are plotted. The dielectric coating enhances the radiation in the shadow region, but does not render the pattern omnidirectional. Variations of field with angle increase with increasing dielectric constant and coating thickness; the positions of the minima appear to be determined only by the diameter of the metal cylinder. Measurements in good agreement with the theory are reported.

621.396.677.8 683
Antenna Reflections—the Reflex Antenna—F. J. Charman. (*R.S.G.B. Bull.*, vol. 32, pp. 59–60; August, 1956.) Short account of the construction and performance of "reflex" antennas of a type described by von Trentini (1310 of 1956), based on multiple reflections between a main reflector sheet and a grating. With a model scaled for 3 kmc the half-power beam widths in the E and H planes were 26° and 30° respectively.

621.396.677.8:621.396.96 684
Antennas for Fire-Control Radar—L. Thourel. (*Ann. Radioélect.*, vol. 11, pp. 216–229; July, 1956.) Discussion indicates that the parabolic reflector is frequently preferable to a lens. Methods of calculating the radiation pattern are derived for the defocused paraboloid; the results obtained agree closely with experimental findings.

621.396.677.83:621.396.11.029.6 685
The Deflection of Short Electromagnetic Waves—Megla. (See 921.)

AUTOMATIC COMPUTERS

681.142 686
A New Computing Method using High-Frequency Currents—H. J. Uffler. (*Ann. Radioélect.*, vol. 11, pp. 187–199; July, 1956.) An electromechanical process is described for performing algebraic operations in analog computers. The system operates at 472 kc and comprises only passive components. Considerable accuracy and stability are achieved.

681.142 687
Applications of a Transformer Analogue Computer—J. R. Barker. (*Brit. J. Appl. Phys.*, vol. 7, pp. 303–307; August, 1956.) Use of the Blackburn analyzer for extracting latent roots of matrices, locating zeros of polynomials, and solving linear and nonlinear simultaneous equations is discussed.

681.142:621.383 688
Photoformer Analysis and Design—E. Elgeskog. (*Chalmers tek. Högsk. Handl.*, No. 172, pp. 40; 1956.) Analysis is presented permitting the design of a photoformer with a bandwidth of several kc, suitable for use in a high-speed repetitive electronic differential analyser. Difficulties due to the short response time are discussed in detail. Results obtained with an experimental system using a plane cr tube screen and photocathode are in good agreement with the theory.

CIRCUITS AND CIRCUIT ELEMENTS

621.314.2:621.372.512.3 689
Design Charts for Tuned Transformers—M. J. Hellstrom. (*Electronics*, vol. 29, pp. 182–186; November, 1956.)

621.316.86:621.372 690
Calculation of Circuits with Indirectly Heated Semiconductor Thermoresistors [thermistors]—N. P. Udalov. (*Avtomatika i Telemekhanika*, vol. 17, pp. 340–342; April, 1956.) The effect of a variation in the heating current on the resistance characteristics may be replaced, in calculations, by an equivalent temperature change. Examples are given.

621.318.57:537.311.33 691
Microwave Semiconductor Switch—M. A. Armistead, E. G. Spencer, and R. D. Hatcher. (PROC. IRE, vol. 44, p. 1875; December, 1956.) The performance of switches comprising crystal diodes arranged inside waveguides is discussed. Useful degrees of isolation can be attained with n-type Ge more readily than with p-type Si, probably on account of the smaller effective mass of the carriers in the former case.

621.318.57:621.314.63 692
Fast Switching with Junction Diodes—Scobey, White, and Salzberg. (See 958.)

621.318.57+621.317.769.029.3]:621.314.7 693
Three New Transistor Circuits—N. Hekimian. (*Electronics*, vol. 29, pp. 178–181; November, 1956.) Descriptions are given of a temperature-stabilized flip-flop, a tone keyer and an af frequency meter using junction transistors.

621.318.57:621.314.7 694
An Asymmetrical Bistable Circuit using Junction Transistors—(*Mullard Tech. Commun.*, vol. 2, pp. 254–278; July, 1956.) Conditions for the stable states of the basic switching circuit are analyzed and an empirical method for investigating the dynamic operation is presented. Detailed procedure for the design of a particular modified circuit is indicated. Reliable switching times of the order of 4 μs may be obtained, with repetition rates up to a fifth of the grounded-base cutoff frequency; trigger sensitivity is good.

621.318.57:621.387 695
The Design of Cold-Cathode-Valve Circuits—J. E. Flood and J. B. Warman. (*Electronic Engng.*, vol. 28, pp. 416–421, 489–493, and 528–532; October–December, 1956.) Switching circuits using cold-cathode diodes and triodes are discussed. Circuit elements are described for performing logical operations, counting, information storage, etc. The effects of tolerances on circuit operation are examined. Applications of multicathode tubes are mentioned. 37 references.

621.372.4 696
Representation of a Type of Lossy Network in Standard Form—B. Gross. (*Arch. Elekt. Übertragung*, vol. 10, pp. 299–302; July, 1956.) The method is limited to two-terminal networks consisting of regular arrangements of two groups of identical sub-systems, with either lumped or distributed elements. A lossy uniform transmission line is treated as an example.

621.372.4 697
Partial Equivalence of Two-Terminal Networks—K. H. R. Weber. (*Hochfrequenztech. u. Elektroakust.*, vol. 65, pp. 1–4; July, 1956.) Analysis is given for networks including R, C, and L. The term "partial equivalence" is used to indicate that, while the equivalence extends over a complete frequency range, it applies only for certain combinations of circuit parameters.

621.372.4:621.314.7 698
Graphical-Analytic Method of Constructing Voltage/Current Characteristics of a Two-Pole Network containing a Semiconductor Triode—N. I. Brodovich. (*Avtomatika i Telemekhanika*,

vol. 17, pp. 335–339; April, 1956.) The construction of the dynamic characteristics of a point-contact transistor in an earthed-base circuit is described.

621.372.413 699
Formula for Calculating the Frequency of a Toroidal [cavity] Resonator—V. A. Teplyakov and B. K. Shembel'. (*Radiotekhnika i Elektronika*, vol. 1, pp. 443–446; April, 1956.) An empirical formula accurate to within 5 per cent is given for re-entrant cylindrical resonators.

621.372.5 700
The Derivation of the Parameters of a Loss-Free Quadripole from the Reactive Transformation Diagram—F. Gemmel. (*Arch. Elekt. Übertragung*, vol. 10, pp. 273–274; July, 1956.) The quadripole parameters are determined from the position of the perspective axis in the reactive transformation diagram. See also 372 of 1957.

621.372.5 701
Nonreciprocal Quadripoles and the Gyrator—E. Cambi. (*Ricerca Sci.*, vol. 26, pp. 2049–2070; July, 1956.) Theory and application of the gyrator [301 of 1951 (Tellegen and Klauss)] are summarized. Natural systems with gyrator properties normally have considerable insertion losses; an approximation to the ideal gyrator can be achieved by introducing active circuit elements.

621.372.54 702
Minimum Signal Distortion and Noise Power in Linear Filters—R. Kulikowski. (*Bull. Acad. Polon. Sci., Classe 4*, vol. 4, pp. 123–126; 1956. In English.) Analysis is presented facilitating the design of physically realizable filters for transferring, with minimum distortion, signals with random and nonrandom components. See aslo 1327 of 1956 (Kulikowski and Plebánski).

621.372.54 703
Bridged-T Filters with One or Two Cut-Off Frequencies—J. E. Colin. (*Câbles and Transm.*, vol. 10, pp. 165–206; July, 1956.) Formulas for the image parameters are derived and the filters are classified according to operational type. Both symmetrical and nonsymmetrical forms are examined and a proof is given of the nonexistence or limitations of special types. Examples are given of symmetrical filters which are more advantageous for high-pass or band-elimination purposes than the very restricted ladder-type and which can be realized as crystal filters. Filter-design tables are appended.

621.372.543.2:538.652:621.396.41 704
Electromechanical Filters for Single-Sideband Applications—D. L. Lundgren. (PROC. IRE, vol. 44, pp. 1744–1749; December, 1956.) The design of filters comprising cylindrical resonators with coupling necks or slugs is discussed; both longitudinal and torsional modes of vibration are considered. From the point of view of production, the preferred frequencies are from 200 to 250 kc. The frequency characteristics of typical filters with various numbers of sections are described; a typical 9-section torsional-mode filter for 250 kc provides a carrier rejection of 27 db.

621.372.553 705
Phase-Adjusting Circuits—J. W. R. Griffiths and J. H. Mole. (*Electronic Radio Eng.*, vol. 34, pp. 26–30; January, 1957.) "A well-known phase-adjusting circuit is shown to be a special form of a more general type of circuit. Various other forms of this generic circuit are discussed, and shown to be of practical use under certain conditions of load, where the original circuit would not be suitable. The results are presented in a form useful for reference."

621.372.56.029.6 706
Wide-Band Coaxial Magnetic Attenuators—G. W. Epprecht. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, vol. 34, pp. 281–285; July 1, 1956.) The matching properties of attenuators of the type described *e.g.*, by Reggia and Beatty (964 of 1953) are improved by arranging the magnetic material, particularly ferrite, as axially spaced disks. The marked frequency dependence of the attenuation is eliminated by combining units in such a way that their frequency characteristics compensate one another.

621.373 707
Self-Oscillations in a System with Delayed Feedback—Yu. M. Az'yan and V. V. Migulin. (*Radiotekhnika i Elektronika*, vol. 1, pp. 418–427; April, 1956.) A system comprising an amplifier and a time-delay feedback circuit is considered, taking into account the dispersion of the cirduit. The predicted effects on the oscillation characteristics of changes in the circuit parameters were confirmed experimentally.

621.373:621.396.822 708
Influence of Slow Fluctuations on an Oscillator—V. I. Tikhonov and I. N. Amiantov. (*Radiotekhnika i Elektronika*, vol. 1, pp. 428–432; April, 1956.) Expressions are derived for the statistical characteristics of the amplitude and phase; the phase fluctuations are calculated.

621.373.4.029.6 709
The Influence of an External Force on Self-Oscillating U.H.F. Systems—E. S. Vorobeichikov and F. M. Klement'ev. (*Radiotekhnika i Elektronika*, vol. 1, pp. 335–338; March, 1956.) Amplitude and stability characteristics of triode and klystron oscillators are derived, taking into account the effect of the finite transit time of electrons.

621.373.421.11 710
A Class of Self-Oscillating Systems—I. M. Volk. (*C.R. Acad. Sci. U.R.S.S.*, vol. 110, pp. 189–192; September 11, 1956. In Russian.) Analysis is presented covering the case of a triode-tube oscillator with two coupled tuned circuits in the grid circuit.

621.373.44:621.317.755 711
A High-Voltage Pulse Generator and Tests on an Improved Deflecting System of a Cold-Cathode Oscillograph—Cones. (See 883.)

621.373.5:621.314.7 712
Transistor Pulse Generator—F. Rozner. (*Electronic Radio Eng.*, vol. 34, pp. 8–10; January, 1957.) The use of *p-n-p* and *n-p-n* transistors in combination to generate pulses with short rise time is discussed. Minority-carrier storage is used to broaden the pulse. The rise and fall times obtainable with l.f. medium-power transistors are of the order of 0.7 μs at a repetition frequency of 100 kc with a peak power output of 1 w.

621.373.52:621.314.7 713
Determination of Steady-State Conditions in Transistor Oscillators—G. Raabe. (*Nachr-Tech.*, vol. 6, pp. 295–302; July, 1956.) Mathematical treatment using Poincaré's method to find the conditions for steady-state oscillations from the circuit parameters and the nonlinear characteristics of the driving device (*e.g.*, point-contact transistor). The effect of parameter changes on the form of oscillation is examined and the frequency deviation and harmonic content are determined. The use of harmonics with frequencies above the transistor cutoff is possible in some circuits.

621.375:51 714
Application of Orthogonal Step Polynomials in the Analysis of Transient Processes in Multistage Amplifiers—S. V. Samsonenko. (*Radiotekhnika i Elektronika*, vol. 1, pp. 269–273; March, 1956.) The application of Hermite and Laguerre polynomials is discussed. The analysis is simpler than by operational-calculus methods, particularly when the number of stages is large.

621.375.2 715
Cascode Characteristics—W. Grant. (*Wireless World*, vol. 63, pp. 33–36; January, 1957.) A graphical method of constructing the characteristics is presented.

621.375.2.018.756 716
Bandwidth/Rise-Time Chart [for design of pulse amplifiers]—M. D. Prince. (*Electronics*, vol. 29, p. 188; November, 1956.)

621.375.2.029.3 717
Audio Amplifier Delivers 3000 W—A. B. Bereskin. (*Electronics*, vol. 29, pp. 162–163; November, 1956.) A push-pull amplifier for deriving a loudspeaker mounted in an aircraft, for direct communication with the ground, uses two Type 4-1000A air-cooled tetrodes. The transformer primary is bifiar wound and is divided into two halves with the secondary sandwiched between them.

621.375.221:621.396.61 718
Linear-Power-Amplifier Design—W. B. Bruene. (PROC. IRE, vol. 44, pp. 1754–1759; December, 1956.) Methods of minimizing distortion and improving reliability in class-AB rf amplifiers for multichannel ssb transmitters are discussed. The significance of the tube characteristics is analyzed. A high-gain, three-stage amplifier using tetrodes is briefly described.

621.375.232.9 719
A Wide-Band Differential Amplifier of Unity Gain—J. C. S. Richards. (*Electronic Eng.*, vol. 28, pp. 499–501; November, 1956.) Description of a circuit with a single-ended output, capable of handling signals as large as 100 V rms over the frequency range 5 cps-500 kc, with a rejection ratio $>$500.

621.375.3 720
Design of Magnetic Amplifiers with Toroidal Cores—O. A. Sedykh. (*Avtomatika i Telemekhanika*, vol. 17, pp. 445–459; May, 1956.) The application of the formulas derived is illustrated by the design of a 50-w amplifier with internal feedback having an amplification factor of 500 and minimum weight.

621.375.3(47) 721
List of Russian and Foreign Literature on Magnetic Amplifiers for 1951–1954—G. B. Subbotina. (*Avtomatika i Telemekhanika*, vol. 17, pp. 471–487; May, 1956.) A bibliography of over 300 references, including 77 to Russian sources.

621.375.4:621.314.7 722
Minimizing Gain Variations with Temperature in *RC*-Coupled Transistor Amplifiers—T. A. Prugh. (PROC. IRE, vol. 44, p. 1880; December, 1956.) A method depending on the appropriate choice of external circuit conductance is discussed briefly.

621.372.5 723
Vierpoltheorie und Frequenztransformation [Book Review]—T. Laurent. Springer, Berlin, 1956, 299 pp., D.M. 34.50. (*Brit. J. Appl. Phys.*, vol. 7, pp. 310–311; August, 1956.) German edition of an authoritative work on network theory originally published in Swedish.

GENERAL PHYSICS

537.21:513 724
Calculation of Capacitance—D. Harrison. (*Electronic Radio Eng.*, vol. 34, pp. 21–25; January, 1957.) "The method of geometrical inversion is applied to the determination of the capacitance between long parallel circular conductors. Formulas are derived for the capacitance between a long cylindrical conduc-

calculated. The expressions obtained for the ordinary and extraordinary components are compared with the corresponding expressions for the case of a dielectric medium. The intensity maxima in one case correspond to zero intensities in the other.

538.566:535.42 745

The Airy Pattern in Systems of High Angular Aperture—B. Richards and E. Wolf. (*Proc. Phys. Soc.*, vol. 69, pp. 854–856; August 1, 1956.) The possibility is discussed that when the angular aperture of the image-forming pencil in an aberration-free optical or microwave system is sufficiently increased, the diffraction image undergoes a more substantial modification than that of a simple diminution predicted by the classical formula of Airy.

538.566:535.43]+534.24 746

Scattering Theorems for Bounded Periodic Structures—V. Twersky. (*J. Appl. Phys.*, vol. 27, pp. 1118–1122; October, 1956.) Analysis is presented for the scattering of a plane wave by periodic arrays of infinite extent in two dimensions and of finite extent in the third. Green's theorem is used.

538.566:537.226 747

On the Reflection of Electromagnetic Waves from a Dielectric Cylinder—H. Wilhelmsson. (*Chalmers Tek. Högsk. Handl.*, pp. 16, 1955.) An exact solution is presented for the general case of incidence of a plane wave at an oblique angle, with either the magnetic or the electric vector perpendicular to the cylinder; the solution for an arbitrarily polarized wave is obtained by superposing the two special solutions. The coupling between the TM and TE modes produced vanishes for normal incidence.

538.569.4.029.6 748

Radiospectroscope for Observing Electronic Paramagnetic Resonance at Centimetre Wavelengths—A. A. Manenkov and A. M. Prokhorov. (*Radiotekhnika i Elektronika*, vol. 1, pp. 469–477; April, 1956.) Two types of equipment are described, one using a crystal detector and af amplifier and the other a superheterodyne circuit with an IF of 75 mc. Block diagrams are shown.

538.569.4.029.6:535.33.08 749

High-Resolution Microwave Zeeman Spectrometer—R. W. R. Hoisington, C. Kellner, and M. J. Pentz. (*Nature, Lond.*, vol. 178, pp. 1111–1112; November 17, 1956.) Brief description of apparatus in which a Q-band klystron, whose reflector is modulated simultaneously by a 25-cps sawtooth voltage and a 100-kc sinusoidal voltage, supplies power at a frequency of 35 kmc to a resonant-cavity absorption cell located between the poles of a large electromagnet; the power reflected from the cavity is detected by a Si crystal.

537.56 750

Physics of Fully Ionized Gases. [Book Review]—L. Spitzer, Jr. Interscience, New York and London, 1956, pp. 105. (*Nature, Lond.*, vol. 178, p. 1083; November 17, 1956.) An accurate and reliable source of theoretical information for researchers in fields combining hydrodynamics and electromagnetism.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523:538.69 751

Suggestions Concerning the Nature of the Cosmic-Ray Cut-Iff at Sunspot Minimum—F. Hoyle. (*Phys. Rev.*, vol. 104, pp. 269–270; October 1, 1956.) A mechanism based on an interstellar magnetic field is discussed. See also 1699 of 1956 (Davis).

523.16:551.510.535 752

The Measurement of Cosmic Radio Emission for Ionospheric Studies—M. Laffineur and J. D. Whitehead. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 347–349; November, 1956.) Technique developed at Meudon for automatically recording cosmic noise over the frequency band 24–24.1 mc is briefly described.

523.3:621.396.9 753

Radio Echoes from the Moon—I. C. Browne, J. V. Evans, J. K. Hargreaves, and W. A. S. Murray. (*Proc. Phys. Soc.*, vol. 69, pp. 901–920; September 1, 1956.) Report of observations made at a frequency of 120 mc, with the moon in transit. The mean echo intensity is compared with theoretical estimates, and the nature of the rapid fading of the echoes is investigated in relation to various laws of scattering from the moon's surface. An investigation was made of slow fading of the echoes in relation to the rotation of the plane of polarization of the radio waves during their passage through the ionosphere. The effect is used to estimate the total electron content of unit cross section between the observer and the moon.

523.5:621.396.11.029.62 754

Meteoric Echoes Observed Simultaneously by Back-Scatter and Forward Scatter—McKinley and McNamara. (See 923.)

523.72:538.566:551.51 755

Solar Temperature and Atmospheric Attenuation in the 7–8-mm Wavelength Range—R. N. Whitehurst and F. H. Mitchell. (PROC. IRE, vol. 44, pp. 1879–1880; December, 1956.) Measurements made using a Dicke-type radiometer (475 of 1947) indicate a solar temperature of 6000°±500°K and values of total vertical attenuation between 0.3 and 0.6 db depending on the weather.

55:621.396 756

Perturbations of a Satellite's Orbit due to the Earth's Oblateness—L. Blitzer, M. Weisfeld, and A. D. Wheelon. (*J. Appl. Phys.*, vol. 27, pp. 1141–1149; October, 1956.) Measurements by conventional radio techniques of the perturbations of a satellite's orbit can be used for a new determination of the earth's oblateness.

550.371+550.386 757

Seasonal Distribution of Short-Period Fluctuations of the Earth's Electromagnetic Field—M. V. Okhotsimskaya. (*Bull. Acad. Sci. U.R.S.S., Sér. Géophys.*, no. 8, pp. 999–1000; August, 1956. In Russian.) The frequency of fluctuations is greatest at the equinoxes. See also 108 of 1955 (Troitskaya).

550.38 758

Vertical Extrapolation of Geomagnetic Field Components—A. Zmuda and L. McClung. (*Trans. Amer. Geophys. Union*, vol. 36, pp. 939–942; December, 1955.) "In the region external to sources of magnetism, both the divergence and the curl of the magnetic intensity are zero. From the corresponding analytic expressions, the derivatives in the vertical direction of each field component are obtained in terms of the values of the components on a surface surrounding the sources. A Taylor series is formed with these derivatives and then used to continue, either upward or downward, the respective components. The rapidity of the convergence depends on the complexity of the surface field and on the distance of the computed point from the surface."

550.385 759

The Average Electric Current System for the Sudden Commencements of Magnetic Storms—J. A. Jacobs and T. Obayashi. (*Geofis. Pura Appl.*, vol. 34, pp. 21–35; 1956. In English.) Report of a statistical investigation of world-wide current systems.

550.385:523.74 760

Relationships between Geomagnetic Micropulsations and Solar *UM* Regions—Y. Kato and S. Akasofu. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 352–354; November, 1956.)

551.510.535 761

Convective Diffusion in the Equatorial F Region—J. W. Dungey. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 304–310; November, 1956.) Phenomena discussed by Johnson and Hulburt (3045 of 1950) are considered further. The usual formula for the conductivity is inappropriate when diffusion is involved. The convective motion is regarded as that of a gravity-driven dynamo and its speed is controlled by the current flowing along the lines of force into lower levels of the ionosphere. The speed is inversely proportional to the east-west scale of the ionization irregularities present, and may be a few m for a scale of 100 m.

551.510.535 762

On the Deviate from the Mean Value of f_0F_2—M. Mambo. (*J. Radio Res. Labs, Japan*, vol. 3, pp. 181–187; July, 1956.) The relative deviation of the midnight value of f_0F_2 from its 27-day running-mean value is determined for the period 1949–1955 for Washington, Tokyo, Huancayo, and Christchurch. Good correlation is found between all these stations; correlation with sunspot numbers is also good.

551.510.535 763

A Universal Formula for the Morning F_2 Ionization at European Stations—O. Burkard. (*Geofis. Pura Appl.*, vol. 34, pp. 207–210; In German. 1956.) The following formula is derived from analysis of observations at 11 stations:—$(f_0F_2)^2 = K(\cos\chi/\cos^2\phi)^x\cos^2\phi$, where χ is the sun's zenith angle, ϕ the geographical latitude, x a parameter depending on locality and K a factor independent of locality.

551.510.535 764

The Occurrence of High Multiple Reflections from the F_2 Region of the Ionosphere based on a Study of the Ahmedabad Records—R. G. Rastogi. (*Proc. Indian Acad. Sci.*, Section A, vol. 41, pp. 253–260; June, 1955.) Observations are reported indicating that even at night the F_2 layer cannot be considered as a simple plane reflecting surface; dynamic changes are taking place most of the time. Photographs and graphical analyses of the records are presented.

551.510.535:523.3 765

The Measurement of the Electron Content of the Ionosphere by the Lunar-Radio-Echo Method—J. V. Evans. (*Proc. Phys. Soc.*, vol. 69, pp. 953–955; September 1, 1956.) Preliminary results are reported of determinations made using the technique indicated by Browne *et al.* (753 above). They indicate that the total electron content of the ionosphere is about twice that expected on the basis of a simple parabolic height distribution of electrons.

551.510.535:621.3.087.4 766

Design and Development of a Simple Ionospheric Equipment—T. V. S. Murty. (*J. Sci. Industr. Res.*, vol. 15A, pp. 70–74; February, 1956.) Medium-power sounding equipment is described. The transmitter oscillator is excited by a separate pulser with variable pulse width. The equipment is manually operated and can be used with simple horizontal receiving dipoles.

551.510.535:621.396.11 767

Investigation of Winds in the Ionosphere by the Spaced-Receiver Method—B. R. Rao, M. S. Rao, and D. S. Murthy. (*J. Sci. Industr. Res.*, vol. 15A, pp. 75–81; February, 1956.) Results of measurements made during 1954 are presented in the form of polar diagrams showing the seasonal variations of wind movements in the E and F regions, and are compared with results obtained by several other workers.

551.510.535:621.396.11 768
Comparison of the Values of (M3000)F_2 at the Four Observatories in Japan—I. Kasuya, K. Sawada, and I. Yamashita. (*J. Radio Res. Labs, Japan*, vol. 3, pp. 161–175; July, 1956.) A statistical analysis for the period 1948–1954 shows that variations of the (M3000)F_2 factor and hpF_2 are closely correlated and are considerably influenced by solar activity; seasonal and diurnal variations are regular functions of latitude.

551.510.535:621.396.11 769
The Focusing of Short Radio Waves Reflected from the Ionosphere—Whitehead. (See 914.)

551.510.535:621.396.11 770
The Absorption of Radio Waves in an Ionospheric Layer—Whitehead. (See 915.)

551.510.535:621.396.11 771
The Connection between Ionospheric Patterns and Field Strengths Reflected on the Ground—Drummond. (See 916.)

551.577/.578:621.396.96 772
Factors Influencing Radar-Echo Intensities in the Melting Layer—R. Wexler and D. Atlas: B. J. Mason. (*Quart. J. R. Met. Soc.*, vol. 82, pp. 349–351; July, 1956.) Comments on a previous paper by Mason (*ibid.*, vol. 81, p. 262, 1955), together with his reply.

551.594.11 773
Short-Period Variations in the Atmospheric Electric Potential Gradient—W. S. Whitlock and J. A. Chalmers. (*Quart. J. R. Met. Soc.*, vol. 82, pp. 325–336; July, 1956.) Measurements of the vertical gradient close to the earth's surface indicate that variations over periods of the order of minutes are generally caused by the horizontal motion of wind-borne space charge.

551.594.2 774
The Vertical Electric Current during Continuous Rain and Snow—J. A. Chalmers. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 311–321; November, 1956.) Investigations of the total current below nimbo-stratus clouds are reported.

551.594.2 775
Visible Electrical Discharges Inside Thunderclouds—D. J. Malan. (*Geofis. Pura Appl.*, vol. 34, pp. 221–223; In English. 1956.) Photographs of lightning flashes in a transparent thundercloud are reproduced. They show that the negatively charged region reaches to an altitude of at least 8 km above ground and has a horizontal width of about 8 km.

551.594.22 776
The Relation between the Number of Strokes, Stroke Intervals and the Total Durations of Lightning Discharges—D. J. Malan. (*Geofis. Pura Appl.*, vol. 34, pp. 224–230; In English. 1956.)

551.594.221:621.396.96 777
The Radar Observation of Lightning—M. G. H. Ligda. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 329–346; November, 1956.) Report of observations made with a horizontally scanning system. Various types of echo are distinguished, and radar displays are reproduced; one of the displays corresponds to a discharge estimated to be >100 miles long.

551.594.5 778
V.H.F. Auroral Noise—T. R. Hartz, G. C. Reid, and E. L. Vogan. (*Canad. J. Phys.*, vol. 34, pp. 728–729; July, 1956.) A typical record is reproduced showing enhanced 32-mc radiation observed at a time when the presence of aurora was indicated by other observations. Repeated observations of enhanced radiation on 32, 50, and 53 mc give support to the view that the phenomenon has an auroral origin.

551.594.5:550.385 779
On the Ring-Current Hypothesis—N. Wax. (*Chalmers Tech. Högsk. Handl.*, no. 171, pp. 32, 1956.) Discussion indicates that experimental proof of the ring-current hypothesis is still lacking and that theories so far advanced are inadequate.

551.594.6:621.396.11.029.45 780
Investigation of the Propagation of Long and Very Long Radio Waves by the Analysis of Atmospherics Waveforms—Al'pert and Borodina. (See 920.)

LOCATION AND AIDS TO NAVIGATION

621.396.93 781
Contribution to the Theory of Goniometers and Coordinate Transformers—K. Baur. (*Frequenz*, vol. 10, pp. 213–221; July, 1956.) Solutions for the magnetostatic boundary conditions and equations for the rotor-stator coupling factor and rotor voltage are derived and used to analyze the operation of a simple goniometer and of a coordinate transformer. The latter is a goniometer with two rotor coils at right angles to provide the two signals for the cro display when a multi-element antenna system is used. The reduction of system errors by a method of harmonic compensation is detailed and examples are given. The whole df system is treated as a network to derive error equations. Errors introduced by the calibration equipment are also examined.

621.396.93.029.62/.63 782
Design of Height-Diversity U.H.F. Direction Finders—J. A. Fantoni and R. C. Benoit, Jr. (*Tele-Tech and Electronic Ind.*, vol. 15, pp. 90–92, 203; June, 1956.) Description of the U.S.A. Type-AN/CRD-6 equipment for the frequency range 225–400 mc.

621.396.96 783
Design of ASDE Radar Equipment—J. E. Woodward and D. R. Kirshner. (*Tele-Tech and Electronic Ind.*, vol. 15, pp. 86–87, 186; June, 1956.) A description of "airport surface detection equipment" for operation at a frequency of 24 kmc, under development for the U. S. Air Force.

621.396.96:551.577/.578 784
Factors Influencing Radar-Echo Intensities in the Melting Layer—Wexler and Atlas: Mason. (See 772.)

621.396.96:621.376.23 785
A Radar Detection Philosophy—W. M. Siebert. (IRE TRANS., vol. IT-2, pp. 204–221; September, 1956.) Discussion of possible meth ods for specifying radar system parameters from the desired performance, in terms of accuracy, freedom from ambiguity, and resolution. Signal energy and waveform are considered separately.

621.396.96:681.142 786
Radar Simulator trains Missile-Master Crews—G. W. Oberle. (*Electronics*, vol. 29, pp. 155–157; November, 1956.) The 30-target simulator described is used to train personnel working with a special anti-aircraft fire-control computer.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 787
Elementary Analogies between Vacuum and Electricity—A. D. Degras. (*Le Vide*, vol. 11, pp. 155–162; July/August, 1956.) The use of equivalent circuits for analyzing complex problems in rarefaction technique is discussed.

533.5:621.3.032.73 788
Equilibrium between Glass and Water Vapour at Bake-Out Temperatures—B. J. Todd. (*J. Appl. Phys.*, vol. 27, pp. 1209–1210; October, 1956.) "The diffusion of water from glass is shown to be a reversible process. The equilibrium partial pressure of water for a soda-lime glass was 10 mm (Hg) at 500° C. and 12 mm (Hg) at 550° C. In a very dry atmosphere the diffusion of water from the glass proceeded as well as in vacuum." For previous work, see 767 of 1956.

533.583:621.385 789
Absorption of Oxygen and Carbon Monoxide by Barium Alloy Getters—R. N. Bloomer. (*Nature, Lond.*, vol. 178, pp. 1000–1001; November 3, 1956.) Results obtained by Wagener (3012 of 1951 and 2685 of 1954) are discussed, and a brief account is given of experiments made to clear up uncertainty regarding the influence of an ionizing discharge on the pumping speed. Only a slight influence was observed in gettering oxygen, the magnitude of the effect being directly proportional to the electron current. The variation of the pumping speed with time and temperature is shown graphically. A catalytic effect is exercised by an incandescent tungsten filament. Results with CO differed from those obtained by Morrison and Zetterstrom (2632 of 1955).

535.215:537.311.33:546.482.21 790
The Photoelectric Properties of Cadmium Sulphide—G. Wlérick. (*Ann. Phys., Paris*, vol. 1, pp. 623–679; July/August, 1956.) Report of a comprehensive investigation. The effect of asymmetrical illumination on the photoconduction in CdS(Cu) was studied using a symmetrical arrangement of Au electrodes; an asymmetrical (rectifying) effect was observed. The properties of the CdS/Au barrier were studied in detail, as were also those of the internal barrier separating the illuminated and nonilluminated regions of the CdS. From these results the relations between the surface and volume properties of the material were established. The observations can be explained satisfactorily in terms of a simple model introducing only donors and traps.

535.215:[546.482.21+546.482.31 791
Photoconductivity Speed of Response for High-Intensity Excitation in Cadmium Sulphide and Selenide—R. H. Bube. (*J. Appl. Phys.*, vol. 27, pp. 1237–1242; October, 1956.) Measurements on single crystals, sintered layers and evaporated layers, using a source at a temperature of 1900° K giving an illumination of 1740 fc., indicated minimum rise times of 250 μs and minimum decay times of 300 μs for CdS with 17 μs and 8 μs as the corresponding figures for CdSe. These figures are for relatively insensitive materials; the variations with sensitivity are discussed. The results can be used to determine trap densities in the materials.

535.215:546.482.21:537.311.33 792
Diffusion Length of Charge Carriers in CdS—J. Auth and E. A. Niekisch. (*Z. Naturf.*, vol. 10a, p. 1035; December, 1955.) A simple method is described for measuring the distribution of charge-carrier concentration in an illuminated photoconducting crystal with a pair of ohmic contacts, and formulas are presented from which the diffusion length can be determined.

535.215:546.817.221:539.232 793
Photoconductivity of Lead Sulphide Films—G. W. Mahlman. (*Phys. Rev.*, vol. 103, pp. 1619–1630; September 15, 1956.) Measurements are reported of the temperature variation of conductivity, the transient response, the dependence on illumination intensity, and the spectral characteristics, for chemically oxidized films.

535.215:546.817.221:539.232 794
Barrier Theory of the Photoconductivity of Lead Sulphide—J. C. Slater. (*Phys. Rev.*, vol.

mechanical stresses affect differently the magnetic properties of various materials thus strengthened or impregnated; some measurements are reported. The use of special resins and methods can eliminate these effects; on the other hand, magnetic characteristics may be improved by the deliberate application of stresses [see also 2586 of 1954 (Williams *et al.*)]

538.221 838
Dynamax—a New Crystal and Domain-Oriented Magnetic Core Material—G. H. Howe. (*Elect. Eng.*, vol. 75, pp. 702–704; August, 1956.) Details are given of the properties of a high-μ Ni-Fe alloy produced in the form of a thin tape.

538.221 839
Magnetic Properties of Electrolytically Precipitated Thin Layers of Nickel—L. Reimer. (*Z. Naturf.*, vol. 10a, pp. 1030–1031; December, 1955.) Brief report of experimental results. The variation of the coercive force with layer thickness is shown graphically for the freshly prepared layer at 20°C and for the layers after treatment at 200°C for two hours.

538.221:538.569.4 840
Multiple Ferromagnetic Resonance in Ferrite Spheres—R. L. White and I. H. Solt, Jr. (*Phys. Rev.*, vol. 104, pp. 56–62; October 1, 1956.) Measurements of microwave absorbtion in Mn- and Mn-Zn-ferrite spheres are reported; five major and seven minor resonance lines were observed over a 700-oersted range of variation of magnetic field. The results are interpreted in terms of complicated modes of precession of the bulk magnetization.

538.221:539.234:538.61 841
Magnification of the Magneto-optical Kerr Rotation by Means of Evaporated Films—J. Kranz. (*Naturwissenschaften*, vol. 43, pp. 370–371; August, 1956.) Technique for observing domain structure at the surface of ferromagnetic bodies is described.

538.221:621.318.13 842
Hysteresis in Magnetically Soft Materials—R. Feldtkeller and H. Wilde. (*Elektrotech. Z., Edn A*, vol. 77, pp. 449–453; July 1, 1956.) Extension of the method of calculation indicated by Preisach (*Z. Phys.*, vol. 94, pp. 277–302; 1935) based on the statistical distribution of the elementary loops. By taking account of the elastic nature of the wall movements and assuming a Gaussian law of distribution the reversible permeability can be calculated [see also 1469 of 1956 (Wilde)] and an explanation found for special or anomalous forms of the hysteresis loop.

538.221:621.318.134 843
Study on Ferrites for Use in Magnetostriction Vibrators: Part 1—Ni-Zn Ferrite. Part 2—Ni-Cu Ferrite—Y. Kikuchi, N. Tsuya, H. Shimizu, M. Terajima, A. Sugiyama, T. Hirone, S. Maeda, and J. Shimoiizaka. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B.*, vol. 7, pp. 1–7 and 171–178; June and December, 1955.)

538.221:621.318.134 844
The Temperature-Dependent Resistivity of certain Iron-Deficient Magnesium Manganese Ferrites—L. C. F. Blackman. (*J. Electronics*, vol. 2, pp. 199–200; September, 1956.) Critical comment on a paper by Osmond (3456 of 1956).

539.23 845
Preparation and Electron-Diffraction Study of Thin Films of Metal Alloys—P. Michel. (*Ann. Phys., Paris*, vol. 1, pp. 719–744; July/August, 1956.)

539.23:537.311.31 846
Resistivity of Very Thin Metal Films—G. Darmois. (*C.R. Acad. Sci., Paris*, vol. 243, pp. 1024–1026; October 8, 1956.) An approximate calculation indicates that the resistivity variations observed by Mostovetch (3031 of 1953) can be explained on the basis of very small differences in the size of the metal grains.

539.32:[546.72+546.621+549.514.51 847
Dynamic Elastic Moduli of Iron, Aluminium and Fused Quartz—D. S. Hughes and C. Maurette. (*J. Appl. Phys.*, vol. 27, pp. 1184–1186; vol. 27. pp. 1184–1186; October, 1956.) Measurements at pressures between 1 and 9000 b and temperatures between 25° and 300°C (200° for quartz) were made using an ultrasonic pulse technique; linear variation of the elastic moduli with both pressure and temperature was observed.

621.315.61 848
X-Ray-Induced Conductivity in Insulating Materials—J. F. Fowler. (*Proc. Roy. Soc. A*, vol. 236, pp. 464–480; September 11, 1956.) "A model based on conduction by free electrons and including the presence of electron traps is proposed, and the theoretical predictions based thereon are shown to be in good agreement with the experimental results. The dependence of induced conductivity and of the subsequent decay upon temperature and dose rate have been investigated. Physical parameters are given for each material: recombination cross section, number of traps and their distribution in energy, mean distance diffused by free electrons and probability factors of release from traps. The results suggest that when crystalline regions are present in a material (*e.g.*, polyethylene), the boundaries of these regions provide trapping sites in addition to traps of unspecified nature which are present in completely amorphous materials."

621.315.61 849
The Effect of Air Inclusions on the Dielectric Strength and Losses of Insulating Materials—Yu. M. Volokobinski. (*Zh. Tekh. Fiz.*, vol. 26, pp. 568–575; March, 1956.) The energy dissipation in an agglomeration of pores is investigated.

621.315.61:621.317.3 850
Investigations on Dielectric Materials for Component Development—P. Henninger, G. Kremmling, and H. Eisenlohr. (*Frequenz*, vol. 10, pp. 241–252 and 286–291; August and September, 1956.) A survey of methods for detecting the effects of moisture, temperature, and mechanical stresses on the performance and aging of solid and liquid dielectrics. To obtain an accurate assessment of material behavior, tests should extend beyond the rating limits of temperature and frequency. The use of optical methods in conjunction with dielectric and magnetic measurements is increasing in importance.

621.315.612:546.28–31 851
Filaments of Silica—B. E. Vassiliou. (*Nature, Lond.*, vol. 178, pp. 1131–1132; November 17, 1956.) Brief note reporting observations of silica filaments of various diameters down to molecular dimensions, formed incidentally in the course of experiments with oxide compounds containing silica. The filaments exhibited strong es charges which decayed slowly.

621.315.612.6:666 852
Dielectric Losses in Boro-alkaline Glasses at Low Temperatures—V. A. Ioffe. (*Zh. Tekh. Fiz.*, vol. 26, pp. 516–525; March, 1956.) Report of measurements over the temperature range 12°–300°K at frequencies of 2.4×10^5 and 10^6 cps.

621.315.615:537.226:621.317.33 853
Dielectric Properties of Polar Solutes in Non-polar Solvents at Microwave Frequencies—D. E. Clarke and S. N. Kumar. (*Brit. J. Appl. Phys.*, vol. 7, pp. 282–284; August, 1956.) Report of measurements on solutions of benzophenone over the temperature range 10°–30°C at a frequency of 9.2×10^9 cps, and of nitromethane over the range 10°–40°C at 3.5×10^{10} cps.

621.315.616 854
Transient Electric Currents from Plastic Insulators—R. J. Munick. (*J. Appl. Phys.*, vol. 27, pp. 1114–1118; October, 1956.) Measurements were made on a number of plastics, at times ranging from 10 to 10^4 sec after a step voltage variation. The results are discussed in relation to possible mechanisms.

MATHEMATICS

512.831:621.372 855
The *N*th Power of a 2×2 Transfer Matrix—D. T. Swift-Hook. (*Electronic Eng.*, vol. 28, p. 505; November, 1956.) Relations useful in the analysis of iterated networks are discussed.

513:537.21 856
Calculation of Capacitance—Harrison. (See 724.)

517:519.2 857
Some Discontinuous Stochastic Processes—A. Dalcher. (*Z. Angew. Math. Phys.*, vol. 7, pp. 273–304; July 25, 1956.) Methods are discussed for determining the distribution of a function x at time t, where x is subjected to random discontinuous variations and $x(t)$ is a solution to the differential equation $dx/dt = x(x, t)$ between the discontinuities.

517:535.42 858
Some Definite Integrals involving Conical Functions—L. B. Felsen. (*J. Math. Phys.*, vol. 35, pp. 177–178; July, 1956.) A method is presented for evaluating integrals occurring in the solution of diffraction problems.

MEASUREMENTS AND TEST GEAR

529.786 859
Stark-Modulation Atomic Clock—I. Takahashi, T. Ogawa, M. Yamano, A. Hirai, and M. Takeyama. (*Rev. Sci. Instr.*, vol. 27, pp. 739–745; September, 1956.) Details are given of an instrument in which particular attention has been paid to long-term stability.

621.3.018.41(083.74):621.314.7 860
A Transistor-Driven Tuning-Fork Frequency Standard—F. Haas. (*Toute la Radio*, vol. 23, pp. 282–283; September, 1956.) The economical circuit described includes a junction transistor and is suitable for a 1.5-V battery supply.

621.314.7.001.4 861
A Transistor Tester—(*Mullard Tech. Commun.*, vol. 2, pp. 248–253; July, 1956.) The instrument described permits the dc determination of current gain, collector leakage current and collector turnover voltage for grounded-emitter *p-n-p*-junction transistors, to an accuracy within about 5 per cent.

621.317.3:621.315.61 862
Investigations on Dielectric Materials for Component Development—Henninger, Kremmling, and Eisenlohr. (See 850.)

621.317.3:621.372.41 863
A New Method for Precise Measurement of Quadrantal Frequency-Difference [of resonators] by applying Carrier-Suppressed Modulation—Y. Kikuchi, H. Shimizu, and M. Terajima. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, vol. 7, pp. 17–21; June, 1955.) The carrier frequency is made equal to the natural frequency of the resonator under test; the frequency of the modulating signal is adjusted so so that the voltage across the resonator is $1/\sqrt{2}$ of that produced by the same current in a resistance equal to that of the resonator at

Post-deflection acceleration is not used. The timebase is designed to provide for writing speeds of $10^4-5\times10^2$ km; it can be synchronized with each pulse, or it can be triggered by a single pulse to give a single sweep or 50 sweeps per sec. Complete circuit diagrams are given. The application of the oscilloscope in the study of pulsed magnetron oscillations is described and illustrated by oscillograms.

621.317.769.029.3+621.318.57]:621.314.7 886
Three New Transistor Circuits—Hekimian. (See 693.)

621.317.78.029.6:621.316.825 887
A Microwave Thermistor Calorimeter—M. J. Smith and J. R. M. Vaughan. (*J. Sci. Instr.*, vol. 33, pp. 353–356; September, 1956.) A continuous-flow instrument is described, for measuring powers down to a few mw. The circuit is a direct-reading four-thermistor dc bridge indicating degrees of temperature rise and watts at a standard rate of water flow on a specially calibrated microammeter. Mathematical analysis is presented covering compensation for variations of ambient temperature and correction for internal heat losses. The error is probably $<\pm10$ per cent for powers >1 w at 8–9 mm λ with a reading time of a few seconds.

621.317.794.029.6 888
A Semiconducting Antimony Bolometer—E. J. Gillham. (*J. Sci. Instr.*, vol. 33, pp. 338–341; September, 1956.) Description of an instrument using a thin film of "amorphous" Sb sputtered on to a plastic pellicle; the resulting resistance of the film is suitable for matching into a tube amplifier.

621.389:539.155.082.7 889
Non-magnetic Mass Spectrometers—L. W. Kerr. (*J. Electronics*, vol. 2, pp. 179–198; September, 1956.) A general theory of rf "energy-gain" mass spectrometers is presented, based on a Fourier analysis of the rf analyzer field. The performance of instruments in this class is compared with that of magnetic mass spectrometers.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.232–8:538.652 890
Magnetostrictive Transducers with Mechanical Loads—R. R. Whymark. (*Acustica*, vol. 6, pp. 277–287; 1956.) The influence of loads comprising thin stubs on a window-type transducer is considered theoretically and checked by measurements. Liquid loads are simulated with high-loss structures. An optimum value of 42 per cent is observed for the electromechanical efficiency; this is in good agreement with the value predicted from theory.

550.8:621.387.4 891
An Airborne Computer-Controlled Detector for Radioactive Ores—E. J. Frank. (*J. Brit. IRE.*, vol. 16, pp. 633–645; Correction, p. 621. November, 1956.)

621–52:621.395.625.3 892
Delay Unit using Magnetic Recording—V. A. Ivanov. (*Avtomatika i Telemekhanika*, vol. 17, pp. 324–328; April, 1956.) A variable-speed recorder used in conjunction with a set of magnetic-tape loops of various lengths can be used for producing delays between 0.5 s and 20 min in signals of frequencies up to 10 cps for purposes of automatic control. The signal is recorded using a carrier frequency of 820 cps with AM, or, in the case of a dc signal, fm.

621–52:621.9 893
Machine Tool Control—C. K. Marklew. (*Elect. Rev., Lond.*, vol. 159, pp. 189–193; August 3, 1956.) A brief survey of electronic control systems using punched tape, cinematograph film, magnetic tape, photocell devices, etc.

621.317.39:531.71:538.63:537.311.33 894
A Magnetoresistance Displacement Gauge—I. M. Ross and E. W. Saker. (*Nature, Lond.*, vol. 178, p. 1196; November 2, 1956.) The stylus displacement to be measured causes an InSb crystal to move relative to a permanent magnet whose poles are shaped to provide a strong field gradient; in a practical arrangement, two InSb crystals are used in a bridge. A galvanometer deflection of 5 mm has been obtained for a stylus movement of 1μ.

621.317.79:531.7 895
Transducer Characteristics—H. G. M. Spratt. (*Electronic Radio Eng.*, vol. 34, pp. 2–8; January, 1957.) "The principles of transducers employed for the measurement of the vibration of and strain in mechanical bodies are explained and some representative types are described."

621.384.6 896
Fixed-Field Alternating-Gradient Particle Accelerators—K. R. Symon, D. W. Kerst, L. W. Jones, L. J. Laslett, and K. M. Terwilliger. (*Phys. Rev.*, vol. 103, pp. 1837–1859; September 15, 1956.) Radial-sector and spiral-sector types of fixed-field alternating-gradient accelerators are described. The former are simpler to construct; the latter occupy a smaller volume for a given particle energy. Analysis for the orbits is presented. The principles discussed have applications in the design of fixed-field synchrotrons, betatrons, and high-energy cyclotrons.

621.384.6 897
On Exceeding the Critical Energy in a Strong-Focusing Accelerator—A. A. Kolomenski and L. L. Sabsovich. (*Zh. Tekh. Fiz.*, vol. 26, pp. 576–584; March, 1956.) When the energy of particles in an accelerator reaches a certain critical value the normal operating conditions are disturbed. Under certain conditions, a transition through this value is possible without affecting the operation of the accelerator.

621.384.612 898
Radiation Resonance in Synchrotrons—A. N. Matveev. (*Zh. Eksp. Teor. Fiz.*, vol. 30, p. 804; April, 1956.) The resonance due to radiation results in an increase in the amplitude of betatron oscillations.

621.384.622.1 899
Ion-Beam Focusing in a 200-kv [linear] Accelerator—D. Kamke and H. Seguin. (*Z. Naturf.*, vol. 10a, pp. 1036–1038; December, 1955.) Results of calculations and measurements are compared.

621.385.833 900
Focal Properties and Chromatic and Spherical Aberrations of the Three-Electrode Electron Lens—G. D. Archard. (*Brit. J. Appl. Phys.*, vol. 7, pp. 330–332; September, 1956.) "Published theoretical and experimental values of the focal lengths of simple three-electrode electron lenses are compared and found to be in general agreement. These are presented in the form of graphs which show directly the dependence of focal length on lens geometry and voltage ratio. Analogous graphs are derived for spherical and chromatic aberrations in the form of the radios C_S/S and C_C/S (S being the separation of adjacent electrodes), and the general relation between the various curves is discussed."

621.385.833 901
Study of the Leakage Fields of a Four-Pole Magnetic Lens—A. Septier. (*C.R. Acad. Sci., Paris*, vol. 243, pp. 1026–1029; October 8, 1956.)

621.385.833 902
Use of a Four-Pole Magnetic Lens to reduce Image Distortion in Reflection Electron Microscopy—C. Fert and R. Saporte. (*C.R. Acad. Sci., Paris*, vol. 243, pp. 1107–1110; October 15, 1956.)

621.385.833 903
Study of the Effective Length of a Four-Pole Magnetic Lens and of its Variations over the Gap—A. Septier. (*C.R. Acad. Sci., Paris*, vol. 243, pp. 1297–1300; October 29, 1956.)

621.385.833 904
Improvement of the Resolving Power of the Emission-Type Electron Microscope—C. Fert and R. Simon. (*C.R. Acad. Sci. Paris*, vol. 243, pp. 1300–1303; October 29, 1956.) Description of an instrument with which a resolving power of 300 Å has been attained. The cathode emission is produced by ion bombardment, and a beam-limiting diaphragm is used.

621.385.833 905
Scanning Electrometer for Electron Microscopy—G. F. Bahr, L. Carlsson, and G. Lomakka. (*Rev. Sci. Instru.*, vol. 29, pp. 749–750; September, 1956.) An arrangement for direct measurement of electron intensities in the image plane of an electron microscope comprises probe with pre-amplifier, mechanical drive, amplifier, and recording device.

621.395.625.3 906
Magnetic Head has Megacycle Range—O. Kornei. (*Electronics*, vol. 29, pp. 172–174; November, 1956.) A range of recording heads for high frequencies have ferrite cores with the gap defined by means of thin pole shoes made of 16-alfenol, with coatings of silicon monoxide. Recording at pulse densities up to 2500/in. is feasible.

621.398:629.13 907
"Jindivik"—Radio-Controlled Aircraft—E. W. Baynton, B. S. Deegan and R. W. Leslie. (*Proc. IRE, Aust.*, vol. 17, pp. 267–277; August, 1956.) "The Jindivik is a jet powered target aircraft, which is controlled by radio either from a ground station or from a shepherd aircraft. The switching method of control is used; the control signals transmitted over the radio link specify the required flight manoeuvres which are carried out under the control of the automatic pilot system. An fm/am telemetry system based on inductance-type transducers transmits flight data back to the ground controller. Twin-track magnetic tape equipment records the control and telemetry signals throughout each trial for subsequent study."

681.84:621.37/.38 908
Punched-Card Reader for the Blind—F. Dado, V. Proscia, and M. Raphael. (*Electronics*, vol. 29, pp. 148–149; November, 1956.) Brushes attached to a slider move in parallel lines over the punched card, closing circuits through the performations so as to produce coded tones. A linear braille scale is provided on the card.

621.38 909
Static and Dynamic Electron Optics. [Book Review]—P. A. Sturrock. University Press, Cambridge, 1955, 30 s. (*Proc. Phys. Soc.*, vol. 69, p. 962; September 1, 1956.) "This book will be indispensable to all those who want to do original work in electron optics or on particle accelerators. . . . "

PROPAGATION OF WAVES

621.396.11 910
The Calculation of Radio [-wave] Refraction—D. M. Vysokovski. (*Radiotekhnika i Elektronika*, vol. 1, pp. 274–276; March, 1956. Expressions for the angle of refraction in the form of an integral and of a power series are

briefly considered. Conditions are formulated for the refraction to be independent of the height distribution of the refractive index.

621.396.11 911
Influence of the Height Distribution of the Permittivity of Air on the Refraction of Radio Waves in the Lower Layers of the Atmosphere—A. V. Shabel'nikov. (*Radiotekhnika i Elektronika*, vol. 1, pp. 277–280; March, 1956.) Results of calculations indicate that the angle of refraction is practically independent of the function $\epsilon(h)$ for angles of elevation $\gtrsim 10°$ on arrival.

621.396.11 912
Comparative Performances of High-Frequency Radio-Telegraph Circuits during Disturbed Conditions—R. J. Hitchcock. (*Electronic Eng.*, vol. 28, pp. 476–481; November, 1956.) The performances of the Melbourne-London and Nairobi-London circuits for the sunspot-minimum period September, 1952 to January, 1955 were analyzed by the superposed-epoch method, the performance of the New York-London circuit being taken as the basis for defining disturbed ionospheric conditions. Curves are also shown for the Montreal-London and Capetown-London circuits for the sunspot-maximum period July, 1946 to October, 1949. Examination of the data indicates that while high-latitude routes deteriorate during disturbed periods, the performance of other routes may be unaffected or may even improve.

621.396.11:550.385:523.7 913
On Radio Propagation Disturbances—K. S[h]inno. (*J. Radio Res. Labs., Japan*, vol. 3, pp. 155–160; July, 1956.) Propagation disturbances associated with solar M regions, having a 27-day recurrence period, are more severe than those associated with nonrecurrent magnetic storms. Forecasts of disturbance are more accurate during the decreasing half-cycle of solar activity.

621.396.11:551.510.535 914
The Focusing of Short Radio Waves Reflected from the Ionosphere—J. D. Whitehead. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 269–275; November, 1956.) "It is shown that short-lived increases in the mean amplitude of waves reflected from the F region of the ionosphere are the result of focusing effects caused by large-scale distortions moving horizontally. The velocity of the movement can be determined from simultaneous observations of changes in the amplitude and the phase. It proves to be of the same order as that found at the same time for the small irregularities. There is a marked diurnal variation in the number of amplitude increases caused by focusing."

621.396.11:551.510.535 915
The Absorption of Radio Waves in an Ionospheric Layer—J. D. Whitehead. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 276–281; November, 1956.) "The absorption of radio waves traveling vertically through or reflected in a Chapman layer is investigated by a method which takes into account the presence of the earth's magnetic field. The distribution of the absorption along the path is considered, and it is shown that when the wave is reflected inside the layer an important contribution to the absorption occurs near the level of reflection."

621.396.11:551.510.535 916
The Connection between Ionospheric Patterns and Field Strengths Reflected on the Ground—J. E. Drummond. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 282–294; November, 1956.) "If the ionosphere is regarded as a plane, patchy reflector, it can be shown by using Doppler shift theory and wave theory that drift and turbulent processes with periods less than $\lambda/2 \sin \alpha$ (λ is the radio wavelength and 2α is the angle subtended by the reflecting area on the ground) do not produce patterns on the ground and other short-length processes are attenuated. The correlogram of the reflected signal is also correspondingly modified, and some ionospheric observations are examined in the light of this theory."

621.396.11:551.510.535 917
The Calculation of Group Velocity in Magnetoionic Theory—R. F. Mullaly. (*J. Atmos. Terr. Phys.*, vol. 9, pp. 322–325; November, 1956.) "It is shown how the magneto-ionic group refractive index μ' may be calculated as a function of the direction of propagation θ by expressing both these quantities in simple form in terms of a parameter λ, which is given a series of values. A similar method gives μ' as a function of the electron density for a fixed value of θ. Whereas most computations of μ' made up to the present have required electronic calculating machines, the simplified formulas given here are suitable for use with a desk calculator. Throughout, the effect of collisions in the medium is neglected."

621.396.11:551.510.535 918
On the Degree of Suitability of Ionospheric Prediction—H. Shibata, Y. Arima, and T. Oguchi. (*J. Radio Res. Labs, Japan*, vol. 3, pp. 177–180; July, 1956.) A statistical method for comparing quantitatively the predicted values of f_0F_2 with observed values is described. An example taken from figures for Tokyo shows satisfactory results.

621.396.11.029.45 919
Calculation of the Field Strength of Long and Very Long Radio Waves above the Earth's Surface in Practical Conditions—Ya. L. Al'pert. (*Radiotekhnika i Elektronika*, vol. 1, pp. 281–292; March, 1956.) Results are presented of calculations of propagation at frequencies of 500 cps-30 kc taking into account the inhomogeneity of the ionosphere and the frequency dependence of the conductivity. Analysis of the wave interference factor $Ae^{i\phi}$ shows that A and ϕ depend on distance and frequency in a complex and irregular way; the various functions involved in the calculations are tabulated and graphs are shown of the modulus of the interference factor and the differential and mean phase velocities, all as functions of distance. Graphs are also shown of the field strength at various distances as a function of frequency. The calculated results are in fair agreement with published experimental results.

621.396.11.029.45:551.594.6 920
Investigation of the Propagation of Long and Very Long Radio Waves by the Analysis of Atmospherics Waveforms—Ya. L. Al'pert and S. V. Borodina. (*Radiotekhnika i Elektronika*, vol. 1, pp. 293–308; March, 1956.) The method described involves the harmonic analysis of oscillograms of single lightning discharges and the goniometric determination of their origin. The equipment used is briefly described and block diagrams are given. The results obtained for the dependence of the field strength and mean phase velocity on the frequency and distance are, in general, in good agreement with calculated results.

621.396.11.029.6:621.396.677.83 921
The Deflection of Short Electromagnetic Waves—G. Megla. (*Hochfrequenztech. u. Elektroakust.*, vol. 65, pp. 15–36; July, 1956.) Control of field-strength at points beyond the horizon by using diffraction and refraction effects and deflecting reflecting antennas is discussed. Various shapes and combinations of reflectors are examined. Quantitative assessments are made of the limits within which these effects are usable. Measurements made using wavelengths of 6.2, 3.2 and 1.5 m and 20 and 10 cm are reported.

621.396.11.029.62 922
Taking Account of Antenna Height in the Theory of Tropospheric Scattering of Metre-Wavelength Radio Waves—O. I. Yakovlev. (*Radiotekhnika i Elektronika*, vol. 1, pp. 309–313; March, 1956.) A development of the work of Booker and Gordon (1757 of 1950) and Gordon (1136 of 1955) is reported. A formula is derived for the attenuation of the scattered power relative to the free-space power; the diffraction field and the effect of super-refraction are neglected, but the effect of the ground-reflected wave is taken into account. This function is plotted against antenna height in wavelengths, and against the distance between the antennas, with $f(H)$, a function of the dielectric inhomogeneity of the troposphere, as parameter.

621.396.11.029.62:523.5 923
Meteoric Echoes Observed Simultaneously by Back-Scatter and Forward Scatter—D. W. R. McKinley and A. G. McNamara. (*Canad. J. Phys.*, vol. 34, pp. 625–637; July, 1956.) "Simultaneous observations of back-scatter and forward-scatter meteoric echoes have been made by means of a high-power 33 mc pulse transmitter at Ottawa, with identical receiving systems at Ottawa and at Scarboro, 337.8 km distant. Two-way transmissions, employing a low-power transmitter at Scarboro, were also used to measure absolute time delays. The approximate position of each meteor was plotted from the observed time delays, which enabled corrections to be applied to the echo durations for variations in antenna patterns and other factors, and which also determined the forward-scatter angle, 2ϕ, for each meteor. In the majority of cases an enhancement was observed in the forward-scatter duration relative to the back-scatter duration. The data were divided into a short-duration or underdense group and a long-duration or overdense group. Assuming a theoretical forward-scatter enhancement proportional to $\sec^m\phi$, it was found that the exponent, m, was 1.73 for the underdense group and 1.13 for the overdense group."

621.396.812.3.029.6 924
Fading of Ultra-short Waves and its Relation to the Meteorological Conditions—K. Hirao. (*J. Radio Res. Labs, Japan*, vol. 3, pp. 189–255; July, 1956.) Observations taken in Japan on a frequency of 65.82 mc are discussed in relation to interference fading due to large-scale irregularities of refractive index, and scintillation fading caused by atmospheric turbulence. A specially designed semi-manual record reader and a relay computer used for autocorrelation analysis of data are described and the results are related to meteorological conditions in the lower atmosphere, quantitative conclusions being drawn.

RECEPTION

621.376.23:621.3.018.7 925
Detection of Pulses with Complex Form—E. L. Gerenrot. (*Radiotekhnika i Elektronika*, vol. 1, pp. 438–442; April, 1956.) Transient processes in an ideal pulse detector are considered, assuming that the source impedance cannot be neglected. A general method of calculating the voltage appearing across the load is given for signal pulses of arbitrary shape.

621.376.233:621.3.018.7 926
Transient Processes in the Detection of Weak Signals—L. S. Gutkin. (*Radiotekhnika i Elektronika*, vol. 1, pp. 433–437; April, 1956.) Analysis of the transient processes in a crystal-diode detector is presented. A formula is derived for the distortion of a pulse of arbitrary shape.

621.396.3:621.376.4 927
Elimination of "Reverse Operation" in an Amplitude-Phase Detector due to Fluctuation

Interference—Yu. S. Lezin. (*Radiotekhnika i Elektronika*, vol. 1, pp. 329–334; March, 1956.) The probability is calculated of the occurrence of a change of polarity of the output voltage in an amplitude-phase telegraphy detector due to fluctuation-type interference. Results indicate that if the signal/noise ratio at the input of the detector is greater than unity, then the probability can be made negligibly small by narrowing the pass band of the tuned detector circuit relative to that of the IF amplifier. A circuit diagram of the amplitude-phase detector is given and its operation is briefly described.

621.396.62:621.396.41 928
The Phase-Shift Method of Single-Sideband Signal Reception—D. E. Norgaard. (PROC. IRE, vol. 44, pp. 1735–1743; December, 1956.) Analysis complementary to that for ssb signal generation (955 below) is presented. Zero-frequency signals derived by demodulators from a transmitted pilot carrier may be used for control of gain and frequency in the receiver.

621.396.621:621.396.41 929
Factors Influencing Single-Sideband-Receiver Design—L. W. Couillard. (PROC. IRE, vol. 44, pp. 1750–1753; December, 1956.) The factors considered include frequency stability, cross modulation, gain distribution, and diversity combining.

621.396.621.54:621.376.3:621.314.7 930
Transistorized Receiver for Mobile F.M.—A. M. Booth[e]. (*Electronics*, vol. 29, pp. 158–161; November, 1956.) A receiver for mass production uses printed circuits and 19 available-type transistors. It operates on a 12.5-mc signal produced as an IF by a tube tuner covering the band 20–70 mc. Temperature variations from −67° to +149°F and simultaneous supply-voltage variations from 22 to 30 v are tolerable. A limiter stage and a Foster-Seeley discriminator are included.

621.396.822:621.376.23:519.2 931
On the Distribution of the Product of Diode Detector Waveforms—E. L. R. Webb. (*Canad. J. Phys.*, vol. 34, pp. 679–691; July, 1956.) "The probability distribution of the product of two waveforms such as come from the diode second detectors of radio receivers is examined over the whole range of signal-to-noise ratios. Computed curves of probability density are given for small and moderate values of signal-to-noise ratio and the limiting form for large signal to noise indicated. The pure noise case is the only one immediately available in terms of tabulated functions. Compared to the Rayleigh distribution it rises much faster, reaches its maximum sooner and lower, and decays much more slowly. The very large signal-to-noise ratio case approaches an impulse function. Estimates of mean and variance are given."

STATIONS AND COMMUNICATION SYSTEMS

621.3.018.7 932
Signals of Finite Duration, containing Maximum Energy for a Given Bandwidth—M. S. Gurevich. (*Radiotekhnika i Elektronika*, vol. 1, pp. 313–319; March, 1956.) A mathematical paper on a problem similar to that discussed by Chalk (1518 of 1950).

621.39.001.11 933
1956 Symposium on Information Theory—(IRE TRANS. vol. IT-2, September, 1956.) The text is given of papers presented at a symposium held at the Massachusetts Institute of Technology in September 1956, including the following:—

The Zero Error Capacity of a Noisy Channel—C. E. Shannon (pp. 8–19).

A Linear-Circuit Viewpoint on Error-Correcting Codes—D. A. Huffman (pp. 20–28).

Theory of Information Feedback Systems—S. S. L. Chang (pp. 29–40).

A Linear Coding for Transmitting a Set of Correlated Signals—H. P. Kramer and M. V. Mathews (pp. 41–46).

On an Application of Semi-group Methods to some Problems in Coding—M. P. Schützenberger (pp. 47–60).

An Extension of the Minimum Mean-Square Prediction Error Theory for Sampled Input Signals—M. Blum (pp. 176–184).

A New Interpretation of Information Rate—J. L. Kelly, Jr. (pp. 185–189).

An Outline of a Purely Phenomenological Theory of Statistical Thermodynamics: Part 1—Canonical Ensembles—B. Mandelbrot (pp. 190–203).

Abstracts of these papers appear in *Proc. Inst. Radio Engrs.*, vol. 44, pp. 1643–1644. November, 1956.

621.39.001.11 934
Theory of Ideal Coding of a Binary Transmission—V. I. Siforov. (*Radiotekhnika i Elektronika*, vol. 1, pp. 407–417; April, 1956.)

621.39.001.11 935
The Formation of Code Words—R. Schauffler. (*Arch. Elekt. Übertragung*, vol. 10, pp. 303–314; July, 1956.) Formulas and charts are developed to facilitate the detection and correction of common forms of mutilation in code transmission. The formation of more general code systems is discussed.

621.396.41 936
Synchronous Communications—J. P. Costas. (PROC. IRE, vol. 44, pp. 1713–1718; December, 1956.) The performance of a system using synchronous detection with dsb a.m. is compared with that of a ssb system. The dsb system is less susceptible to jamming and is equal to the ssb system as regards the efficient use of power; the dsb system also shows an advantage by virtue of the greater simplicity of the equipment, especially at the transmitter. The number of usable channels is not necessarily doubled, and in some practical situations may not be increased at all by the use of ssb. In a synchronous receiver designed for the U.S.A.F., phase information for controlling the local oscillator is derived from the sidebands alone, no pilot carrier or synchronizing tone being required.

621.396.41 937
Single-Sideband Technique—(PROC. IRE, vol. 44; December, 1956.) The main part of this issue is devoted to a group of papers constituting a survey of the technique of communication by ssb. Abstracts of some of the papers are given individually; titles of the others are as follows:—

An Introduction to Single-Sideband Communications—J. F. Honey and D. K. Weaver, Jr. (pp. 1667–1675).

Early History of Single-Sideband Transmission—A. A. Oswald (pp. 1676–1679).

Synthesizer-Stabilized Single-Sideband Systems—B. Fisk and C. L. Spencer (pp. 1680–1685).

A Suggestion for Spectrum Conservation—R. T. Cox and E. W. Pappenfus (pp. 1685–1688).

Power and Economics of Single Sideband—E. W. Pappenfus (pp. 1689–1691).

Application of Single-Sideband Technique to Frequency-Shift Telegraph—C. Buff (pp. 1692–1697).

Frequency Control Techniques for Single Sideband—R. L. Craiglow and E. L. Martin (pp. 1697–1702).

Comparison of Linear Single-Sideband Transmitters with Envelope-Elimination-and-Restoration Single-Sideband Transmitters—L. R. Kahn (pp. 1706–1712).

Automatic Tuning Techniques for Single-Sideband Equipment—V. R. DeLong (pp. 1766–1774).

Single-Sideband Operation for International Telegraph—E. D. Becken (pp. 1782–1788).

S.S.B. Receiving and Transmitting Equipment for Point-to-Point Service on H.F. Radio Circuits—H. E. Goldstine, G. E. Hansell, and R. E. Schock (pp. 1789–1794).

Conversion of Airborne H. F. Receiver-Transmitter from Double Sideband to Single Sideband—H. A. Robinson (pp. 1794–1799).

Problems of Transmission to Single-Sideband Operation—N. H. Young, Jr. (pp. 1800–1803).

The Problems of Transition to Single-Sideband Techniques in Aeronautical Communications—J. F. Honey (pp. 1803–1809).

Single-Sideband Techniques applied to Coordinated Mobile Communication Systems—A. Brown (pp. 1824–1828).

Single Sideband in the Amateur Service—G. Grammer (pp. 1829–1833).

Comparison of S.S.B. and F.M. for V.H.F. Mobile Service—H. Magnuski and W. Firestone (pp. 1834–1839).

Design of a High-Power Single-Sideband V.H.F. Communication System—J. W. Smith (pp. 1848–1853).

621.396.41 938
S.S.B. Performance as a Function of Carrier Strength—W. I. Firestone. (PROC. IRE, vol. 44, pp. 1839–1848; December, 1956.) "This paper shows the important part that the carrier plays in over-all system performance and in particular compares the various systems using full carrier, reduced carrier, suppressed carrier and controlled carrier. It is concluded that as the carrier is reduced, the factors of modulation splatter, transmitter efficiency, available peak sideband power, desensitization, and intermodulation all tend to improve. It is also pointed out that due to system stability requirements, complete suppression at the higher radio-frequencies is not feasible. Because there are many types of s.s.b. receiving systems, each requiring a different amount of carrier for synchronizing purposes, it is necessary to consider all values of transmitted carrier to compare the resulting systems and to gain a better understanding of the system characteristics considered. The characteristics of the controlled carrier system are discussed for completeness."

621.396.41:621.396.11 939
Single-Sideband Techniques in U.H.F. Long-Range Communications—W. E. Morrow, Jr., C. L. Mack, Jr., B. E. Nichols, and J. Leonhard. (PROC. IRE, vol. 44, pp. 1854–1873; December, 1956.) A comparison of f.m. and s.s.b. AM techniques for communication systems based on beyond-horizon propagation indicates that the s.s.b. technique affords advantages in respect of spectrum conservation, performance in the presence of multipath propagation, and power requirements for a given signal/noise ratio. The design of equipment for the frequency band 300–400 mc is described; methods of achieving efficient operation with high-power klystrons are indicated.

621.396.41:621.396.931 940
The Application of S.S.B. to High-Frequency Military Tactical Vehicular Radio Sets—R. A. Kulinyi, R. H. Levine, and H. F. Meyer. (PROC. IRE, vol. 44, pp. 1810–1823; December, 1956.) Advantages obtainable by the use of s.s.b. rather than d.s.b. communication systems for military purposes include improved signal/noise ratio, leading to in-

creased range and intelligibility, reduced interference, improved spectrum utilization, quasi-duplex operation, reduced heat generation, greater reliability, and amelioration of maintenance problems. Compatible s.s.b. and d.s.b. systems are discussed.

621.396.41.029.6:621.3.018.78 941
R.F. Bandwidth of Frequency-Division Multiplex Systems using Frequency Modulation—R. Hamer: R. G. Medhurst. (PROC. IRE, vol. 44, p. 1878; December, 1956.) Comments on a paper by Medhurst (1547 of 1956) and author's reply.

621.396.41.029.63:621.318.57 942
Subcarrier Switch for Microwave Party Line—B. Harris. (*Electronics*, vol. 29, pp. 175–177; November, 1956.) Circuit arrangements are described for ensuring that in any channel of a multichannel radio-communication system only one station shall have its carrier operating at any time, the rectified output from the receiver providing a bias which cuts off the carrier at the local transmitter unless the outgoing a.f. signal is greater than the incoming one.

621.396.662:621.396.61/.62 943
Automatic Tuning Mechanisms using Instantuners Type SZT 201 and/or 202—W. L. Vervest and L. van Gorkom. (*Philips Telecommun. Rev.*, vol. 17, pp. 2–16; August, 1956.) Detailed description of equipment incorporating improvements over that described previously [2941 of 1949 (Vervest)].

621.396.712.029.62:621.376.3 944
High-Quality Sound Broadcasting—G. H. Russell. (*Wireless World*, vol. 63, pp. 31–32; January, 1957.) Discussion of a report published by the European Broadcasting Union on *The Present Position and Prospectives of V.H.F. Sound Broadcasting in Europe.* Both technical and economic aspects of the development of v.h.f. f.m. transmitting networks are examined, and the stage reached in various countries is indicated.

SUBSIDIARY APPARATUS

621.311.6:621.316.722:621.314.7 945
Regulated Transistor Power-Supply Design—J. W. Keller, Jr. (*Electronics*, vol. 29, pp. 168–171; November, 1956.) Simple analysis is presented for series and shunt-regulated circuits for low-voltage power supplies.

621.352 946
Recent Patents on Electric Cells—L. Jumau. (*Rev. Gén. Élect.*, vol. 65, pp. 401–418; July, 1956.) Continuation of previous review (264 of 1954).

TELEVISION AND PHOTOTELEGRAPHY

621.397.26:621.397.6 947
A British Microwave Television Link in Canada—A. D. Hodgson and G. M. B. Wills. (*G.E.C.J.*, vol. 23, pp. 123–129; July, 1956.) A description is given of the London-Windsor radio link, in the province of Ontario. The route length is 120 miles; there are four repeater stations. Operation is in the frequency band 1.7–2.3 kmc; frequency modulation is used, with a transmitter deviation of 6 mc peak-to-peak and a receiver bandwidth of 16 mc. Disk-seal tubes are used in the transmitter uhf circuits.

621.397.5/.6:535.623 948
Colour Television—G. N. Patchett. (*J. Brit. Instn Radio Engr.*, vol. 16, pp. 591–620; November, 1956.) "Theory of color mixing and of colorimetry is discussed briefly. Various systems for color television are outlined and studio and receiver equipment described. The N.T.S.C. system and its modification to British standards are discussed." Over 150 references.

621.397.5:535.623:778.5 949
Recent Improvements in Black-and-White Film Recording for Colour-Television Use—W. L. Hughes. (*J. Soc. Mot. Pict. Telev. Eng.*, vol. 65, pp. 359–364; Discussion, p. 364; July, 1956.) Account of the development of a system suitable both for producing films by mechanical camera for flying-spot scanning, and for making kinescope recordings. Similar material is presented in 1955 IRE CONVENTION RECORD, Part 7, pp. 69–80.

621.397.5:535.623:778.5 950
Colour Kinescope Recording on Embossed Film—C. H. Evans and R. B. Smith. (*J. Soc. Mot. Pict. Telev. Eng.*, vol. 65, pp. 365–371; Discussion, pp. 371–372; July, 1956.)

621.397.5:621.39.001.11 951
Television Systems with Statistical Encoding—B. B. Gurfinkel'. (*Radiotekhnika i Elektronika*, vol. 1, pp. 478–496; April, 1956.) Theory of encoding systems using a nonlinear transformation of the signal-function time-scale is presented, and several practical systems reported in the literature are briefly discussed

621.397.6.001.4:621.317.74 952
Measuring Colour-Television Luminance vs Chroma Delay—Ettlinger. (See 880.)

621.397.61:771.35 953
Optics Before the Camera—C.; Burns. (*J. Telev. Soc.*, vol. 8, pp. 117–120, 122; July–September, 1956.) Practical details are given regarding the nature and adjustment of optical systems used with television cameras.

621.397.621.2:621.385.832 954
Frequency Characteristics of Kinescopes—L. M. Selyakov. (*Radiotekhnika i Elektronika*, vol. 1, pp. 525–534; April, 1956.) The dependence of M, the ratio of the modulation coefficient of the visual brightness of the sinusoidal signal on the screen to the modulation coefficient of the signal at the modulation tube, is calculated as a function of the video frequency, taking into account the effect of the halo. Calculated and experimentally determined characteristics of typical Russian picture tubes are tabulated and presented graphically.

TRANSMISSION

621.396.61:621.396.41 955
The Phase-Shift Method of Single-Sideband Signal Generation—D. E. Norgaard. (PROC. IRE, vol. 44, pp. 1718–1735; December, 1956.) A general expression is derived for sideband suppression obtained by the phase-shift method. The suppression ratio is expressed in terms of four system parameters, three of which depend on the wide-band phase-shift networks used. A simple dual-channel s.s.b. generator is described. Use of the phase-shift method in conjunction with band-pass filters is discussed. The effects of intermodulation distortion and the performance stability are examined.

621.396.61:621.396.41:621.375.221 956
Distortion-Reducing Means for Single-Sideband Transmitters—W. B. Bruene. (PROC. IRE, vol. 44, pp. 1760–1765; December, 1956.) Methods of reducing intermodulation distortion products from rf power amplifiers used in multichannel s.s.b. transmitters are discussed. Direct rf feedback is adjudged preferable to the method of envelope distortion cancelling modulation. A circuit combining the two techniques is described.

TUBES AND THERMIONICS

537.533 957
Single-Component Stationary Electron Flow under Space-Charge Conditions—B Meltzer. (*J. Electronics*, vol. 2, pp. 118–127; September, 1956.) It is shown analytically that for all electron beams issuing from a cathode with zero or negligible velocity the flow can be treated as one with a single velocity component. The differential equation describing the flow is relatively simple, particularly when the coordinates are chosen to be orthogonal. Explicit expressions are obtained for the current, charge densities and potentials in a flow from a space-charge-limited cathode constrained to a circular path.

621.314.63:621.318.57 958
Fast Switching with Junction Diodes—J. E. Scobey, W. A. White, and B. Salzberg. (PROC. IRE, vol. 44, pp. 1880–1881; December, 1956.) By taking as operating point the reverse breakdown voltage rather than zero voltage, switching speed and upper frequency limit can be increased. Special selection of diodes is necessary for this class of operation.

621.314.632 959
On the Anomalous Rectification of Cuprous Sulphide Detectors—M. Anastassiades and D. Ilias. (*Proc. Phys. Soc.*, vol. 69, pp. 958–960; September 1, 1956.) Reversal of the direction of rectification is observed in CuS rectifiers as the applied voltage passes through the value 0.3 v rms. The effect is attributed to rectification at the contact with the holder, acting in opposition to that of the unit proper.

621.314.7 960
Developmental Study on Point-Contact Transistors—M. Aida. (*Rep. Elect. Commun. Lab., Japan*, vol. 4, pp. 18–28; May, 1956.) Aspects of the assembly relevant to reliability of subsequent operation are discussed, and a cartridge designed to ensure correct contact pressure is described. Measured temperature variations and noise characteristics are presented, as well as results of life and humidity tests.

621.314.7 961
Measurement of the Parameters Determining the High-Frequency Performance of Transistors. Elements of the "Natural" Equivalent Circuit—J. Riethmüller. (*Ann. Radioélect.*, vol. 11. pp. 239–248; July, 1956.) Test methods are described. A comparison between measured values and those derived from the "natural" equivalent circuit [607 of 1956 (Zawels)] confirms the validity of this network.

621.314.7:621.396.822 962
Temperature Dependence of Flicker Noise of *p-n-p* Junction Transistors—K. Amakasu and M. Asano. (*J. Appl. Phys.*, vol. 27, p. 1249; October, 1956.) Measurements over the temperature range $-150°$ to $+43°$C are presented graphically and discussed briefly.

621.314.7.001.4 963
A Transistor Tester—(See 861.)

621.383.27 964
An Improved Photomultiplier Construction—A. E. Jennings and C. E. F. Misso. (*J. Sci. Instru.*, vol. 33, pp. 323–324; August, 1956.) A slatted design described by Sommer and Turk (2085 of 1950) is discussed and a modified construction giving improved focusing is proposed.

621.383.4 965
Alternating-Current Measurements on Cadmium Sulphide Photocells—E. Klier. (*Ann. Phys., Lpz.*, vol. 18, pp. 163–170; August 15, 1956.) Cells with ohmic and with non-ohmic contacts were investigated. The results indicate that cells treated in a glow discharge exhibit the same behavior with alternating current as with direct current.

621.385.029.6 966
The Focusing of Electron Beams by an Alternating Longitudinal Magnetic Field—O. Cahen. (*Ann. Télécommun.*, vol. 11, pp. 142–150; July/August, 1956.) Analysis is developed based on the equations of motion of the electrons. Formulas are derived relating ripple length and beam diameter. A focusing arrangement is discussed comprising two interleaved sets of magnetizable members associated with coils and surrounded by an iron tube to complete the magnetic circuit. The arrangement was tested in a traveling-wave tube.

621.385.029.6 967
Study of the Oscillation Modes of the M-Type Carcinotron: Part 2—M. de Bennetot. (*Ann. Radioélect.*, vol. 11, pp. 230–238; July, 1956.) Extension of the analysis given in part 1 (3255 of 1956) to beams of arbitrary thickness. Expressions are derived for the energy exchange between beam and delay line and the boundary conditions in the interaction space.

621.385.029.6–712 968
Air Cooling a Finned Magnetron—M. Mark. (*Tele-Tech and Electronic Ind.*, vol. 15, pp. 100–101, 178; June, 1956.) A light-weight forced-air cooling system is described, suitable for use in airborne equipment.

621.385.032.73.001.4 969
The Control of Thermionic Valve Envelope Quality by Thermal-Shock Testing—G. D. Redston. (*Electronic Eng.*, vol. 28, pp. 470–475; November, 1956.) Failures of all-glass tubes in thermal shock tests are discussed; different defects are brought out by different tests. Results of the "downward" thermal-shock test are more nearly correlated with service life than those of the "upward" thermal-shock test. Tempering the tube base improves resistance to "downward" thermal shock but produces no significant change in the number of failures on life test.

621.385.3/.5:621.396.822 970
Uncorrelated Grid Noise—D. A. Bell. (*Electronic Radio Eng.*, vol. 34, pp. 36–37; January, 1957.) An explanation is advanced of the absence of correlation observed *e.g.*, by Houlding and Glennie (1261 of 1954) between grid and anode noise.

621.385.8+621.317.755].029.63/.64 971
Oscillograph for Investigating U.H.F. Oscillations and Some Results of its Application in the Study of Pulsed Magnetrons—Chernushenko. (See 885.)

621.387:621.316.722.1:621.396.822 972
Gas-Filled Voltage Stabilizers—F. A. Benson. (*Electronic Radio Eng.*, vol. 34, pp. 16–20; January, 1957.) Tubes manufactured specially for previous investigations [640 and 1263 of 1956 (Benson and Bental)] were used for measurements of the effects on the noise characteristics of varying the cathode material, the gas filling and the gas pressure. Results are presented and discussed.

621.387:621.318.57 973
The Design of Cold-Cathode-Valve Circuits—Flood and Warman. (See 695.)

MISCELLANEOUS

621.3(47) 974
[Russian] Books on Radio Engineering and Electronics in 1956—P. O. Chechik. (*Radiotekhnika i Elektronika*, vol. 1, pp. 537–539; April, 1956.) The list includes over 70 titles of books by Russian authors.